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"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."

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THE 1957 CONVENTION

Most members will be familiar with the pattern of the Conventions organized by the Institution in the post-war years. Arranged in order to review current developments and future trends in the specialized divisions of radio engineering, the Conventions have provided an excellent opportunity for national and, indeed, international exchange of ideas.

Experience has shown that such meetings are desirable about once every three years; certainly the success of previous gatherings may be judged from requests constantly received, particularly from overseas, for as much notice as possible to be given of any future Conventions. For this latter reason, and to assist members who will wish to contribute to the proceedings, arrangements are now being made to hold the next Convention from 26th June to 1st July, 1957.

The venue selected is King's College in the University of Cambridge—the centre of the Television Session of the 1951 Radio Convention.* Next year all Sessions of the Convention will be centred on Cambridge and the Council of the Institution gratefully acknowledges the advice and help of the authorities of King's College and the University in securing the necessary facilities.

In deciding that the general theme of the Convention should be "Electronics in Automation", the Council has been influenced by the success and general appeal of the 1954 Convention at Oxford†, which was devoted to "Industrial Electronics". The majority of papers presented at that Convention, and the subsequent discussions, centred upon what is now generally regarded as "automation".

Many members will agree with the Earl of Halsbury‡ " . . . that those who coined these new terms did not include a philologist," but the word "automation" is now widely used—if not properly understood—in describing the ever increasing process of mechanization. Indeed, the word now has official blessing by being the subject of a report by the Department of Scientific and Industrial Research.§

Understandably, this report deals most usefully with the problems of management in introducing automation and perhaps on this account was well received in the House of Commons on 17th May last. Reviewing well known techniques, the report has, however, emphasized that electronics has made a major contribution to the development of automation in the fields of automatic control and, more particularly, in the present and some possible uses of electronic computers. The report also states that engineers with a variety of technical backgrounds will be needed and that special courses in electronics will be required for technicians.

The purpose of the 1957 Convention will be to hear and discuss the demands made upon the electronics engineer to design equipment capable of processing information and initiating action. There is, therefore, every promise of an outstandingly interesting Convention which will usefully implement the principal object of all such Institutions as our own—"the advancement of science for the benefit of man".

* "The 1951 Radio Convention", *J.Brit.I.R.E.*, **11**, pp. 205, 274, June and July 1951.

† 1954 Convention Programme, *J.Brit.I.R.E.*, **14**, p. 238, June 1954.

‡ Rt. Hon. the Earl of Halsbury, "Automation". *J. Royal Soc. Arts*, **104**, p. 535, 8th June, 1956.

§ Department of Scientific and Industrial Research, "Automation: a Report on the Technical Trends and their Impact on Management and Labour." (London, H.M.S.O., 1956.)

NOTICES

Birthday Honours List

The Council of the Institution congratulates Mr. L. H. Bedford, M.A., B.Sc. (Past President), on his promotion to Commander of the Most Excellent Order of the British Empire, announced in Her Majesty's Birthday Honours List. Mr. Bedford, who is Chief Engineer of the Guided Weapons Division of the English Electric Company, was appointed an O.B.E. in 1943. A note on his career appears in the *Journal* for January 1956.

New Edition of the List of Members

A new edition of the Brit.I.R.E. List of Members is in course of preparation. As in previous years, the List will include the names of all corporate members, Associates and Graduates (but not Students), and will contain the Articles of Association, membership of standing and local Committees, and other reference material. Copies of the List will be sent to all members whose names are included in it, and the publication date will be announced in due course.

In order that the information contained in the List shall be as up-to-date as possible, members are asked to notify the Institution if their latest addresses, and personal details such as rank, degrees and other qualifications, honours, etc., are not correctly shown on their present address plates, i.e. as used for the despatch of the *Journal*.

Recent Appointments

Professor H. W. Melville, F.R.S., Mason Professor of Chemistry in the University of Birmingham for the past eight years, has succeeded Sir Ben Lockspeiser, K.C.B., F.R.S., as Secretary of the Department of Scientific and Industrial Research. Sir Ben, who has retired on reaching the age of sixty-five, was appointed Secretary of the D.S.I.R. in 1949; two years ago he was elected first president of the European Organization for Nuclear Research (CERN).

The position of Director of the British Scientific Instrument Research Association is to be filled by Dr. J. Thomson, M.A., D.Sc., at present Deputy Director of Physical Research at the Admiralty. Mr. A. J. Philpot, C.B.E., M.A., B.Sc., is retiring shortly after having been with the Association for thirty-six years.

New Committee of North Eastern Section

The new Committee of the Institution's North Eastern Section for the session 1956-7 has recently been elected. The Chairman is H. Brennan, B.Sc. (*Member*), and L. G. Brough (*Associate Member*) is Treasurer; H. Armstrong (*Associate Member*) is Membership Secretary, and J. Bilbrough (*Associate*) Programme Secretary. Other members of the Committee are:—J. Beer (*Associate Member*), W. A. Davis (*Associate Member*), O. B. Kellett (*Member*) and J. W. Osselton (*Graduate*).

Television Developments

During June, the B.B.C.'s new television transmitting mast at Rowridge, Isle of Wight, came into operation. Rowridge originally opened in November 1954 with a temporary 200-ft. aerial tower, with transmitting frequencies of 56·75 Mc/s for vision and 53·25 Mc/s for sound (channel 3); the permanent aerial system on the new 500-ft. mast, using the same frequencies, will increase the effective radiated power of the station by more than three times—to 32 kW at its greatest directivity. This should extend the coverage to link up with the service areas of the B.B.C.'s Wenvoe and Crystal Palace stations. The temporary transmitter of Truleigh Hill, near Brighton, will continue in operation for the time being.

Recent television broadcasts from liners in the English Channel and from a submarine of the Royal Navy, have made use of the permanent radio link between Rowridge and London.

The B.B.C. also announces the acquisition of two new sites for television stations. One is to be at Sandale, Cumberland, 1,200 ft. above sea level and some 14 miles south-west of Carlisle, and is expected to be completed during 1957. It was stated in the House of Commons that the B.B.C. and the I.T.A. had agreed to co-operate to avoid the duplication of masts at this site; the development programme of the I.T.A. indicates, however, that it will be some years before their station is set up. The other new B.B.C. station, which should also be completed during 1957, is to be at Sheriffs Mountain, near Londonderry, and will extend the service to more than 130,000 people in the western part of Northern Ireland. Both stations will be suitable for v.h.f.-f.m. transmission of sound radio when it becomes possible to extend the system to these areas.

EXTENDING THE LIMITS OF RESISTANCE MEASUREMENT USING ELECTRONIC TECHNIQUES*

by

Gerald Hitchcox (Member)†

Read before a meeting of the Institution in London on 28th September, 1955

In the Chair: Mr. A. G. Wray, M.A. (*Associate Member*)

SUMMARY

In general, this paper is confined to a discussion of devices and techniques which are suitable for inclusion in commercial instruments, rather than of those which, by their complication, delicacy, or specialization, are restricted to the laboratory. It opens with a description of the basic ohmmeter in its two dual forms, refers to three low resistance systems, two using direct and one alternating test currents, and then goes on to describe what is thought to be a novel instrument which can measure very low resistance using pulsed test currents to reduce thermal dissipation in in the test sample.

At the very high resistance end of the range where extension has been much greater, the three principal systems are discussed, together with the application of modern devices such as differential constant current generators, electronic stabilizers, dynamic capacitor modulators, and so on; and the paper ends with a detailed description of a widely-used general purpose commercial megohmmeter.

1. Introduction

Up to the outbreak of the last war the measurement of electrical resistance was confined to a range extending from perhaps one tenth of one milliohm up to one thousand megohms. Even so, a coverage of ten billion to one is impressive; it has only one parallel in electrical engineering—the related quantity current—and very few in either mechanical engineering or physics. In spite of the magnitude of this range, its measurement could be satisfactorily effected by conventional devices, conventional here meaning non-electronic. No distinction need be drawn in this sense between ohmmeters in their dual forms, ratiometers, and bridges, since in general the limitations to their ranges are decided by the same factors, principally the sensitivity of the indicating device.

Early in the war, the measurement of resistance values much higher than 1,000 megohms became necessary. One such case lay in the development and testing of the capacitor discharge projectile fuse; here one charged capacitor is discharged through a high resistance network into another which ultimately triggers a cold cathode tube; the operational merit is the ability to vary the pre-set delay time right up to the moment of firing. Capacitors were used with a parallel leakage resistance so high that their natural time constant approached one year. Since the war, more drastic requirements have arisen. The generation and measurement of the minute currents involved in nuclear physics demand the accurate measurement of very high resistors and the detection of even higher leakage resistances. Determination of moisture content in synthetic compounds such as nylon requires indication of resistance up to 10^{16} ohms. Such figures lie so far outside ordinary experience that they are difficult to grasp; 10^{16} ohms is the resistance of one billion miles of 26 s.w.g. nichrome wire.

At the other end of the range extension has been less dramatic; the measurement of very low resistance has increased in importance rather than descended in magnitude, helped by the tendency to larger electrical machines and transformers, the growth of railway electrification, and the increasing use of high conductivity cells in chemical process control.

The character of this paper is essentially quantitative—its object is the discussion, not of

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† Mr. Hitchcox is a Consulting Engineer.

U.D.C. No. 621.317.732/6:621.37/8.

a new field, but rather of extensions to an old one using new techniques such as electronic constant current devices, high performance stabilizers, modulator d.c. amplifiers, time sampling, and differentiators both to generate minute currents and to analyse complex impedances. It may be useful first to consider those factors which limit the range of non-electronic instruments.

1.1. *Limitations to the Range of Non-Electronic Instruments*

Figure 1 shows the direct-reading ohmmeter in its two dual forms. In the series type (*a*) a constant voltage E is applied across the unknown resistance R and the resultant current I indicated by the series ammeter. This meter could be linearly calibrated in terms of conductance $G=I/E$; conventionally it reads resistance $R=E/I$, reciprocally and therefore less conveniently. Fig. 1b shows the less well-known dual form where a constant current I is applied in series with the unknown; here the meter is a voltmeter and naturally reads resistance linearly since $R=IV$ where V is the potential drop across the unknown. It is usual though not essential to insert the ballast resistors shown in broken lines, in series and in parallel with the unknown respectively. Their effect is to extend the indicated resistance range from infinity to zero, or vice versa in type (*b*), with a double object: partly to provide two "natural" calibration points, and partly to prevent damage to the meter since no reading can possibly lie outside such a range. Such protective resistors destroy the constancy of the test voltage and test current supplies respectively, and with it the possibility of strictly linear calibration; their justification is additional safety and convenience.

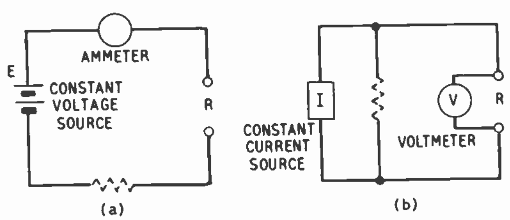

Fig. 1.—The two dual forms of ohmmeter, (a) constant voltage, (b) constant current.

With the series type (*a*) the upper limit to the measurable range is decided by the maximum test voltage considered reasonable and by the minimum current readable on a practical meter. Taking these to be 1,000 volts and 1 microampere respectively postulates an upper limit of 1,000 megohms. The lower limit is governed both by the current capacity and the internal resistance of the test voltage supply; using a dry cell a common value is 0·1 ohm.

In the parallel type (*b*) the lower resistance limit—taking this first to preserve duality in the comparison—is decided by the maximum level of test current which can be conveniently generated and by the minimum potential drop readable on a practical meter. Taking these as 10 amperes and 1 millivolt respectively gives a lower limit of 0·1 milliohm. The upper end is determined by the properties of the constant current supply; current must be high enough to permit some diversion by the meter, and this increases the difficulty of providing a constant current supply whose output is unaffected by the voltage drop across a high test resistance. A common upper limit is 10,000 ohms.

It will be seen that these limitations to the two dual forms of ohmmeter, though differing in magnitude, are duals of each other. None of them is rigid, of course. Any one can be exceeded by sacrificing accuracy, or reliability, or convenience, or by expedients normally considered uneconomical, or by exceptionally ingenious design. Sometimes limitations may be imposed external to the construction of the instrument; for example, in type (*a*) a test voltage of 1,000 volts may be unacceptable because of the danger of breakdown in the circuit being measured.

2. The Measurement of Very Low Resistance

2.1. *Use of the Four Terminal Network*

Nearly all direct reading instruments are either ratiometers or ohmmeters of the type skeletonized in Fig. 1b in which a constant current is injected into the unknown and resistance measured linearly in terms of potential drop. Four test terminals and leads are used, two carrying the test current and two connecting the test resistance to the voltage measuring device. Separation in function largely obviates the effect of lead resistance. Resistance in the current leads cannot affect the test current if a constant current source is used; alternatively, this effect is nullified if the instrument is a ratiometer. Resistance in the voltmeter leads is unimportant provided that it is low relative to that of the voltmeter itself, no problem where an electronic amplifier is used.

2.2. *Instruments Using Direct Test Currents*

An ingenious non-electronic instrument is the Evershed and Vignoles "Ducter," a true ratiometer. The meter is a "crossed-coil" device, one section carrying a current proportional to that injected into the unknown and the other a current proportional to the potential dropped across it. The meter, a floating zero type without mechanical restraint, automatically indicates the quotient of resultant potential drop divided by causative current, i.e., resistance.

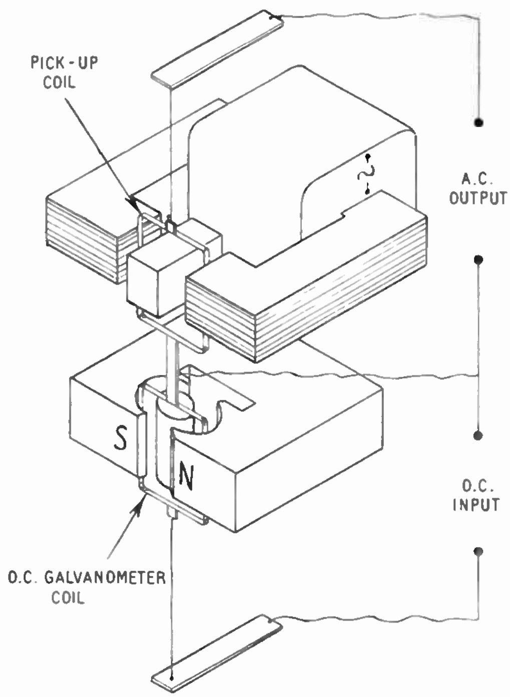

Fig. 2.—Construction of the Pye "Microvoltmeter."

With a ratiometer system, since accuracy is unaffected within broad limits by variations in the absolute magnitude of the test current, much higher values can be used than could be conveniently generated in a constant current source, this resulting in increased sensitivity and a lower bottom range. That of the "Ducter" using nominally 100 amperes test current is 0 to 100 microhms; an otherwise very similar Russian instrument (Technopromimport M-246) uses only 20 amperes, attained by some sacrifice in mechanical robustness.

Equally low resistance values can be measured with much smaller test currents using the Pye "Microvoltmeter" as the potential detection device. On the second most sensitive range, 0 to 100 μV, resistance from 0 to 100 microhms can be covered with a test current of one ampere. The attraction of so low a current is the ease with which it can be generated and the negligible risk of thermal or magnetic damage to the unknown. The "Microvoltmeter" is at present available only as a voltmeter; apparatus for current generation, calibration, and so on, must be provided by the user.

Its principle is suggested in Fig. 2. A current proportional to the input voltage is fed to the d.c. galvanometer coil, mechanically coupled to a pick-up coil which can rotate in a powerful a.c. field. Any d.c. input turns this coil away from its neutral position, so that a voltage is induced in it whose magnitude and phase are respectively proportional to the magnitude and direction of the d.c. input. The a.c. signal is then fed to a high-gain two-stage electronic amplifier (not shown) the output from which is rectified in a phase-sensitive network and returned to the d.c. galvanometer in such a manner that it opposes the input signal. The whole arrangement acts as a high gain, high stability servo loop, the output indication being picked off from a high level point and fed to a centre-zero meter. In its commercial form the most sensitive range corresponds to an input of 0 to $\pm$ 10 μV with a long term zero stability of better than $\pm 0{\cdot}5$ μV.

2.3. *Instruments Using Alternating Test Currents*

In a different class of instrument, alternating test currents are used at mains frequency, the potential drop across the unknown resistance being taken to an electronic alternating voltage amplifier, then rectified and fed to a panel meter directly calibrated in terms of resistance. The advantages of a.c. operation are the ease with which large test currents can be generated—simply the use of a suitable step-down transformer—and the inherently attractive characteristics of an a.c. amplifier: implicit freedom from zero drift, insusceptibility to damage from accidental overload, and potentially high sensitivity. The higher the sensitivity of the amplifier, the less can be the test current for a given range of unknown resistance.

In the Electronic Instruments Model 47A Micro-ohmmeter (Fig. 3), the test current is generated in a high resistance network fed from a conventional transformer; the internal series resistance is never less than 125 times the maximum of whatever range is in use. Saturated

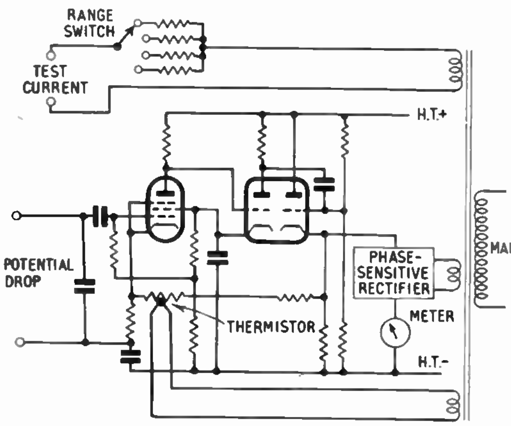

Fig. 3.—Micro-ohmmeter using alternating test currents.

regulator constant voltage transformers could have been used to avoid dependence of test current upon variations in mains voltage; their disadvantage is a high stray field, unacceptable if the potential drop amplifier is to combine very high gain with commercial simplicity. Instead, a current proportional to mains voltage and hence to test current is arranged to modify (inversely) the gain of the potential drop amplifier, using an indirectly heated thermistor in the feedback circuit, so arranged that over an adequate range the product of mains voltage and amplifier gain is constant, the system within these limits acting as a ratiometer.

In the amplifier, d.c. locking is applied to the first two stages to stabilize operating conditions to ensure maximum amplification unaffected by changes in steady state valve parameters; by this means, together with the use of a high gain low noise pentode in the input stage, the overall magnification from these two stages is about 20,000. Heavy degeneration is used to obtain stability in calibration; with feedback, the amplifier requires an input signal of 12 millivolts for full-scale deflection. On the lowest range, test current is 10 amperes, implying a resistance coverage from 0 to 1,200 microhms.

A phase-sensitive rectifier links the amplifier to the meter. Such a circuit responds only to that component of the input signal in phase with the test current, and can be used to suppress the effect of reactance, inevitably present to some degree in the test circuit. Generally, reactance at 50 c/s is negligible relative to resistance; exceptions where Q approaches or exceeds unity are large iron-cored chokes, drums of low resistance cable, and even single loops where the enclosed area is unusually large. Examples are installed lightning conductors together with the return test lead, and long stretches of bonded railway track.

2.4. *Instruments Using Pulsed Test Currents*

Recently a system has been developed which permits the measurement of samples where both expected resistance and power dissipative ability are exceptionally low. The dilemma here is that very high test currents are necessary to raise the signal voltage above the "noise" level while at the same time heavy continuous currents cannot be tolerated because they may burn out the sample. Using either d.c. or a.c. detection, it is difficult to reduce noise below about 50 μV; it may be due to thermo-electric potentials, to pick-up signals, or to drift in the amplifier itself. Where the test resistance is one microhm, to develop a voltage well above

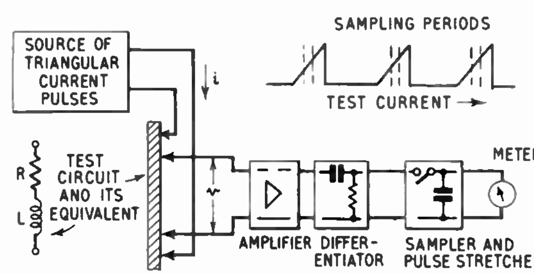

Fig. 4.—Micro-ohmmeter using triangular test current pulses.

the interference level may require a test current between 500 and 1,000 amperes and this latter current even in one microhm corresponds to a power of one watt, which may well be inadmissible. One solution which has been found satisfactory is shown in Fig. 4. The test current is generated as a series of isolated triangular pulses, each lasting for perhaps one millisecond and recurring at a rate of 10 c/s. The use of triangular pulses is dictated by the need to eliminate the effects of inductive reactance, crippling with any other form of pulsed wave-train, especially since, with such low resistance values, the Q of the test circuit may be high.

The action of the system may be followed from Fig. 4. If during the course of each pulse the test current i follows the law:

$$i = At$$

where with a triangular pulse A is constant, then the voltage v developed across a test circuit of resistance R and series inductance L is given by:

$$\begin{aligned} v &= Ri + L\mathrm{d}i/\mathrm{d}t \\ &= RAt + LA \end{aligned}$$

since $\mathrm{d}i/\mathrm{d}t = A$. If this signal is fed to a differentiating network, the resultant output signal v_o is seen to be:

$$v_o = B.\mathrm{d}v/\mathrm{d}t = BRA$$

where B is some constant which depends upon the performance of the differentiating network. It will be seen that the output voltage v_o is proportional to the series resistance R and that the component due to inductive reactance has been eliminated. In practice, measurement is made by sampling over a short period intermediate in each pulse, the indication being maintained during the long dead intervals by a pulse-stretching arrangement, thereby suppressing the severe disturbance at the end of each pulse. Since the magnitude of the signal v_o depends upon rate of change of test current, and not upon momentary amplitude, exact location in time of the sampling period is not required.

The merit of this system is the ability to apply test currents of very high momentary value with at the same time low mean thermal dissipation in the test sample because of the relatively long interval between successive pulses. It has been found possible to measure resistances of less than one microhm at a mean power level not exceeding 50 microwatts.

3. The Measurement of Very High Resistance

Instruments in this field, their range, it may be said, extending "as far as thought can reach," are now widely used in industry as well as in the laboratory. Up to 10^{16} ohms can be measured with fair accuracy; up to 10^{13} ohms within ± 1 per cent. All models use some form of electrometer amplifier, either measuring voltage across or current through the unknown, and these amplifiers are described in sub-section 3.2.

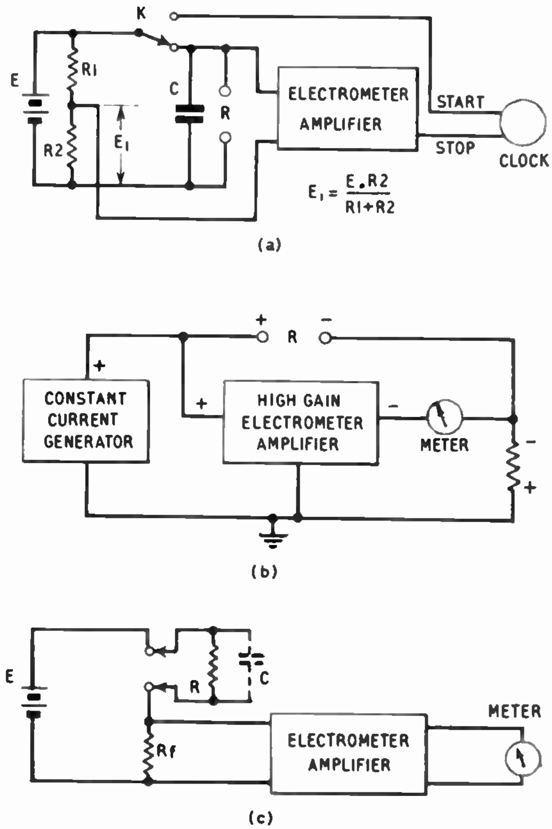

Fig. 5.—Three types of high resistance meter:
(a) capacitance-resistance discharge.
(b) constant current, variable voltage.
(c) constant voltage, variable current.

3.1. *Three Primary Systems*

These are: (*a*) the capacitance-resistance discharge type,
(*b*) the constant current, variable voltage type,
(*c*) the constant voltage, variable current type.

The last two are developments from the direct-reading ohmmeters shown in Fig. 1. The first can be thought of as a ratiometer in that it relates an observed change in voltage to a change in current deduced in terms of known parallel capacitance and measured time. By their very nature, neither types (*a*) nor (*b*) can be readily applied to the measurement of high resistance in parallel with high capacitance, as occurs in large capacitors and long lengths of cable, and this limitation seriously restricts their use in industry.

3.1.1. Capacitance-resistance discharge type (Fig. 5a)

In operation, a known capacitor C connected in parallel across the unknown resistance R is charged from a fixed voltage source E. The key K is then opened, and the capacitor C allowed partially to discharge through R. When the potential across it falls to some predetermined value E_1 corresponding to zero voltage input to the bridge-connected electrometer amplifier, a timing clock started when key K was opened is stopped and the time interval t noted. The value of resistance R can be deduced from:

$$\log_e E/E_1 = t/CR.$$

The use of a potentiometric comparison circuit together with null operation of the electrometer amplifier removes any need for accurate knowledge of either the absolute value of voltage E or gain of the electrometer. The parallel capacitance C need only be high enough to "swamp" variations in that of the unknown resistance; from 10 to 100 pF is usual. The fraction $(E - E_1)/E$ by which the initial voltage is allowed to run down can be as low as perhaps 2 per cent. to reduce both variation in test voltage during the test and its duration.

The Harwell 1055-B instrument (Fig. 6) covers resistance up to $1{\cdot}5 \times 10^{15}$ ohms, this with 12 pF in parallel and a run-down time of three minutes. Using larger capacitances and with shorter intervals, its bottom limit can be brought down to 10^8 ohms. Rate of decay is always measured rather than rate of growth of charge, to permit adequate electrification of the sample before the test; the resistance of some synthetic insulants can rise ten times during the first minute after applying the test voltage.

The advantages of this system are the ease with which high resistance can be measured accurately in terms of time, and the fact that test voltage can be low since all of it is available for the electrometer amplifier; cf. type (*c*) below. Its disadvantages are inconveniently long duration of the test, inability to detect rapid changes in resistance, and restriction to components themselves substantially free from parallel capacitance.

3.1.2. Constant current variable voltage system (Fig. 5b)

This type obviates the first two disadvantages above but retains the third. Another disadvantage is wide variability in test voltage, which for any given current is directly proportional to unknown resistance, and this may debar the measurement of highly voltage-sensitive resistors.

Fig. 6.—High Resistor Meter Harwell Type 1055-B. An advanced instrument in the capacitance-resistance discharge class, type (a).

The upper measurable limit is decided inversely by the magnitude of the constant test current and directly by the maximum potential drop which can be tolerated. Fig 5b illustrates the use of a Miller amplifier. Its merit is that degeneration can reduce equivalent input resistance and greatly simplify the design of the constant current circuit. This point is clear if it is remembered that very high gain in the electrometer amplifier implies infinitesimal input voltage to produce any given output; however high the unknown test resistance R, no significant voltage is developed across the input terminals of the amplifier and there is nothing to disturb the value of the test current. Another approach is to consider that the output circuit of the amplifier injects a voltage equal and opposite to that dropped across the unknown resistor and therefore leaves the equivalent voltage of the constant current source undisturbed.

This source can be a stable voltage in series with an equally stable fixed resistance which, with adequate degeneration in the electrometer system, need have no unreasonably high value. Cascode electronic circuits are in general unsuitable for the generation of very small currents; an interesting alternative is the saw-tooth differentiator. If a voltage source of low internal resistance, its output voltage E rising continuously at a uniform rate, is connected in series with a small fixed capacitance C to a load also of small impedance, the resultant current i is given by:

$$i=C.\mathrm{d}E/\mathrm{d}t$$

and is constant. Since E cannot in practice increase indefinitely it is periodically returned to zero, the source waveform being a very low frequency saw-tooth. The merit of this arrangement is that the minute current i can be uniquely defined in terms of capacitance and rate of change of voltage, both of which it is possible to specify with high initial accuracy and stability. For example, if C is one picofarad and if $\mathrm{d}E/\mathrm{d}t$ is 0·1 volt per second, test current is 10^{-13} ampere; it would be difficult stably and accurately to generate so small a current using resistive control.

Since potential drop across the unknown is proportional to resistance, the meter scale can be linearly calibrated in terms of resistance. High parallel capacitance is inadmissible because, with necessarily minute test currents, the time to charge it may be intolerably long.

3.1.3. Constant voltage variable current system (Fig. 5c)

This is widely used because it is suitable for the measurement of highly capacitive samples, such as capacitors and long lengths of cable.

If the unknown resistance R has a capacitance C in parallel, and if the test voltage E, though nearly constant, is drifting at $\mathrm{d}E/\mathrm{d}t$ volts per second due to imperfect stabilization, current through the test circuit will have two components, one due to leakage resistance:

$$\text{resistive component} \quad i_r=E/R \quad \ldots\ldots\ldots(1)$$

and one due to displacement of charge in the parallel capacitance C:

$$\text{displacement component} \quad i_d=C.\mathrm{d}E/\mathrm{d}t \quad \ldots(2)$$

Taking as an illustration values typical of a high grade capacitor where C is 10^{-6} farad and R is 10^{12} ohms, if the test voltage E is nominally 100 volts, then from (1) resistive current $i_r=10^{-10}$ ampere. If displacement current i_d is to be no greater, from (2) it follows that $\mathrm{d}E/\mathrm{d}t$ must not exceed 10^{-4} volt (1 part in 10^6) per second, no easy requirement.

Displacement current must be kept small since, in a 2-terminal network, it is inseparable from resistive current, and, if large, could mask it. It is therefore essential that the short term stability of the test voltage supply be exceedingly high; it is this factor which dominates the design of any practical instrument and which at present decides the upper limit to the measurable range.

In the circuit of Fig. 5c leakage current is measured in terms of voltage drop across a relatively low reference resistor Rf. Detection of current implies a meter calibration reciprocal in resistance or linear in conductance. If Rf is small, to preserve constancy in test voltage, so is the input voltage to the electrometer amplifier; it follows that sensitivity must be high. A commercial instrument based upon the constant voltage variable current system is described in section 3.4.

3.2. *The Electrometer Amplifier*

As is implied above, the specification for the electrometer amplifier varies with its application. In instrument types (*a*) and (*b*) the amplifier is connected directly across either part or the whole of the unknown resistance; its own input resistance, especially if variable, must be much higher than the maximum which is to be measured, but sensitivity can be low since the whole of the test voltage may be available as the input signal. In type (c) where both the reference resistance and the voltage developed across it are relatively low, amplifier resistance can be reduced while sensitivity must be increased, for any given test voltage.

Apart from the well-known d.c. valve amplifier, conversion devices are sometimes used, where the d.c. signal modulates an alternating voltage which, after heavy amplification, is rectified back to d.c. power. Modulator amplifiers have two advantages: isolation of the a.c. amplifier from the input circuit prevents the injection of grid current, and, using the powerful weapon of frequency discrimination, the electronic amplifier need contribute no significant zero instability or drift to the system. Two general points should be stressed. While improved zero stability can permit higher useful sensitivity, an increased upper limit to the measurable resistance range does not necessarily follow; extension may be prohibited by other factors, such as imperfect stability in test voltage. Secondly, the maximum resistance across which an electrometer amplifier can be connected is determined both by its own input resistance and by any input current which it may generate. The former is reducible by negative feedback, the latter is not; high amplifier gain can, however, be used to reduce it incidentally, by limiting the excursion of the input grid to the optimum part of the grid voltage-current characteristic.

3.2.1. D.c. valve amplifier

The d.c. valve amplifier is popular because it combines performance adequate for most applications with simplicity, reliability, and low first cost. Maximum useful sensitivity is governed by zero level stability, with good circuit design about ± 1 millivolt referred to the input terminals. The grid current of an electrometer valve can be as low as 10^{-15} ampere with negligible ohmic conductance; less expensive and more convenient semi-electrometer valves are available for use in type (*c*) instruments. One such amplifier is described in some detail in section 3.4.

3.2.2. Capacitor modulator

The capacitor modulator permits rather higher useful sensitivity because of its improved zero stability, from 2 to 10 times better than that of the d.c. valve amplifier. Input resistance varies with different constructions from 10^{13} to 10^{16} ohms. In Figure 7, the input signal E is fed through a high resistance R to a small parallel capacitor of mean capacitance C whose momentary capacitance is periodically varied above and below its mean value, either by a motor-driven vane or by a vibrating reed device, the reed being one plate of the capacitor. If variation frequency is f, usually between 100 and 500 c/s, and if the input time constant CR is long relative to $1/f$, then the mean charge $Q = EC$ remains substantially constant throughout each cycle. The momentary voltage across capacitor C therefore varies (inversely) in sympathy with the momentary capacitance, pulsating above and below its mean value E. The alternating component, proportional to d.c. input, is filtered to an a.c. amplifier, heavily amplified, and then rectified. A phase-sensitive demodulator is used to relate output polarity to that of the input and thereby permit a degeneration loop which includes both modulator and rectifier. Input resistance is determined only by the physical insulation of the modulator; residual instability of zero level is normally due to imperfections in the surfaces of the electrodes which form the variable capacitor.

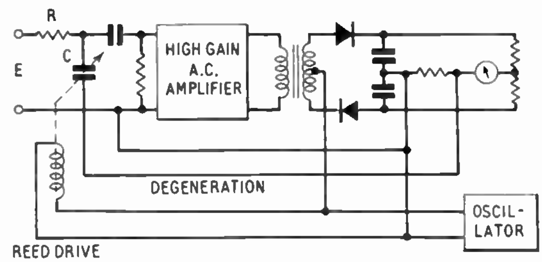

Fig. 7.—The capacitor modulator electrometer amplifier.

3.2.3. Mechanical or chopper modulator

This is also a converter device where the modulator is a switch which periodically connects the a.c. amplifier either to the d.c. input or to a short-circuit, generating a rectangular waveform, its peak-to-peak amplitude equal to that of the input signal. With careful design this system can display exceedingly good zero stability; $\pm 0{\cdot}1$ microvolt has been achieved commercially. Its disadvantages are limited life of the converter due to mechanical wear in the contact assembly, and fairly low input resistance, which effectively restricts the use of this system to type (*c*) instruments. Resistance is low partly because of the periodic discharge of the self-capacitance of the converter, and partly because the energy dissipated in the grid leak preceding the first valve in the a.c. amplifier must be derived from the d.c. input source. A mechanical modulator is essentially a switching device and, unlike the capacitance modulator, it cannot convert mechanical energy into electrical.

3.3. *Guarding and Sampling*

In the input circuit of the electrometer amplifier conductive paths are not, of course, confined to the active elements—the input valve or the modulator converter. They may arise in the wiring and in the insulation of any switching, round the terminals and the valve-holders, or across the electrode separator in the modulator capacitor. Two devices which can be used either singly or together to reduce such leakages are guarding and time sampling.

Guarding is the increase of equivalent resistance not by further raising its physical or "cold" value but rather by reduction of voltage across any leakage paths by circuit techniques. In a practical application to electrometer amplifiers, all leakage paths from the live side of the input circuit are arranged by suitable mechanical constructions—guard rings, screened wiring, isolated chassis—to terminate on a uni-potential conductive plane itself insulated from the base of the circuit. This conductive plane is connected to a voltage source of low impedance, its potential arranged to follow or "track" that of the live side of the input. To the extent that potential difference across the composite leakage path is reduced is the equivalent resistance of the system raised. A tracking point of low source impedance is naturally available in the degenerative Miller amplifier; here the reduction in input conductance is equal to the degeneration factor, which may well be several hundred times. There is no theoretical limit to the effectiveness of guarding; only the imperfect accuracy and stability of practical components need prevent it from being infinite.

The second device, periodic sampling, simply means intermittent connection of the electrometer amplifier to the input circuit, connection being periodically made and broken, perhaps by a pulsed high insulation relay. The lower the mark/space ratio, the lower the average current or energy absorption and the higher the equivalent input resistance. Some form of pulse stretcher such as a capacitor storage device is necessary to maintain the indication during blank periods. Sampling can not only raise equivalent resistance; it can also reduce the mean value of current injected by the amplifier into the input circuit, such as grid current in the d.c. coupled valve amplifier. It has already been mentioned that such currents cannot be directly reduced by degeneration since normally they are not linearly related to input voltage.

In instruments of type (*c*), using the constant voltage variable current system, it is important to note that while the requirement that amplifier input resistance should be high is relaxed, minimal grid current is as important as in any other system. Reduction in the value of the reference resistor relative to that of the unknown is not accompanied by any increase in current through it, which in all cases is determined only by the resistance under test and by the test voltage applied to it. It follows from this that is is important that any current generated in the input stage of the amplifier should be small compared to that in the unknown resistance, and that the degree of importance is not affected either by the value of the reference resistance or by the sensitivity of the amplifier connected across it.

3.4. *A Commercial High Resistance Meter*

In any instrument intended for general use there are requirements which do not trouble the designer of the laboratory model, operated only by skilled staff familiar with its adjustments and limitations. The commercial instrument must use internationally available valves without need for selection inside any one type. Its accuracy must be unaffected by reasonable changes in mains voltage or frequency or by humid and dusty atmospheres. It must not require frequent or complicated adjustment, and its calibration should be independent of ordinary variations in characteristics of valves throughout their life. Display of the reading must be direct and non-ambiguous; the instrument should be robust mechanically and insusceptible to damage from faulty connections or disadjustment of the range switch. Finally, the operator must be adequately protected from shock or injury arising from carelessness. A practical circuit is shown in Fig 8, which, though simplified, is essentially that of the Electronic Instruments Model 29A Megohmmeter.

The test potential of 500 volts is derived from an electronic stabilizer preceded by a saturated regulator (constant voltage) transformer. This latter device reduces amplitude variations about 30 times, the ratio of the percentage change in mains voltage to the percentage change in output voltage across any secondary. Unhappily, variations in mains frequency generate changes in output amplitudes with an exaggeration factor of roughly 1·3, a 2 per cent. shift in frequency producing a 2·6 per cent. change in output voltage. However, mains frequency usually varies much less than voltage and in any case can only change slowly, so that a substantial improvement in short term output stability is realized.

The electronic stabilizer is an example of the parallel type, presenting a high d.c. resistance and a very low incremental impedance between

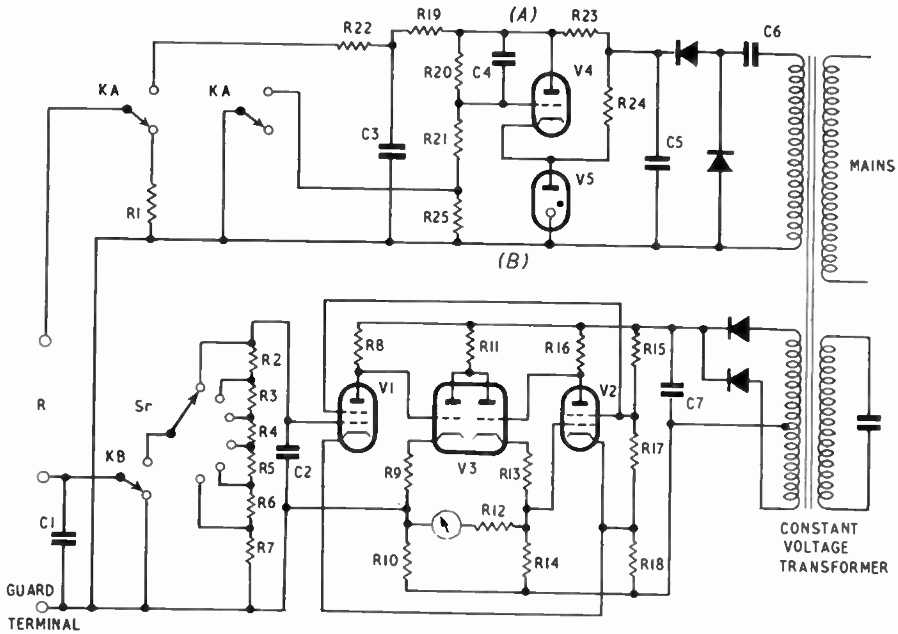

Fig. 8.—A commercial high resistance meter.

(A) and (B) in Fig. 8. If $N=(R_{20}+R_{21})/R_{21}$ is the reduction ratio of the potential divider which couples the anode of V4 to its grid, if g_m is the mutual conductance of V4 and if r_t is the incremental impedance of the high stability reference tube V5, it can be shown that the incremental impedance between (A) and (B) is given by $Z_{AB}=N(1/g_m+r_t)$ which with practical component values can be about 2000 ohms. If the series resistor R23 has a value of 100,000 ohms, fluctuations in the h.t. supply are reduced $R_{23}/Z_{AB}+1=51$ times. For rapid periodic variations such as ripple components, the presence of the capacitor C4 across R20 reduces the effective value of N nearly to unity and increases the suppression ratio of the electronic stabilizer as a whole to nearly 300 times. With a reasonably good mains supply the short term stability of the test voltage voltage between points (A) and (B) is better than 1 part in 100,000. Ripple components are further attenuated by resistor R19 and capacitor C3 while the final series resistor R22 is included to limit momentary short-circuit output current and thereby minimize risk of shock to the operator.

Key KA connects the live side of the test voltage to the unknown resistance while key KB transfers the other end of the unknown to the input of the electrometer amplifier. Separate keys are provided, partly to permit pre-charging of any capacitance across the unknown before the reading is taken, by operating KA before KB, and partly fully to engage the operator's hands while the test voltage is applied. Release of both keys automatically discharges the test circuit through the low resistance R1. Selection of range is effected by the range switch Sr which varies the effective value of the reference resistance through which the test current passes; to prevent zero level shifts as the range is changed, due to grid current in amplifier valve V1, an inverted-L arrangement is used to maintain the grid-cathode resistance of V1 at a nearly constant value. The reference resistance is always small relative to that of the unknown, never more than one part in 3000.

On this instrument the highest resistance which can be measured with acceptable accuracy is 2×10^{13} ohms, twenty million megohms. Through such a resistance the test potential of 500 volts drives a current of $2{\cdot}5\times10^{-11}$ ampere. On the highest resistance range the value of the reference resistance, the sum of R2 to R7, is 81 megohms, and from this it follows that the input signal to the amplifier, the product of test circuit current and reference resistance, is 2·02 millivolts for the maximum test resistance of 2×10^{13} ohms. Grid current from the input valve does not normally exceed 10^{-11} ampere and, provided that it remains constant, has no harmful effect upon either zero level or sensitivity. The amplifier consists of two nearly identical sections, each with a high gain pentode voltage amplifier followed by a voltage to current converter stage; their outputs are so coupled that the effects of power supply and component changes in the two sections are cancelled while their gains are additive, in that the right-hand section acts as an extremely low incremental resistance across the output bridge resistor R14 and permits a bridge efficiency very close to unity. The high sensitivity of the system permits the application of heavy degenerative feedback from R12 together with the meter, directly to the input circuit. The meter is protected from damage by overload—e.g., should the test terminals be connected to a short circuit—since the current through it must always be less than the standing anode current in the left-hand section of the double-triode V3, which in practice can be kept down to a safe value. Not shown in the simplified circuit of Fig. 8 are controls to adjust the zero level of the electrometer amplifier, corresponding to zero input signal and hence infinite test resistance. The output of the amplifier is taken to a meter reciprocally calibrated in terms of resistance; it could equally well be graduated linearly in conductance.

Clearly enough, the upper limit to the range of instruments of this type is governed by magnitude of test potential, by the sensitivity of the electrometer, and by the maximum value of reference resistance which can be connected across it. For the first, 500 volts is widely accepted internationally as a suitable reference value; higher voltages may in any case be inadmissible because of the risk of breakdown in components such as capacitors. Stability of test potential decides the value of higher electrometer sensitivity, an increase serving no useful purpose if it further amplifies fluctuations. A higher reference resistor is practicable if either a full electrometer valve or a capacitor modulator is used, to reduce or obviate grid current. It is, however, found that reference resistances above 500 megohms introduce serious inconvenience both in the construction of the instrument and in its use.

4. Acknowledgments

Information about Harwell instruments and invaluable assistance generally was given by Mr. H. Lysons of the United Kingdom Atomic Energy Authority. The Pye "Microvoltmeter" and the Evershed and Vignoles "Ducter" are, of course, the proprietary products of those companies. As regards instruments with which the author has been concerned, the Model 47A microhmmeter and the Model 29A megohmmeter were developed for Electronic Instruments Limited, to whom thanks are given for permission to describe the work, while the microhmmeter system using pulsed test currents was designed at the Massachusetts Institute of Technology in connection with the measurement of semi-conductive materials.

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RESEARCH IN GREAT BRITAIN

N.P.L. Open Day

Although most of the purely radio research undertaken by D.S.I.R. is carried out elsewhere (by the Radio Research Organization at Slough), there is a considerable amount of work in the more general field of electronic engineering taking place at the National Physical Laboratory at Teddington. The President and other representatives of the Institution invited to attend the recent Open Day thus found much to interest them in every one of the nine Divisions of the Laboratory.

The newly established Control Mechanisms and Electronics Division is currently constructing a new high speed computer (ACE) which will embody experienced gained in logical and engineering techniques from the experimental ACE pilot model and its engineered version DEUCE. The new ACE will operate at a higher speed and will have a considerably larger store than its predecessors—1½ million digits in the mercury delay lines and magnetic drums. The C.M.E. Division has also constructed versatile electronic analogue computers which enable the operation of servomechanisms to be simulated rapidly and accurately, and the responses to be displayed on c.r. tubes.

Ever since its foundation in 1900, the N.P.L.'s main concern has been in the precise measurement of physical constants, and the Electricity Division is collaborating with the Royal Greenwich Observatory in measuring the variations in the rate of rotation of the earth in terms of an atomic standard. Mention has already been made in the *Journal* for May 1955 of the caesium atomic frequency standard to be built by the Massachusetts Institute of Technology; the basically similar N.P.L. instrument uses highly developed microwave techniques in dividing the natural frequency (9192·632 Mc/s) of the caesium atom for checking standard frequencies such as those radiated by the Post Office's Rugby station.

While most of the Divisions use electronic instruments for measurement purposes, some of the ultrasonics work being done in the Physics Division employs techniques which may well have wider application—here equipment has been developed for displaying the Argand diagram of transducers up to megacycle frequencies which would be suitable for other types of complex networks.

Atomic Energy Research at Harwell

Recent changes in the security classification of atomic energy research projects enabled the Atomic Energy Research Establishment to show some of the achievements during the first ten years of its operation at a recent Open Day. A striking feature of the research work, which embraces the fields of pure and applied physics and chemistry and mathematics, is the wide use of electronic techniques.

In the Electronics Division, a number of items were exhibited which highlighted the important uses of electronics. A notable Harwell instrument is the transistor digital computer CADET, which has a power consumption of only 70 watts and will eventually be fitted in two 19-in. racks, 6 ft. high; in the programme demonstrated the computer searched for numbers equal to the sum of the cubes of their digits. Analogue computers were also shown in the form of reactor simulators, which are of great value in design and in the training of operators.

Among the other exhibits were many electronic instruments, for instance for geological surveying, for clinical and industrial use, as well as items such as scintillation counter spectrometers, used in the fundamental investigations into radioactive substances.

An Industrial Research Laboratory

While many of the largest companies engaged in work in the electronics and allied fields maintain their own research departments, smaller firms have in the past tended to rely on outside research bodies for investigation outside the immediate matter in hand. It is therefore of interest to record the setting up of research and development laboratories by the Solartron Electronic Group Ltd., which are both administratively and geographically separate from the main manufacturing centre.

The new laboratories are in a large mansion outside Dorking, and a representative of the Institution was recently invited to visit them. The work to be undertaken will include industrial applications such as the testing of control devices, and the design of computers and training devices. The development of an electronic reading machine having considerable potentialities for business equipment, is also being carried out.

FERRITES IN WAVEGUIDES *

by

G. H. B. Thompson, B.A. †

SUMMARY

The gyromagnetic mechanism which controls the permeability of a ferrite at microwave frequencies is investigated theoretically in several simple cases, and the tensor form of the permeability in the general case is indicated. The theory is first applied to circular waveguides containing a ferrite magnetized along the axis. Experimental results are presented which confirm the expected presence of resonance absorption but also show low field losses. A possible explanation of these losses is suggested. Applications of the effects in the construction of isolators and circulators are described. These include resonance and reflection isolators for circularly polarized modes, and also devices which make use of Faraday rotation. The effects of transverse magnetization of the ferrite is then discussed. Applications in rectangular guide in the form of resonance isolators, non-reciprocal phase circulators and field displacement devices are described. Bi-refringent behaviour in circular guide is also described. The relative merits of the different types of isolator and circulator are compared at different frequencies in terms of ease of construction, and performance at one frequency and over a band. The characteristics of most of them can be theoretically calculated from the components of the permeability tensor, and the last part of the paper is devoted to methods for measuring these.

LIST OF PRINCIPAL SYMBOLS

B	Magnetic induction
H	Total magnetic field
h	Small r.f. component of magnetic field
k	Off-diagonal component of permeability tensor of ferrite expressed as multiple of free space permeability
M	Total magnetization per unit volume
m	Small r.f. component of magnetization per unit volume
$N_{x,y,z}$	Shape demagnetizing factors
P	Electric polarization per unit volume
R	Radius of cylindrical waveguide
β	Wave propogation constant
β_0	,, ,, ,, in free space
β_m	,, ,, ,, in infinite ferrite
β_g	,, ,, ,, in waveguide
γ	Gyromagnetic ratio
ε_0	Permittivity of free space
ε	Dielectric constant
θ	Differential phase delay per unit length for the two modes supported by a guide containing ferrite
μ_0	Permeability of free space
μ	Diagonal term in permeability matrix of ferrite expressed as a multiple of the free space permeability
μ_{eff}	Effective transverse permeability of ferrite in the same units
φ	Faraday rotation per unit length
ω	Angular frequency of wave
ω_0	Resonance frequency in ferrite

Rationalized M.K.S. units are used throughout this paper.

1. Introduction

The materials normally dealt with in waveguide work are either conductors or dielectrics, and their permeability is almost universally unity. The ferrites owe their unique position to the fact that they are almost as strongly ferromagnetic as iron, and yet sufficiently non-lossy to support a propagating electromagnetic wave. The permeability of unmagnetized ferrites at microwave frequencies is not necessarily very large since the magnetic effects are caused by spinning electrons, and the gyroscopic forces acting on these tend to resist rapid changes in orientation However a gyroscope under the action of a constant force precesses at a steady rate, and so do these electrons if a steady magnetic field is applied.

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The quantum-mechanical forces which link the electron spins in a ferromagnetic material ensure that they all precess together, and so the magnetic moment of the ferrite as a whole swings about the field direction. This is in fact a very short-lived effect, and dies away quickly after the application of the magnetic field. It can be sustained however by an incident wave of similar frequency, and this condition is called ferromagnetic resonance. At and near this point there is a great deal of reaction on the wave, and the effective permeability is considerable. The great practical importance, however, of the ferrites lies not so much in the possibility there exists of controlling the properties of the material with the applied magnetic field, but in the fact that the effects are non-reciprocal. A certain spatial direction of gyromagnetic precession is created by the field, but from the point of view of the incident wave it depends on the side of approach whether this appears right-handed or left-handed. If it can be arranged that in the region of the ferrite the field components of the wave are circularly polarized, then it is possible to construct a device which behaves differently for the two directions of propagation.

In addition to the non-reciprocal behaviour of both the phase and amplitude of the transmission coefficient it is possible to obtain, even under substantially loss free conditions, reflection coefficients which are non-reciprocal in amplitude. A large and entirely new range of microwave components has been developed by applying these effects not only to two-port devices, but also to waveguide junctions of three or more arms.

2. Description of Ferrites[24]

The ferrites are mixed oxides of trivalent iron with a divalent metal such as manganese, nickel, zinc, copper, cobalt, etc.; in some cases the iron may be partially replaced by trivalent aluminium. They are characterized more by their crystal structure, which is of the spinel type, than by their constituents. In Fig. 1, which is a diagram of the unit cell, the large spheres represent the close-packed oxygen atoms, and the smaller spheres marked A and B represent the metal atoms which fit in the interstices. In a magnetic ferrite half the trivalent atoms fill the sites B, whilst the sites A are filled both by the remaining half, and also by all the divalent atoms. Besides the pure ferrites which contain only two metals, mixed ferrites can be produced. Thus the divalent ion can be replaced stage by stage through the structure by another divalent ion, and a continuous solid solution exists between the two ideal end compositions. The conductivity of the ferrites is very low because the metallic conduction electrons are now occupied in holding the structure together. If, however, there is any excess iron in the lattice, it will be in the divalent form, and the extra electron introduced by each atom is capable of moving around amongst the original iron atoms and considerably increasing the conductivity.

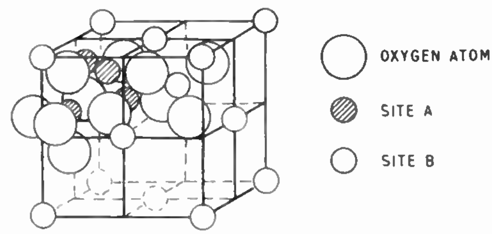

Fig. 1.—Unit cell of the spinel lattice.

Thus it is an advantage to keep the iron content slightly short. The magnetic properties are due to the uncompensated spins of the outer electrons of the metal ions. The saturation, of the order of a few thousand gauss, is considerably less than that of iron on account of the antiparallel coupling of some of these spins, and in zinc ferrite which is nonmagnetic this cancellation is complete. However, in most other cases there is partial parallel coupling, and the materials are ferromagnetic.

3. Circularly Polarized Field

3.1. *Theory*

The precessional motion of a spinning electron in a steady magnetic field is illustrated in Fig. 2a for the case where the damping forces are negligible. The motion is described by the normal gyroscopic equation, which relates the rate of change of angular momentum and hence of magnetic moment, to the applied couple. For a ferromagnetic substance, where all the electrons are coupled by exchange forces, their angular momenta and magnetic moments can be summed and a general equation of magnetic motion can be obtained.[1] Thus if **M** is the total magnetic moment per unit volume, **H** is the field and γ is the ratio of the magnetic moment to the angular momentum of the electron, the equation becomes:

$$d\mathbf{M}/dt = -\gamma\mathbf{M}\times\mathbf{H} \qquad \text{.........(1)}$$

This equation will be investigated in a rectangular coordinate system chosen so that the static magnetic field lies along the z direction. In the presence of an r.f. field **H** is no longer constant. A small high frequency component applied along the z direction only produces very slight perturbations in the precessional frequency, but if it is applied along any direction at right angles it causes an oscillation of the axis of precession as shown in Fig. 2b and, when the two frequencies are similar, can make the magnetic moment swing out considerably from the field direction. On account of the symmetry about the z axis it is simplest to consider the case of a circularly polarized TEM wave propagating along this axis with its transverse components of field rotating in the x-y plane. All other types of polarization in the same plane can be synthesized from appropriate combinations of left-handed and right-handed circularly polarized components, and the treatment is thus quite general. The transverse components of the magnetization and magnetic field will be denoted by the vectors $\mathbf{m}_t$ and $\mathbf{h}_t$ respectively. In a circularly polarized

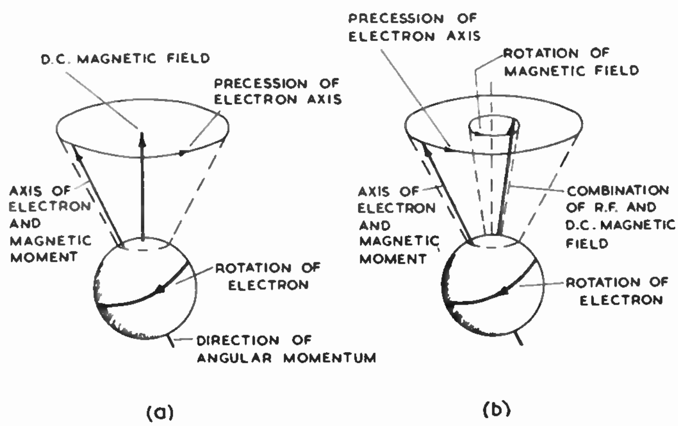

Fig. 2.—Precession of spinning electron. (a) Precession of electron in steady (d.c.) magnetic field. (b) Precession of electron in a combination of a d.c. and a transverse circular r.f. magnetic field for positive polarization.

field these vectors are constant in magnitude and rotate with angular velocity $\pm\,\omega$ about the z axis. Additional vectors $^*\mathbf{m}_t$ and $^*\mathbf{h}_t$ will be used, where the starring operation represents a rotation of the original vector by an angle of $-\pi/2$ about the z axis. For the case where $\mathbf{m}_t$ and $\mathbf{h}_t$ are small compared with $\mathbf{M}_z$ and $\mathbf{H}_z$ and the latter can be considered constant, eq. (1) can be rewritten

$$d\mathbf{m}_t/dt = -\gamma(\mathbf{m}_t \times \mathbf{H}_z + \mathbf{M}_z \times \mathbf{h}_t)$$

or

$$\pm\,\omega\,{}^*\mathbf{m}_t = \gamma({}^*\mathbf{m}_t\,H_z - M_z\,{}^*\mathbf{h}_t) \quad \ldots\ldots\ldots(2)$$

where the positive sign refers to right-handed rotation and the negative sign to left-handed. By re-arranging eq. (2) one can obtain the transverse magnetization in terms of the transverse field, and hence the effective permeability μ_{eff} of the material. Thus:

$$\mathbf{m}_t = \left\{\frac{\gamma M_z}{\gamma H_z \mp \omega}\right\}\mathbf{h}_t \text{ for } \pm\,\omega \quad \ldots\ldots\ldots(3)$$

and

$$\mathbf{B}_t = (\mu_o\mathbf{h}_t + \mathbf{m}_t) = \mu_o\mu_{eff}\mathbf{h}_t$$

where for $\pm\,\omega$,

$$\mu_{eff} = 1 + \frac{\gamma M_z}{\mu_o(\gamma H_z \mp \omega)} = 1 + \frac{\gamma(\mp M_z)}{\mu_o[\gamma(\mp H_z) + \omega]} \quad \ldots\ldots\ldots(4)$$

The effective permeability is therefore a scalar quantity and relates B and H in the normal way. It will be seen from the two expressions for the effective permeability that altering the direction of rotation is exactly equivalent to reversing the steady field direction, with the consequent reversal of magnetization. In the following treatment positive fields will apply to the case where ω and H are of the same sign and negative fields will apply to the case where they are of opposite sign, and ω will be assumed always positive. In Fig. 3, μ_{eff} is plotted against steady magnetic field, according to eq. (4) with, however, a damping term included, for a typical ferrite at 10,000 Mc/s. Up to saturation the equation is not really valid, because the electron spins are not all aligned. Beyond saturation it will be noticed that resonance only occurs for positive steady fields, in which case the r.f. vector is rotating in the same direction as the electron spin precessian, and the field required for resonance is equal to ω/γ. For positive fields less than resonance the r.f. magnetization is oppositely directed to the r.f. magnetic field and the permeability is less than unity. When the steady field is increased beyond the inversion point, as shown in Fig. 3, the magnetization outweighs the magnetic field and the permeability becomes negative. At resonance the magnetization makes a considerable angle with the applied steady field and the damping forces result in a large absorption of energy, as shown by the curve of the imaginary part of the effective permeability μ'_{eff}. In the region beyond resonance the magnetization rotates in phase with the magnetic field and the permeability is positive and greater than unity. This applies also to the whole region of negative steady field. At high fields, both positive and negative, the permeability tends to unity.

In fact circular polarization of the field in the entire region of the ferrite is only possible if the ferrite is either infinite, or occupies a small region in the centre of a guide where the r.f. fields are entirely transverse. The boundary conditions at the edge of a guide, for instance, require that there be no normal component of magnetic induction, and hence any transverse component must be plane polarized parallel to the wall. For an infinite ferrite the propagation constant is proportional, in the normal way, to the square root of the permeability. For positive steady fields the propagation constant therefore decreases as the field is increased, and becomes imaginary when the permeability goes negative. In this region the wave is evanescent. For fields beyond resonance the wave starts propagating again with very short wavelength which increases as the field is increased still further. For negative fields the wave propagates all the time and the wavelength is not greatly affected by the field. The general behaviour of a rod in the centre of an H_{11} circular waveguide is similar, but in this case one is interested in the relation between, not the internal, but the external r.f. field and the magnetization. On account of the demagnetizing effects in the rod which arise from effective magnetic poles being produced on its surface[2] the external field $\mathbf{h}_t'$ is related to the internal field $\mathbf{h}_t$ as follows: $\mathbf{h}_t' = \mathbf{h}_t + \mathbf{m}_t/2\mu_0$.
As will be shown later (Sect. 5.1) resonance now occurs at a frequency ω_0 given by

$$\omega_0 = \gamma\left(H_z + \frac{M_z}{2\mu_0}\right)$$

Also the change in propagation constant produced by putting a thin ferrite rod down the centre of a circular guide for the H_{11} mode can be shown[3,4,5] to be given approximately by the following equation:

$$\delta\beta_g = \left[\frac{\pi U\, N_1'(U)}{4\, J_1''(U)}\right]\frac{1}{\beta_g}\left(\frac{r}{R}\right)^2\left\{\beta_g^2\,\frac{\gamma M_z}{\mu_0(\omega_0 \pm \omega)} + \beta_0^2\,\frac{\varepsilon - 1}{\varepsilon + 1}\right\} \qquad \text{.........(5)}$$

where r is the radius of the ferrite, R the radius of the guide, ω_0 the resonance frequency, β_g the guide propagation constant, β_0 the free space propagation constant, ε the dielectric constant, and U is given by the first root of: $J_1'(U) = 0$. The resonance frequency for a rod is thus different from that in an infinite ferrite, and no evanescent waves appear, but otherwise the behaviour with the former is just a diluted version of that with the latter.

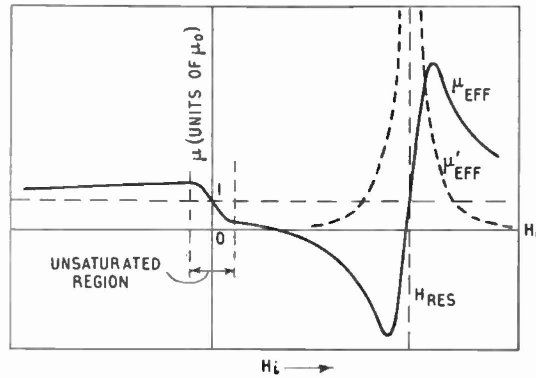

Fig. 3.—Ferrite permeability for circularly polarized waves as a function of internal magnetic field.

This illustration and Figs. 4, 8, 13, 14, 15a, 16, 17, 19, 20, 23, 24 are reproduced from Ref. 8.

For the exact solution of propagation in a waveguide filled with ferrite magnetized along the z direction it is possible to deduce from eq. (1) the following general relations between the three components of induction, $\mathbf{B}_x$, $\mathbf{B}_y$ and $\mathbf{B}_z$, and the three components of magnetic field $\mathbf{H}_x$, $\mathbf{H}_y$ and $\mathbf{H}_z$:[1]

$$\begin{aligned}\mathbf{B}_x &= \mu_0(\mu\,\mathbf{H}_x - jk\,\mathbf{H}_y)\\ \mathbf{B}_y &= \mu_0(\mu\,\mathbf{H}_y + jk\,\mathbf{H}_x) \qquad \text{.........(6)}\\ \mathbf{B}_z &= \mu_0(\mu_z\,\mathbf{H}_z)\end{aligned}$$

where

$$\mu = 1 - \frac{\gamma^2 M_z H_z}{\mu_0(\omega^2 - (\gamma H_z)^2)} \qquad k = \frac{\gamma M_z}{\mu_0[\omega^2 - (\gamma H_z)^2]}$$

$\mu_z = 1$ for a saturated ferrite.

These can be written more compactly in tensor notation as follows:

$$\mathbf{B} = \mu_0 \begin{Vmatrix} \mu & -jk & 0 \\ jk & \mu & 0 \\ 0 & 0 & \mu_z \end{Vmatrix} \mathbf{H};$$

the term enclosed between the lines is called the tensor permeability of the ferrite.

A general wave equation can be derived by inserting these relations into Maxwell's equations. This can be solved numerically for the appropriate boundary conditions of the guide concerned to give the propagation constant and the field patterns.[3,4,6,7] A neat graphical method has been evolved by Suhl and Walker,[7] who have classified possible modes.

3.2. *Experimental Results*

Van Trier[3] carried out some measurements on the change in propagation constant produced by very thin ferrite rods in an H_{11} circular waveguide, using a resonance cavity technique. He found that, provided the radius of the ferrite was less than about a twentieth of that of the guide, good agreement was obtained with the approximate theory quoted above. For most practical applications this is too thin to produce a useful effect. Fig. 4 shows the measured attenuation of a ferrite rod of more usual diameter plotted against the applied steady field[8] and it can be seen that the behaviour departs slightly from the ideal. Thus some resonance absorption occurs at negative fields. A circularly polarized wave was used, but as has been shown above, the magnetic vector itself is only truly circularly polarized at the centre of the guide.

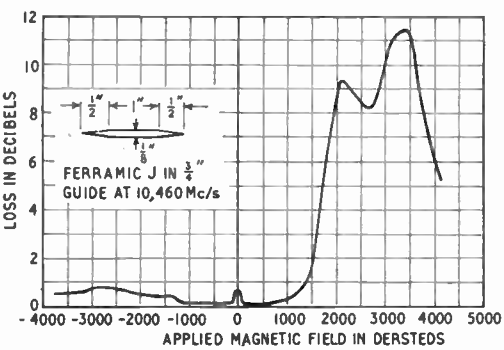

Fig. 4.—Curve showing ferromagnetic resonance in ferrite-loaded circular guide.

At other points a certain proportion of the opposite circular polarization is present and this can induce resonance in a negative field. The second curious feature is the considerable loss which occurs at low fields, in the region where the ferrite is unsaturated. This is typical of most ferrites at frequencies below about 10,000 Mc/s. It can be attributed to demagnetizing effects which allow resonance to take place in zero applied field. Fig. 5 illustrates a model due to Polder and Smit[9] which shows an extreme case of demagnetization that can occur in an unsaturated ferrite if adjacent domains are magnetized in the opposite direction. A cross-section of a rod is shown, and the domains which are magnetized alternately into and out of the paper, are in the form of thin laminae set vertically side by side. The two diagrams represent the state of the magnetization at moments in time separated by

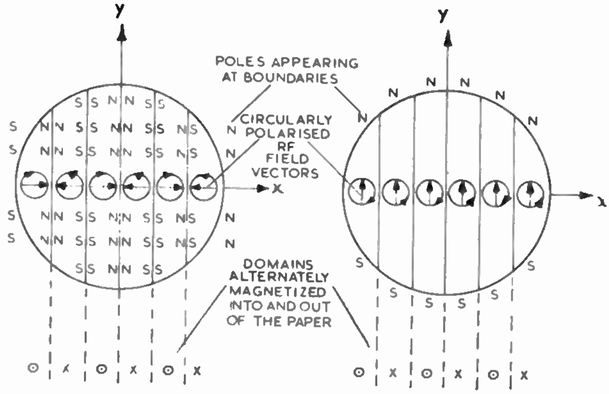

Fig. 5.—Demagnetization across the section of an unmagnetized ferrite rod with parallel domains.

a quarter period when an r.f. magnetic field parallel to the y axis is incident on the specimen. It will be seen that the magnetization precesses in opposite directions in the two types of domain, and is so phased to produce no nett magnetization in the x direction, whilst giving a sinusoidal variation in the y direction. Free magnetic poles marked N and S thus appear successively at the boundary of the ferrite and at the domain walls. The respective demagnetization factor has the normal value for a rod of $\frac{1}{2}$ in the y direction, but on account of the opposition of the magnetization vectors, it has the value of 2 in the x direction. As will be shown later in Sect. 5.1, the resonance frequency is given in terms of the demagnetization factors N_x and N_y as follows:

$$\omega = \gamma\left\{\left[H_z + N_x\frac{M_z}{\mu_0}\right]\left[H_z + N_y\frac{M_z}{\mu_0}\right]\right\}^{\frac{1}{2}}$$

Inserting the values given above and putting the applied field equal to zero; one obtains a natural resonance frequency given by: $\omega = \gamma M / \mu_0$ This is the maximum which can be obtained for any orientation of domain and roughly corresponds to the highest frequency at which low field loss is observed. Large internal anisotropy fields can, of course, increase this frequency.

4. Isolators and Circulators in Circular Guide

4.1. *Circular Polarization*

It has been shown above that reversing the magnetic field has a considerable effect on the propagation characteristics. The same result is of course produced by keeping the field constant and reversing the direction of propogation. This effect can be applied to the construction of non-reciprocal isolators, i.e. waveguides that transmit in one direction only. From Fig. 4 it can be seen that when a resonance field is applied to a partially-loaded circular guide, a wave with one circular polarization is highly attenuated when travelling in one direction and only slightly attenuated in the other. Fig. 6 shows a practical arrangement containing circular to rectangular tapers and quarter wave plate circular polarizers,[10] which gives non-reciprocal properties in a rectangular waveguide run. This device can give 50 to 1 ratios in forward to backward attenuation, but it is not very practical on account of the high magnetic fields that are necessary. An alternative is to use a ferrite loading so that the backward wave is operating in the region of negative effective permeability shown in Fig. 3, and is entirely reflected. High fields are not necessary in this case for frequencies up to about 15,000 Mc/s. The absorption of the reflected wave can be

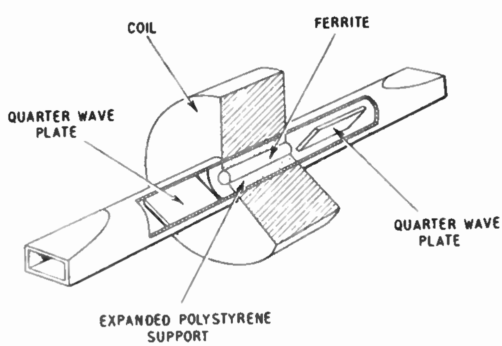

Fig. 6.—Resonance absorption isolator.

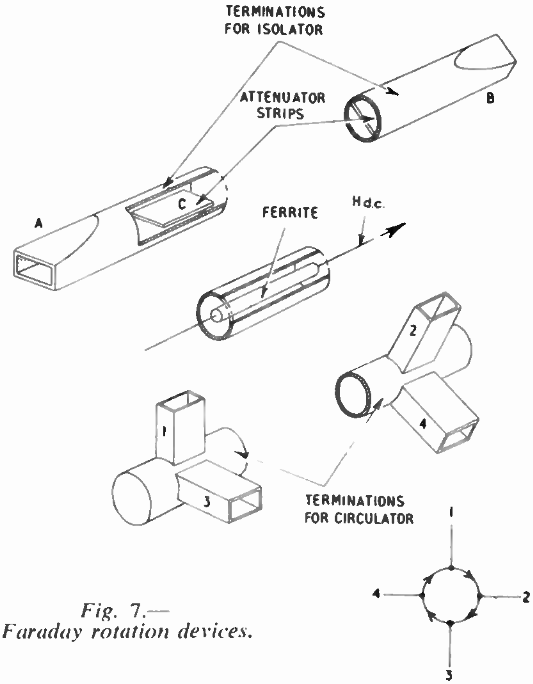

Fig. 7.—Faraday rotation devices.

arranged to occur in a resistive vane set at right angles to the incident electric field on the input side of the polarizer, since the reflected wave emerges with its plane of polarization rotated through one right angle.

4.2. *Faraday Rotation*

A different form of isolator can be constructed by allowing a wave of linear polarization to impinge on the ferrite.[11] Linear polarization can be resolved into two oppositely rotating circular components which propagate through the ferrite with different velocities. If their phases are altered relative to one another by an angle θ, and their amplitudes are unchanged, they recombine on leaving the ferrite to give linear polarization whose plane has been rotated by an angle of $\theta/2$ from that of the input. For steady fields less than the resonance field the rotation is in the same direction as the positive current producing the steady field, for both directions of wave propagation. However, when viewed in the propagation direction, rather than from a constant position, the sense of rotation is opposite for the two waves. The upper part of Fig. 7 illustrates an isolator which operates on this principle. The ferrite is arranged to rotate the plane of polarization by

45 deg., and the input and output guides are also set at 45 deg. to each other. A wave entering at A will be rotated by the correct amount to leave at B without reflection. The reverse wave however will be rotated by the ferrite so that it arrives at A with its electric vector at 90 deg. to the direction this guide will propagate, and it can be absorbed by a resistive strip suitably positioned at C. The lower part of Fig. 7 illustrates a development of this device called a circulator in which four waveguide ports are provided. The diagram shows the coupling between ports. Thus arm 1 couples to 2, 2 to 3, 3 to 4 and 4 to − 1.

The Faraday rotation devices have the advantage that only small magnetic fields are required. In Fig. 3 it can be seen that a large difference between the effective permeability in positive and negative fields can be obtained even in the unsaturated region of the ferrite. For an infinite ferrite the propagation constant β_m is proportional to the square root of the permeability. Using the values for the two permeabilities given in eq. (4) the following relation for the rotation per unit length, φ, is obtained:

$$\varphi=\theta/2=\tfrac{1}{2}(\beta_{m-}-\beta_{m+})$$

$$=\tfrac{1}{2}\beta_0\sqrt{\varepsilon}\left[\left\{1+\frac{\gamma M_z}{\mu_0(\omega+\gamma H_z)}\right\}^{\frac{1}{2}}-\left\{1-\frac{\gamma M_z}{\mu_0(\omega-\gamma H_z)}\right\}^{\frac{1}{2}}\right] \quad \ldots\ldots\ldots(7)$$

For $\omega \gg \gamma H_z$: $\varphi=\tfrac{1}{2}\gamma M_z\sqrt{\varepsilon\,\varepsilon_o/\mu_0}$ ....(8)

And for $\gamma H_z \gg \omega$ it becomes

$$\varphi=-\tfrac{1}{2}\,\frac{\omega^2 M_z}{\gamma\,(H_z M_z)^{\frac{1}{2}}\,H_z^{3/2}}\sqrt{\varepsilon\,\varepsilon_o/\mu_o} \quad \ldots\ldots\ldots(9)$$

For a waveguide filled with a thin pencil of ferrite one can use the propagation constants given by eq. (5) and obtain φ as follows:

$$\varphi \propto \left(\frac{r}{R}\right)^2 \beta\left\{\frac{\gamma M_z}{\mu_o(\omega+\omega_o)}+\frac{\gamma M_z}{\mu_o(\omega-\omega_o)}\right\} \quad (10)$$

where $\omega_o=\gamma\left(H_z+\dfrac{M_z}{2\mu_o}\right)$

For $\omega \gg \omega_0$ this gives: $\varphi \propto \dfrac{\beta_g}{\beta_0} M_z\left(\dfrac{r}{R}\right)^2$

And for $\omega \gg \omega$: $\varphi \propto \dfrac{\beta_g}{\beta_0}\,\dfrac{\omega^2 M_z}{\omega_o^2}\left(\dfrac{r}{R}\right)^2$

It can be seen from these equations that when the field is below that required for resonance the rotation depends very little on frequency, and in the case of the infinite ferrite it is

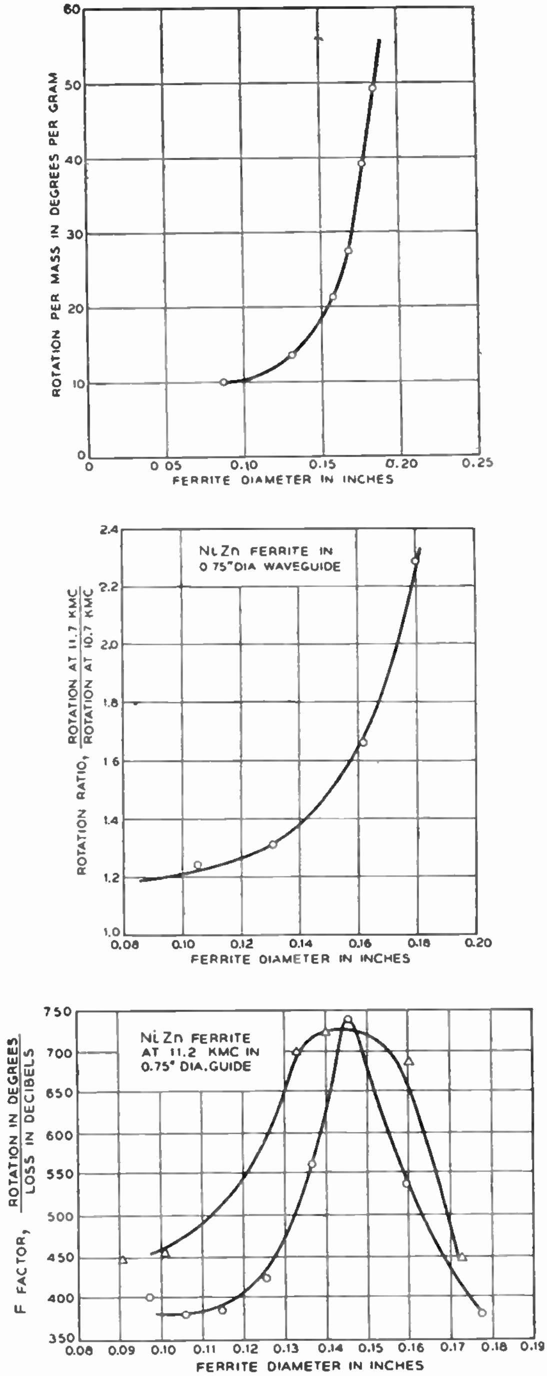

Fig. 8.—Effect of ferrite rod diameter on Faraday rotation characteristics.

independent of frequency. In this region the relation between rotation and field is similar to the magnetization curve of the material. However when the field is greater than the resonance field, or the frequency is less than the resonance frequency, the rotation is proportional to the square of the frequency. It is therefore the usual practice not to work in this range, and to keep the resonance frequency well below the operating frequency. In practice in order to get 45 deg. rotation in a short length of ferrite it is usually necessary to use a rod too thick for eq. (10) to apply. In these circumstances the r.f. field starts concentrating considerably in the rod, and the rotation increases much more rapidly with rod size. This concentration, and hence the rotation also, is frequency sensitive, and the effect can be seen in the experimental curves of Fig. 8 taken from results obtained at the Bell Laboratories.[8] The rotation per unit volume or mass, which for a small sized ferrite is constant, rises rapidly after a certain size is reached. The equivalent increase in frequency sensitivity is also shown. The curve of rotation per unit loss shows that a relatively large ferrite rod is required to give best perfomance in this respect. The choice of ferrite size is thus a compromise between minimum loss, minimum frequency dependence, and minimum physical dimensions. In addition if the ferrite is too large higher order modes can propagate and considerable trouble may be experienced with dimensional resonances giving sharp absorption or reflection losses at particular values of magnetic field.[12]

4.3. *Field Displacement*

The concentration of r.f. energy in ferrite rods of larger radius is confined, in the earlier stages, mainly to the negative wave for which the permeability of the ferrite appears positive (see Fig. 3), whilst the positive wave travels in the region around the ferrite.[12] The possibility exists therefore of selectively attenuating the negative wave by incorporating in the ferrite rod a film of lossy material, and an isolator has been constructed to work on this principle.[13] The bandwidth is found to be superior to that of a Faraday rotation device, and an isolation of better than 15 db for an insertion loss of less than 1·5 db is obtained over a 30 per cent. band about a centre frequency of 10,000 Mc/s. The optimum diameter of the ferrite is inversely proportional to frequency, but when chosen correctly it appears to give simultaneously maximum loss for the positive wave, and minimum loss for the negative wave. The forward loss in this type of isolator is not so dependent on the ferrite losses as in the previous types, since the wave concerned does not propagate through the ferrite.

5. Transverse Fields

The preceding treatment has been concerned only with waves propagating along the direction of magnetization in the ferrite. However any wave which produces a component of magnetic field at right angles to this direction will cause gyromagnetic precession, and this applies to the case shown in rectangular guide in Fig. 9, where the magnetization is at right angles to the direction of propagation and to the plane of polarization of the r.f. magnetic field. The permeability tensor with the system of axes shown is written:

$$\mu_0 \begin{Vmatrix} \mu_x & 0 & 0 \\ 0 & \mu & -jk \\ 0 & jk & \mu \end{Vmatrix}$$

This can be substituted into Maxwell's equations, and solutions for the field distribution and the propagation constant obtained. The latter, in the case where there is no variation of the field quantities in the y direction, behaves[3, 14] as if the ferrite had an effective permeability $(\mu^2 - k^2)/\mu$. This, however, is not a true permeability, as in the case of propagation along the field direction, and there is no simple scalar relation between B and H. For instance in Fig. 10, which shows the variation of the field quantities across the guide, it can be seen that although B_y and B_z

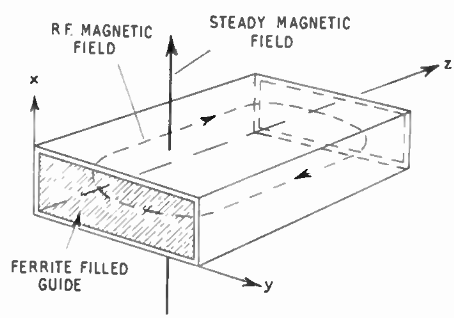

Fig. 9.—Ferrite-filled rectangular guide with transverse magnetic field.

vanish in the normal way at the edge and the centre of the guide respectively, the corresponding values of H are not zero at these points. In fact the field pattern of the two H components have undergone a transverse shift, one to the left and the other to the right. The direction of the shift depends on the sign of the magnetic field or on the direction of propagation, and is therefore a non-reciprocal effect. In practical applications the guide is only partially filled with ferrite and if this filling is unsymmetrical about the centre line, non-reciprocal effects can be achieved not only by field displacement, but also by the boundary conditions at the ferrite surface influencing the forward and backward propagation constants differently, as regards both phase and attenuation.

5.1. *Absorption Isolators in Rectangular Guide*

A resonance absorption isolator in rectangular guide is illustrated in Fig. 11. No

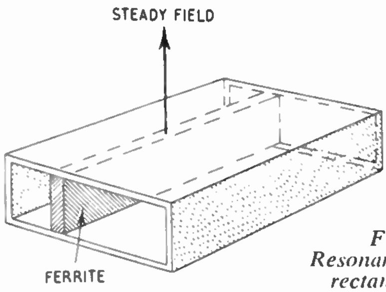

Fig. 11.—Resonance isolator in rectangular guide.

complete theory for its operation has been published, but a simple approach in terms of circular polarization can be used if it is assumed that the ferrite does not appreciably perturb the field pattern in the guide. The normal longitudinal and transverse components of H in a guide are in phase quadrature and therefore combine to produce elliptical polarization in a plane parallel to the broad face. This can be described in terms of a right-handed and left-handed circular component of magnitude A and B respectively where $A+B$ and $A-B$ are the major and minor axes of the ellipse. In Fig. 12 a typical variation of A and B across the guide is illustrated, together with a diagram of the ellipticity at various points. It should be noticed that the direction of rotation is opposite for the two directions of propagation. It has been shown previously (see Fig. 4), that an infinite ferrite shows no resonance effect for one of the senses of circular polarization. Similarly if a ferrite wafer is so placed in the guide that

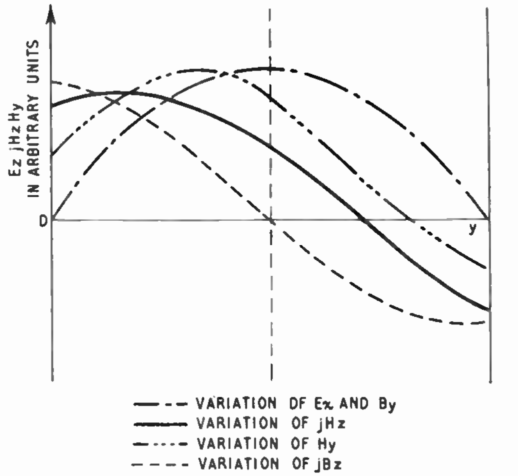

Fig. 10.—Variation of field quantities across ferrite-filled guide.

pure circular polarization of its internal magnetization is generated by the wave no resonance effects will be produced for one direction of propagation, whilst maximum effect will occur for the other. On account, however, of the strong transverse demagnetizing effects in the wafer, a certain degree of ellipticity will be required in the external r.f. field. If the demagnetizing coefficients in the y and z directions are N_y and N_z then the internal fields can be written, in terms of the external:

$$h_y = h_y' - N_y m_y/\mu_0 \quad h_z = h_z' - N_z m_z/\mu_0$$

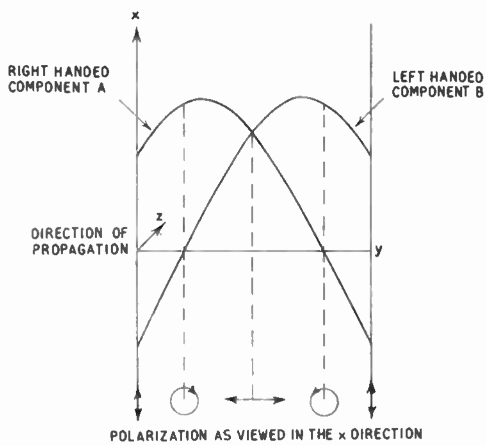

Fig. 12.—Magnitudes of circularly polarized components of magnetic field in the plane of the broad face of a typical rectangular guide.

where h_y' and h_z' are the external fields. Eq. (1) can now be written, in terms of its Cartesian components, as follows:

$$\left.\begin{aligned} j\omega m_y &= -\gamma\left[m_z\left(H_x+\frac{N_zM_x}{\mu_o}\right)-M_xh_z'\right] \\ j\omega m_z &= -\gamma\left[M_xh_y'-m_y\left(H_x+\frac{N_yM_x}{\mu_o}\right)\right] \end{aligned}\right\} \quad \ldots(11)$$

giving:

$$\left.\begin{aligned} m_y &= \frac{\gamma^2 M_x\,(H_x+N_zM_x/\mu_o)}{\omega^2-\gamma^2\left(H_x+\frac{N_zM_x}{\mu_o}\right)\left(H_x+\frac{N_yM_x}{\mu_o}\right)}h_y' - \frac{j\omega\gamma M_x}{\omega^2-\gamma^2\left(H_x+\frac{N_zM_x}{\mu_o}\right)\left(H_x+\frac{N_yM_x}{\mu_o}\right)}h_z' \\ m_z &= \frac{\gamma^2 M_x\,(H_x+N_yM_x/\mu_o)}{\omega^2-\gamma^2\left(H_x+\frac{N_zM_x}{\mu_o}\right)\left(H_x+\frac{N_yM_x}{\mu_o}\right)}h_z' + \frac{j\,\omega\,\gamma M_x}{\omega^2+\gamma^2\left(H_x+\frac{N_zM_x}{\mu_o}\right)\left(H_x+\frac{N_yM_x}{\mu_o}\right)}h_y' \end{aligned}\right\} \quad \ldots\ldots\ldots(12)$$

Resonance therefore occurs at a frequency ω_o given by:

$$\omega_o=\gamma\sqrt{\left(H_x+\frac{N_zM_x}{\mu_o}\right)\left(H_x+\frac{N_yM_x}{\mu_o}\right)}$$

In a thin wafer $N_y=1$ and $N_z=0$, and the resonance frequency becomes: $\omega_0=\sqrt{(\gamma H_xB_x/\mu_0)}$ This is also the condition for the effective permeability referred to above becoming infinite, for according to equation (6) it can be written

$$\mu_{eff}=\frac{\mu^2-k^2}{\mu}=\frac{\omega^2-\left(\frac{\gamma B}{\mu_o}\right)^2}{\omega^2-\gamma^2\,\frac{HB}{\mu_o}} \quad \ldots\ldots\ldots(13)$$

and hence the resonance condition should apply even when the ferrite is not thin. For the thin ferrite however a special case occurs when:

$$h_z'=\pm j\left(\frac{\mu_oH_x}{B_x}\right)^{\frac{1}{2}}\ h_y'$$

Eq. (12) can now be written

$$\left.\begin{aligned} m_y &= \left(\frac{\mu_oH_x}{B_x}\right)^{\frac{1}{2}}\frac{\gamma M_x}{\omega\pm\omega_o}\,h_y' \\ m_z &= \left(\frac{\mu_oH_x}{B_x}\right)^{\frac{1}{2}}\frac{\gamma M_x}{\omega\pm\omega_o}\,h_z' \end{aligned}\right\} \quad \ldots\ldots\ldots(14)$$

and it will be seen that for positive sign in the denominator there is no resonance. Thus if the ferrite is placed at the point in the guide where the ratio of the longitudinal to the transverse component of the elliptically polarized field is $(\mu_0H_x/B_x)^{\frac{1}{2}}$ the wave will only be attenuated in the reverse direction of propagation. Several workers have published experimental results on this subject,[8, 10, 15] and some curves obtained by the Bell Laboratories are shown in Fig. 13. It can be seen that there is an optimum position for the ferrite which gives minimum forward loss, although nowhere is the resonance loss completely absent, and it has been reported that there is no advantage in reducing the thickness of the ferrite below a certain size in spite of the predictions of the theory. A maximum back to front ratio of approximately 25 was obtained and Fig. 14 shows the frequency response of a typical model.

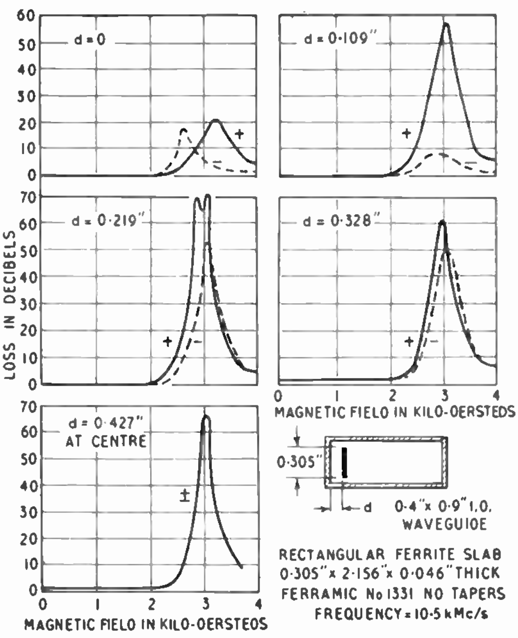

Fig. 13.—Resonance absorption for the forward (+) and reverse (−) directions of propagation as a function of the position of the ferrite element.

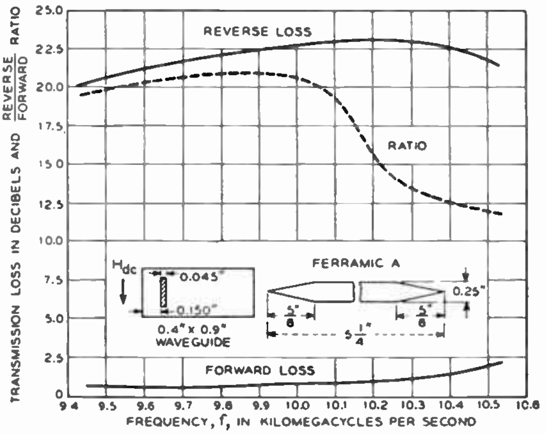

Fig. 14.—Frequency dependence for a 10-kMc/s resonance isolator.

5.2. *Non-reciprocal Phase Effects*

If the configuration of Fig. 11 is used with fields less than the resonance field it can operate as a non-reciprocal phase changer, and according to Suhl and Walker[14] a thin vane is most effective when it is positioned a quarter of the way across the guide. On account, however, of the high dielectric constant of the ferrite, even a vane which occupies only 3 per cent. of the guide distorts the field considerably, and Fig. 15a shows some experimental results[8] which demonstrate that the ferrite is most effective when it is less than a quarter of the way across the guide. These agree very well with calculations due to Lax, Button and Roth which take into account the thickness of the ferrite (Fig. 15b).[16] If the dimensions are chosen so that the difference in phase delay for the two directions of propagation is 180 deg. the device is called a gyrator.

Figure 16 illustrates how a circulator can be constructed by connecting a pair of arms from each of two four arm hybrid junctions and inserting a gyrator in one of the connections.[8, 10]

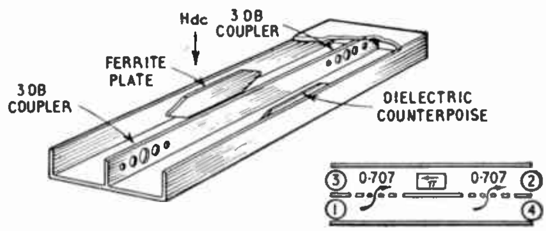

Fig. 16.—Circulator using 3-db couplers and a gyrator.

A compensating pad is inserted in the other connection so that for one direction of propagation the phase delay in the two arms is identical, but for the other there is 180 deg. difference. Thus a signal entering at arm 1 will split equally between the two guides. The relative phases of each will be unchanged when the two signals pass into the second hybrid junction and there they will combine to produce an output in arm 2 only. Similarly a signal entering at arm 3 will leave by arm 4. However signals entering at arms 2 and 4 will leave by arms 3 and 1 respectively on account of the 180 deg. differential phase change which

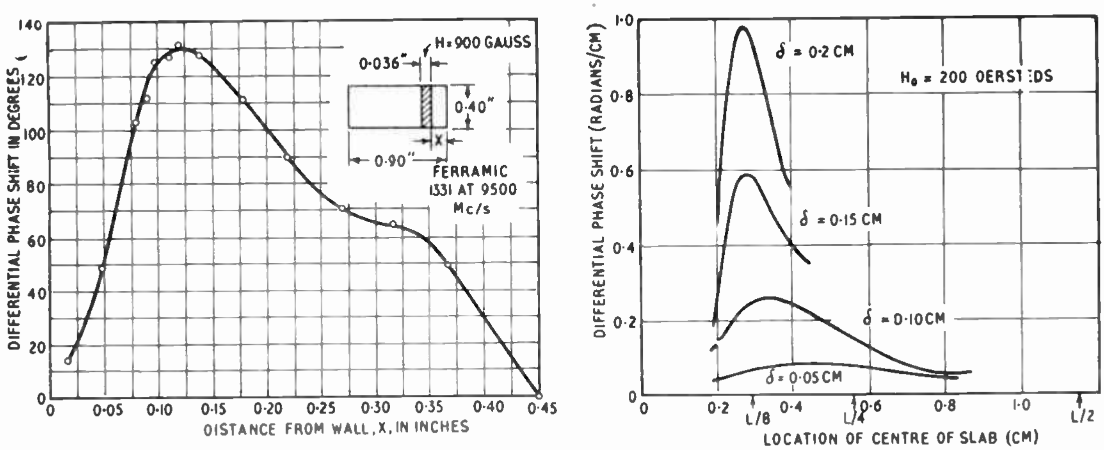

Fig. 15.—Measured and calculated non-reciprocal phase shift for ferrite loaded rectangular guide.

has taken place in the connecting arms between the hybrid junctions. This device can also be used as an isolator if arms 3 and 4 are terminated in matched loads. It will then transmit from left to right only. The non-reciprocal phase effect can also be used by putting the ferrites in the coupling region between the two guides.[8] Two guides, joined side by side as in Fig. 17 and coupled at intervals along the common wall, will only excite each other if the phase velocity is comparable in each. In the arrangement of Fig. 17

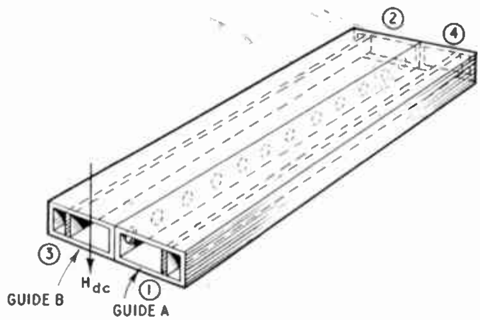

Fig. 17.—Rectangular waveguide circulator using distributed reciprocal coupling between waveguides having non-reciprocal phase constant.

the effect of the different sizes of the two guides on the phase velocity is exactly compensated, for one direction of propagation only, by positioning the ferrite in one in the opposite way to that in the other. For the reverse direction the effect of the ferrite adds to the effect of the guide size to produce a considerable difference in phase velocity. By adjusting the length of the coupler, arm 1 can be made to couple entirely to arm 2 and arm 3 to arm 4, whereas in the reverse direction the power goes straight through from arm 4 to arm 1, and from arm 2 to arm 3, and the device behaves in the same way as the circulator described above.

5.3. *Field Displacement Effects*

Isolators can also be constructed which make use of the non-reciprocal field displacement effect,[8] so that a lossy element may be placed in such a position in the guide that it is completely decoupled from the wave for one direction of propagation, but causes attenuation in the other. One way of accomplishing this is to use the ferrite in a condition where its effective permeability $(\mu^2 - k^2)/\mu$ is negative. This, according to equation 13, occurs when the induction B/μ_0 in the ferrite is greater than ω/γ.

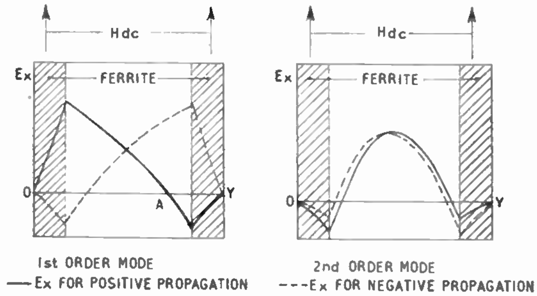

Fig. 18.—Field displacement in ferrite loaded rectangular guide.

In Fig. 18 the electric field E_x for both the first and second order modes is plotted against transverse position in a guide loaded at either side with a thin ferrite slab.[16] It will be seen that the negative permeability causes a reversal of slope of either the curve of electric field or the curve of longitudinal magnetic field at the surface of the ferrite, and enables a zero of electric field to be produced in the centre region of the guide. In the first order mode, which can propagate for all guide sizes, a considerable displacement of the zero takes place when the direction of propagation is reversed. An attenuator strip can be placed at the point A of zero field for the forward direction without appreciably effecting this wave, whereas it will cause considerable loss in the backward direction. The performance of an experimental model made for use at 6 kMc/s by the Bell Laboratories is shown in Fig. 19. At this frequency only a moderate field is required, and a good back to front ratio is obtained. At a frequency of 20 kMc/s a similar device was made to work in the region of positive permeability but its performance was not so good. Useful design equations and typical field patterns are given by Lax, Button and Roth.[16]

The field displacement effect can also be used in the design of circulators.[8] Two examples, one a three arm and one a four arm, are shown in Fig. 20. In both examples the coupling slot is not excited for one direction of propagation in the main guide, but the matching is so arranged that for the other direction the entire energy passes through it. Preliminary performance figures for the three arm circulator are given in the following table:

Path	1 → 2	2 → 3	3 → 1	2 → 1	1 → 3
Insertion loss, db	0·7	1·3	1·3	14	13·5

5.4. *Transverse Fields in Circular Guide*

The field pattern for the H_{11} mode in circular guide, illustrated in Fig. 21, is in general similar to that in rectangular guide, and thus a transverse magnetic field applied along the direction of the electric vector also has a similar effect on an element of ferrite placed in the guide. The effective permeability is still given,

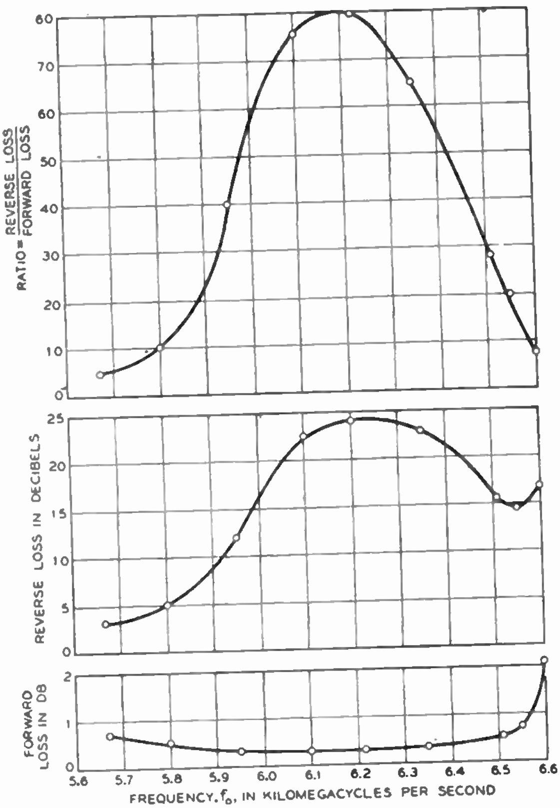

Fig. 19.—Observed performance of a 6-kMc/s resistance-sheet isolator.

though now only approximately, by the term $(\mu^2 - k^2)/\mu$, and the phase and attenuation characteristics are reciprocal only if the ferrite is on the centre line AB. However, in contrast to the rectangular case, this guide will also propagate a mode polarized at right angles whose main magnetic vector is parallel to the static field and is therefore unaffected by it. Thus the two modes propagate with a different velocity and the ferrite has become bi-refringent.[8,17] Three different ferrite configurations are shown in Fig. 22. In the first

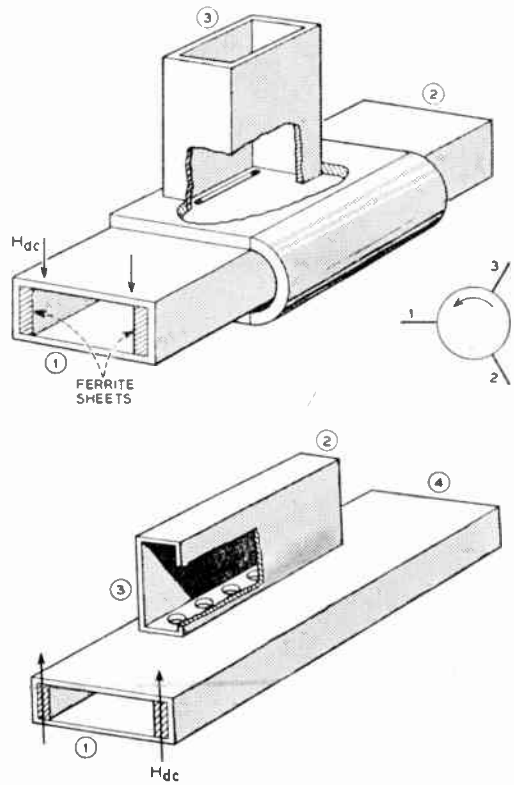

Fig. 20.—Field displacement circulators.

the ferrite is reciprocally bi-refringent and the magnetic field shown only affects the mode polarized in the y direction. In the second the same mode is still affected, but differently for the forward and backward directions of propagation. In the third both modes are

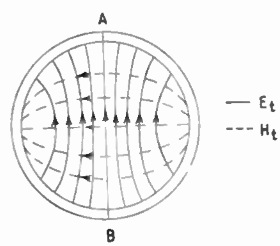

Fig. 21.—Field pattern in the H_{11} mode in circular guide.

affected and the bi-refringence for the backward wave is identical to that for the forward, but its plane is rotated through one right angle. These devices find application both for their non-reciprocal properties, and for the easy way in which the plane of the bi-refringence can be adjusted by an external magnetic field. It has been shown that a non-reciprocal bi-refringent element of the third type with differential phase delay θ deg. can be combined with two reciprocal 90 deg. bi-fringent elements oriented as shown in Fig. 23(i) to reproduce all the properties of a Faraday rotation element and give a non-reciprocal rotation of $\theta/2$.[8] In the

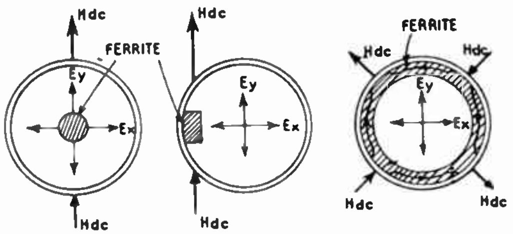

Fig. 22.—Examples of use of ferrite as a bi-refringent medium.

construction of a circulator one of the reciprocal elements can be omitted, as shown in Fig. 23(ii) for the special case where the remaining two are inclined at 45 deg. to the input and output guides respectively, and where the ferrite produces a 90 deg. differential phase shift. Vertically and horizontally polarized waves incident from the left are converted by the ferrite (a) into right-handed and left-handed circularly polarized waves respectively and the

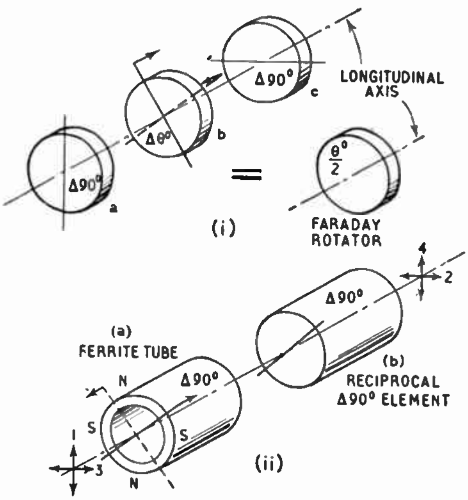

Fig. 23.—Use of bi-refringent elements to simulate Faraday rotation.

reciprocal bi refringent element (b) reconverts them into plane polarized waves of the original orientation. For the backward direction of propagation the reciprocal element performs a similar conversion to circular polarization but the ferrite, which now produces a bi-refringence in a plane at right angles, reconverts the waves into plane polarization but with the plane rotated through 90 deg. The device thus functions like a normal Faraday rotator, with the added advantage that the input and output guides can be set in line, or at any angle to one another, on account of the rotational symmetry of the circularly-polarized waves. A useful application, where the bi-refringent effect alone is required, is illustrated in Fig. 24. This device was originally designed to incorporate a rotating half wave plate located between two quarter wave plates, and gave a frequency change equal to twice the frequency of rotation. Due to mechanical considerations this could not be very large. However, if the half wave plate is replaced by a ferrite and a rotating magnetic field is applied by means of four pole pieces

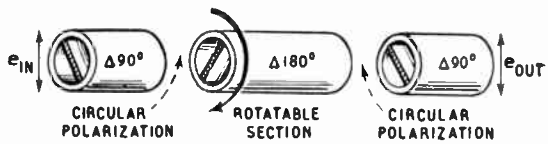

Fig. 24.—Frequency changer.

disposed at 90 degree intervals round the waveguide, frequency changes as great as 20 kilocycles can be produced.[18]

6. Comparison of Ferrite Devices

6.1. *Classification*

It has been shown that not only can several different types of non-reciprocal microwave component be constructed, but that in each case different aspects of the ferrite behaviour may be used to gain the desired result. It might be useful, therefore, to classify the components, and compare the merits of different types of construction. The three non-reciprocal waveguide elements that have been described, the gyrator, the isolator and the circulator, cannot be entirely separated from one another. Thus a gyrator can be combined with two hybrid junctions to produce a circulator, and a circulator can be transformed into an isolator by terminating one or two of its arms with matched loads. If, however, one approaches the matter by considering the means employed, one can immediately pick out the resonance isolator as being quite distinct. This can be constructed in rectangular guide, or in circular guide if means are provided for producing circular polarization, and in both cases the magnetic field must be correctly set to give resonance absorption at the frequency employed. If the field can be adjusted, these two devices, as all isolators, may be used as variable attenuators. The field displacement components also form a separate group. The field is made to couple, in one direction only, to a lossy element in the case of the isolator, and to another guide in the case of the circulator. All but one of the remaining components, which use as their basis a circulator operating on a common fundamental principle, can be included in the third group. This comprises the Faraday rotation isolator and circulator, the non-reciprocal bi-refringent plate isolator and circulator, and the isolator and circulator formed by placing a rectangular guide gyrator in one of the two arms connecting a pair of four arm hybrid junctions together. Each of these can be transformed into a variable coupler or a waveguide switch by providing an adjustable magnetic field. The common principle that operates in all of them is that each input splits up into two modes which are transmitted through the ferrite with different propagation constants. In the Faraday rotator the ferrite is bi-refrigent for the two circularly polarized modes; in the normal bi-refringent element it is the two plane polarized waves which are separated; and in the hybrid junction type the wave is split into two guides, only one of which contains the ferrite. The non-reciprocal effect is due to the fact that the difference between the relative phase of wave one to wave two in the forward and backward directions is 180 deg. Thus in the first two instances wave one is advanced by 90 deg. on wave two in the forward direction, giving 45 deg. positive rotation in the Faraday rotator, and circular polarization in the normal bi-fringent device, and in the reverse direction wave two is advanced by 90 deg. on wave one. In the third instance the phase change is equal for the two waves in the forward direction but differs by 180 deg. in the reverse direction. The remaining component, the total reflecting negative permeability isolator and circulator, must be assigned to a group of its own. It also splits the input up into two circularly polarized modes but it distinguishes between them by transmitting one and totally reflecting the other. The reflected wave has its plane of polarization rotated through 90 deg. on returning through the circular polarizing element, and can therefore be picked up by a guide coupling in at right angles.

6.2. *Comparative Effect of Frequency on Non-reciprocal Components*

Most of the magnetic properties of ferrites are dependent on frequency since they are linked with a resonant process, and it is necessary to take the frequency into account when choosing between different designs of circulator and isolator. In Fig. 25 the internal field required to produce various characteristics for a typical ferrite is plotted against frequency. Curve 1 applies to resonance in an infinite block of ferrite magnetized along the direction of propagation. Curve 3 applies to a thin rod in similar circumstances. The resonance field for a transversely magnetized thin wafer is given by curve 2. Curve 4 shows the field required in all cases for zero permeability, and in the region between this curve and the appropriate resonance curve the permeability is negative. By examining these curves one can see if the fields for a particular case are going to be too

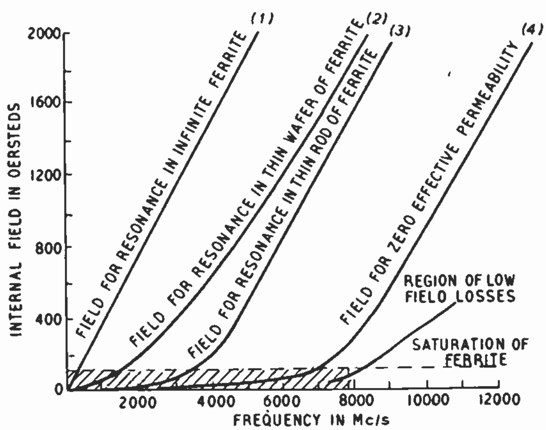

Fig. 25.—Relation between field and frequency for certain characteristics of a typical ferrite.

high or too low for practical use. The limit on the maximum is of course only determined by the magnets available, but at low frequencies especially, the minimum field must be large enough to saturate the ferrite. When the material is only weakly magnetized adjacent domains may not be parallel, and will exert a considerable demagnetizing effect on each other. No very satisfactory quantitative treatment has been produced, but it appears that the demagnetization is greater than in a thin rod and resonance can occur up to a maximum frequency of the order γM, or about 10,000 Mc/s. The corresponding region of unallowable fields is shown in Fig. 25.

Curves 2 and 3 show the fields required by resonance isolators in rectangular and circular guide respectively. At high frequencies the fields become identical, but the rectangular guide has the advantage since it is easier to apply a transverse field. The practical limit in frequency is about 20,000 Mc/s where the fields required are of the order of 5,500 oersteds. At low frequencies the circular guide becomes unworkable first on account of the low field limit being passed, but the rectangular guide component is very frequency sensitive in this region, the field being proportional to the square of the frequency. It is therefore advantageous to use a ferrite with a wide resonance peak.

Curve 4 applies to the two components which require negative permeabilities: the field displacement isolator in rectangular guide, and the total reflection isolator and circulator in circular guide. The low limit of the necessary field is given by this curve, and the high limit is determined by the point at which the resonance losses become appreciable. At high frequencies the minimum field required is less than the resonance field by about half the saturation magnetization of the ferrite, or approximately 1,500 gauss, and for a given field the operating region therefore extends about 5,000 Mc/s beyond that for the resonance isolators. At low frequencies the limiting point comes when the low field losses begin to merge with the resonance losses so that there is no region of loss-free negative permeability. Not much can be said about the broad band properties as they are mainly determined by other factors such as the match for the forward wave in the reflection isolator, and the change in the general field pattern in the guide for the field displacement devices. Experimental curves for the latter are given in Fig. 19.

The three types of isolator and circulator assigned to the third group in the previous section all rely on the difference in the effective permeability of the ferrite for the forward and backward directions of propagation. No exact theory is easily applicable to the bi-refringent element in circular guide or the gyrator in rectangular guide, but the differential phase change follows the general shape of the curve of Fig. 26, which is in fact plotted for Faraday rotation in a waveguide containing a thin pencil of ferrite. The interesting feature is that a constant and useful amount of rotation or differential phase change remains right up to the highest frequencies, whereas in the region of low frequencies, where the magnetic field is greater than that required for resonance, the rotation is proportional to the square of the frequency, and is in general small. This curve

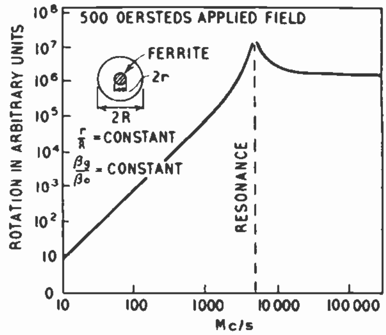

Fig. 26.—Relation between rotation and frequency in a circular waveguide containing a thin rod of ferrite, where the radius of the guide is inversely proportional to frequency and the ferrite occupies a constant proportion of it.

refers to a ferrite pencil whose radius is very small compared with a wavelength, and which occupies a constant proportion of the guide. Faraday rotators which contain larger pencils, and also the two transverse field components, show considerable frequency variation even in the higher range, on account of the high and varying r.f. field concentration in the ferrite, but they still retain the advantage of operating satisfactorily with low applied static fields. For this reason isolators of this type, and in particular the Faraday rotator, are used practically exclusively at frequencies above 30,000 Mc/s.

7. Measurement of the Components of the Permeability Tensor

It has been shown that it is theoretically possible to calculate the behaviour of a ferrite in several of its microwave applications. Simple equations describe the propagation in either infinite ferrite or in a circular waveguide loaded with a thin axial rod, for axial magnetic fields. The same is true in the case of a ferrite-filled rectangular guide with a transverse applied field, although the boundary conditions at the junction with a normal guide introduce complications on account of the transverse displacement of the magnetic field pattern in the ferrite. Numerical methods must be used to obtain solutions for partially-filled rectangular guide, and a systematic graphical approach has been devised for filled circular guide. Partially-filled circular guide, however, has not yet yielded to reasonable mathematical analysis, and this is unfortunate on account of its wide application in Faraday rotators and circulators. The loss mechanism in resonance applications has not been investigated very fully, but the static fields required to produce resonance can be calculated. The basis of all these calculations is the permeability tensor, whose components can be found theoretically in terms of the static field and magnetization of the ferrite as given in equation 6. Methods however have been developed to measure these components directly in terms of microwave quantities. They all rely on the perturbation produced by a small quantity of ferrite upon the field in a resonant cavity.[3, 19-22]

In order to calculate the change in resonance frequency let the perturbed fields and frequency be written with a subscript 1, and the original fields and frequency be written without subscript, and let the ferrite produce a magnetic moment per unit volume of M and an electric polarization of P. Maxwell's equations can then be written, for the perturbed state:

$$\text{Curl } \mathbf{E}_1 = -\mu_o j\omega_1 \mathbf{H}_1 - j\omega_1 \mathbf{M}$$
$$\text{Curl } \mathbf{H} = \varepsilon_o j\omega_1 \mathbf{E}_1 + j\omega_1 \mathbf{P} \qquad \ldots\ldots\ldots(15)$$

and, for the unperturbed state, in complex conjugates:

$$\text{Curl } \mathbf{E}^* = \mu_0 j\omega \mathbf{H}^* \quad \text{Curl } \mathbf{H}^* = -\varepsilon_0 j\omega \mathbf{E}^* \quad \ldots\ldots(16)$$

Making use of the fact that the expression:

$$(\mathbf{H}^*.\text{Curl } \mathbf{E}_1 - \mathbf{E}_1.\text{Curl } \mathbf{H}^*) + (\mathbf{H}_1.\text{Curl } \mathbf{E}^* - \mathbf{E}^*.\text{Curl } H_1)$$

when integrated over the whole volume v of a perfectly conducting cavity, can be shown to be zero,[22] then from eq. (15) and (16):

$$\frac{\omega_o - \omega}{\omega} = \frac{\int_{\Delta v} (\mathbf{M}.\mathbf{H}^* + \mathbf{P}.\mathbf{E}^*)\,dv}{\int_v (\varepsilon_0 \mathbf{E}.\mathbf{E}_1^* + \mu \mathbf{H}.\mathbf{H}_1^*)\,dv}$$

where the integral of the polarisation and the magnetic moment need only be taken over the volume Δv of the ferrite. If the ferrite is placed in a region where the electric field is zero, then $\mathbf{P}$ is also zero. For small perturbations the new fields $\mathbf{E}_1$ and $\mathbf{H}_1$ are approximately equal to the original fields $\mathbf{E}$ and $\mathbf{H}$. Noting that, as the time average of the electrical and magnetic stored energies are equal, $\varepsilon_0 \mathbf{E}.\mathbf{E}_1^* = \mu_0 \mathbf{H}.\mathbf{H}_1^*$, one obtains the following relation:

$$\frac{\delta\omega}{\omega} = \frac{-\int_{\Delta v} \mathbf{M}.\mathbf{H}^*\,dv}{2\int_v \mu_0 \mathbf{H}.\mathbf{H}^*\,dv}$$

If the fields are circularly polarized about the static field direction the effective permeability of the ferrite is $(\mu \pm k)$. For a sphere therefore the magnetic moment per unit volume in a field H is given by:

$$\mathbf{M} = \frac{\mu_o\, 3\,(\mu \pm k - 1)}{\mu \pm k - 2}\,\mathbf{H}$$

and hence the relative change in frequency can be written:

$$\frac{\delta\omega}{\omega} = -\frac{\Delta v\, H^2}{2\int_v H^2\,dv}\left(\frac{3(\mu \pm k - 1)}{\mu \pm k + 2}\right)$$

Several workers have made measurements with this method.[20, 21, 22] An H_{11} cylindrical cavity is used and a small sphere of ferrite is located on the axis close to, but not touching, the end wall. If the cavity is excited with plane polarization two resonances will be observed corresponding to the two circularly polarized components. These can be excited alone, if necessary, and studied individually. Measurements can be made both of shift of frequency and of reduction in Q and hence both the real and imaginary parts of μ and k can be found. It has been found that this method is only satisfactory for very small spheres, i.e. less than 0·5 mm diameter at 10,000 Mc/s. The frequency shift is then very small, particularly at points far from resonance. More accurate measurements can be made using either thin rods[3] or thin discs.[23] Van Trier[3] has developed apparatus for the former case and a variant of eq. (5) is used for determining the permeability components. Thus:

$$\frac{\delta\omega}{\omega} = \frac{\delta\beta_g \cdot \beta_g}{\beta_0^2} = \left[\frac{\pi U N_1'(U)}{4 J_1''(U)}\right] \frac{1}{\beta_o^2} \left(\frac{r}{R}\right)^2 \left\{ \beta_g^2 \left(\frac{\mu - k - 1}{\mu - k + 1}\right) + \beta_o^2 \left(\frac{\varepsilon - 1}{\varepsilon + 1}\right) \right\}$$

Additional relations have been derived by Gintsburg[4] for the case where the cavity is supporting an H_{01} or an E_{01} mode. By this method both μ_z and ε may be measured. Thus for the E_{01} mode:

$$\frac{\delta\omega}{\omega} = \frac{1}{\beta_o^2} \frac{\varepsilon - 1}{2} \left(\frac{r}{R^2}\right)^2 \frac{U_E^2}{J_1^2(U_E)}$$

where U_E is defined by $J_0(U_E) = 0$ (first zero);

For the H_{01} mode:

$$\frac{\delta\omega}{\omega} = \frac{1}{\beta_o^2} \frac{\mu_z - 1}{2} \left(\frac{r}{R^2}\right)^2 \frac{-U_H^2}{J_o^2(U_H)}$$

where U_H is defined by $J_1(U_H) = 0$ (second zero).

8. Acknowledgments

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DIAMOND, A PRACTICAL RADIATION COUNTER*

by

W. F. Cotty (Associate Member)†

SUMMARY

The properties which make diamond an attractive material for a practical conduction counter are discussed and the general theory of conduction counting is outlined. The counting property is thought to be a structural property and this also is responsible for the ill effects of polarization; diamonds, however, have been found which have the natural ability to maintain a steady counting rate for long periods and a method for selecting them is given. It is suggested that, for practical purposes, the diamond should be drilled with a small hole for the inner electrode and mounted at the end of a rigid or flexible cable; finally, the principles of amplification are described in detail.

1. Introduction

The desire of nuclear physicists for a faster counter than the conventional gas-filled types led to the development of the crystal counter. van Heerden[1] pioneered this work, and in 1945 reported success in getting silver chloride to count beta rays. This material, however, would only count when cooled to the temperature of liquid air and this limited its usefulness as a practical counter; consequently, a search was begun for a crystal which would count at room temperature.

In the twenty years prior to van Heerden's discovery a great deal of work had been done on the photo-electric properties of diamonds, and since the phenomena of counting and photoconductivity are very closely related, it is not surprising that scientists saw in diamond a likely material for conduction counting. Nevertheless, two years elapsed before Wooldridge, Ahearn and Burton[2] reported positive results from alpha particles and a month or two later Curtiss and Brown[3] from gamma rays. Thus, the first room-temperature crystal counter was born.

Both teams reported difficulty in finding suitable counting specimens and later another serious difficulty, polarization, arose. Indeed, at one time, polarization received such adverse publicity that it threatened to put an end to all future development of crystal counters. During the past few years, however, valuable experience has been gained in examining counting diamonds and studying their properties, and now it is possible to get diamonds which have the natural ability to maintain a steady counting rate for long periods of time. Fortunately, it is the most sensitive counting diamonds which are the least affected by polarization.

The success of the diamond as a practical counter, therefore, hinges on obtaining a really sensitive diamond for the instrument. The associated equipment is of secondary importance in this respect. This does not mean that a good counting diamond will function efficiently with badly designed equipment, but rather that the best electronic equipment available will not turn a poor counting diamond into an efficient counter.

There are, therefore, two distinct parts to a practical diamond counter, the diamond itself and its associated electronic equipment, and we shall treat each of these parts in detail.

2. Principles of the Diamond Counter

Besides its ability to count, diamond possesses other properties, chemical, physical and electrical, which make it an attractive material for conduction counting and it may serve as an informative introduction to engineers with a limited knowledge of crystals, if we were to outline these properties before we pass on to discuss more technical details of the counter.

2.1. *General Properties of Counting Diamonds*

Chemically, diamond is pure crystalline carbon, carbon in the unstable state. Such is its chemical composition that it will not combine

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U.D.C. No. 621.387.46.

with other substances at ordinary temperatures. For medical purposes, this is a great advantage because it can be sterilized and cleaned to meet the most stringent demands.

Physically, diamond is the hardest mineral known but it can be cut and polished to form the head of a pinpoint counter. It has a specific gravity of 3·5, affording it a stopping power nearly 3,000 times that of air; in other words, a 1-MeV beta particle which would travel through a metre of air would expend all its energy in a diamond 1 mm thick.

Diamond will count at room temperature without any special preparation as distinct from some crystals which will only count when cooled to the temperature of liquid air, and others which have to be annealed. This enables us to mount the diamond in the simplest of holders, which attached to a rigid or flexible lead forms a very satisfactory probe. Diamonds as small as 1 mm side have been used in probes, making them an ideal size for pinpoint counters.

When we analyse its electronic properties we find that the diamond, like most counters, has its advantages as well as its limitations.

It has an extremely fast resolving time, 10^{-8} seconds, and is linear in its output. The energy needed to free a secondary electron in a diamond is about 10 electron-volts which is much less than the 30 eV required to produce an ion pair in a gas. The diamond would therefore seem to be more sensitive than a gas counter but there is no practical evidence of this in the literature. A good counting diamond will, however, count alpha and beta particles of high energy as efficiently as a Geiger-Müller counter.[4]

Under present conditions only a very small percentage of diamonds make really good counters and even among these there is a wide variation in performance. This is due to the electronic traps in the crystal which create an internal space charge. This polarization, as it is called, severely limits the usefulness of many diamond counters but recent research[5] has disclosed diamonds which are not unduly affected by polarization and will function as counters without the cumbersome auxiliary equipment usually needed to counteract the ill effects of polarization.

These were some of the properties which were originally considered promising enough to warrant the development of the diamond as a pinpoint counter. Medically, there was yet another reason; the gamma ray absorption of diamond (carbon) is of the same order as that of human tissue. Thus radiations will penetrate both materials to about the same degree giving, theoretically, a more accurate indication of the dose rate being received by a patient than is possible with other counters.

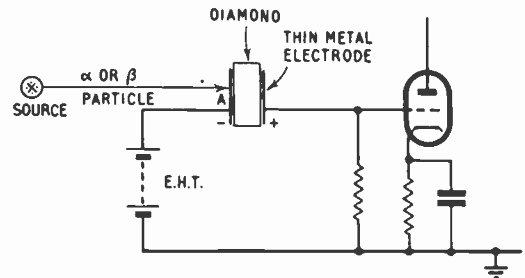

Fig. 1.—Action of a high speed particle on a diamond.

2.2. *General Principles*

From the experiments of Wooldridge, Ahearn and Burton we know that a counting diamond working in an electric field will generate small pulses of current when bombarded by high-speed particles from a radioactive source, and it may be helpful at this stage if we turn our attention to the theory underlying this phenomenon.

When an insulator is bombarded by high-speed particles, according to solid state theory the electrons freed by the radiations are raised to the conduction band. Under the influence of an applied field these electrons, and the positive holes in the normally filled band, move just as they do in the case of photo-conductivity. With a sufficiently high field across the crystal, relatively free of electron traps, it is possible under favourable conditions to detect the movement of these charges and observe the conductivity pulses.

Considering Fig. 1, a high-speed particle, alpha or beta, enters the crystals at A. The penetrating particle liberates secondary electrons in the body of the crystal until it is stopped. The process whereby these electrons are produced along the track of the particle is analogous to ionization in a gas where positive and negative ions are produced. In a crystal a positive hole is produced every time an electron is set free; thus there are an equal number of electrons and positive holes produced each time a high-speed particle enters the crystal. In a crystal like diamond with its symmetrical lattice structure both electrons and holes move with substantially the same velocity. In the diagram the crystal is drawn to represent a thin slice of diamond with thin metal electrodes evaporated on the opposite faces. A high potential is applied to these electrodes and when the diamond is under bombardment the liberated current carriers are attracted to them. The electrons flow towards the anode and the positive holes towards the cathode. This movement of carriers is recorded as a voltage pulse on the grid of the valve.

When a high-speed particle is absorbed by an atom an ion pair is created, and in the process the electron is said to be raised to the conduction band. To explain the mechanism underlying this process it is convenient to refer to the energy-band diagram shown in Fig. 2. In this diagram, A represents the outermost completely filled energy band associated with the insulating crystal, and B is the next higher band which is normally completely empty. As there are normally no vacant levels in A, electronic conduction cannot take place. When an electron is moved from A it leaves an empty space C in this band. This is called a positive hole. It is possible for an electron in a neighbouring level

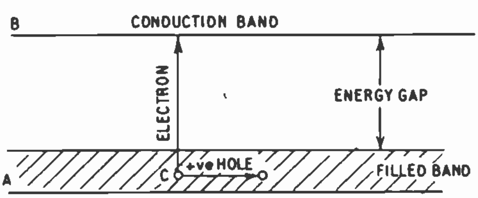

Fig. 2.—The electronic energy bands of an insulator.

in the band A to occupy the vacant site thus leaving a new hole in its previous level. By continued repetitions of this process, the positive hole can be considered as migrating through the crystal lattice. In diamond this effect is known to take place. When the electron which was ejected from C reaches the conduction band it is completely free to wander through the crystal lattice.

In the absence of an electric field both the electrons and their oppositely charged numbers, the positive holes, move in the crystal with random thermal energy. When an electric field is placed across the crystal it has the effect of accelerating both electrons and holes, which move in opposite directions. Acceleration of the current carriers by the field results in a development of a net drift velocity in the direction of the force. The mean drift velocity y developed in unit electric field E is called the mobility v.

Thus $$y = vE \qquad \text{...........(1)}$$

With an applied field across an ideal crystal all the current carriers would reach the terminals. No crystal, however, is perfect and the progress of the current carriers through the crystal is greatly hampered by repeated collisions with the crystal's imperfections. When referring to the crystal's imperfections it is customary to think of these as defects in the lattice, impurity atoms or other defects in the structure of the crystal as distinct from ordinary cracks. Any of these defects may trap an electron. A fuller description of the nature of crystal imperfections and the important role they play in crystal counting will be given later when discussing polarization. If the applied field is weak, the electrons in their flight across the crystal to the anode collide with the ions of the lattice and lose some of their energy. Each electron will try again and repeat the process until it reaches the anode or is captured by a trap in the crystal and becomes immobilized. By increasing the strength of the field sufficiently the velocity of the current carriers is increased and, in all probability, they will move past the traps too quickly to be captured and eventually reach the electrodes.

If we denote the time an electron spends in the conduction band by T we are able to calculate the average distance w travelled by the freed electrons from the expression

$$w = vET \qquad \text{...........(2)}$$

where v is the mobility and E the applied field. The distance w is known as the mean range of the secondary electrons in the field. The mean range w is also known as "Schubweg."

In deriving this formula we assume that (i) the average time T an electron spends in the conduction band is independent of the field strength, (ii) the trap distribution in the crystal is homogeneous, and (iii) the crystal itself is free from space charge effects. It should also be noted that the formula holds good only for traps in the crystal and does not allow for open cracks.

Hecht's[6] work on photoconductivity enables us to determine how the current in the measuring circuit varies with the field applied to the crystal. These theoretical considerations have been set out fully by Mott and Gurney[7] and later summarized by Hofstadter[8] so only the relevant formulae need be stated here. According to Mott and Gurney, if an electron is released a distance X from the anode, the mean distance x drifted by an electron is

$$x = w(1 - e^{-X/w}) \qquad \ldots\ldots\ldots\ldots(3)$$

where $w = vET$ as in the previous equation.

Hecht has shown that the charge measured externally is equal to the charge released in the crystal multiplied by x/d where d is the total distance between the crystal electrodes. If N is the number of electrons released and e the electronic charge, then the ratio φ between the charge released, Ne, and the charge observed, Nxe/d, is

$$\varphi = \frac{w}{d}(1 - e^{-X/w}) \qquad \ldots\ldots\ldots\ldots(4)$$

This formula is for electrons. In diamond the holes move and contribute a like amount to the current. In the case of alpha bombardment the particle only penetrates the crystal to a depth of about 10 microns so the amount the holes contribute is negligible if the ray enters through the cathode. For simplicity, we will consider only alpha radiation in the examples below.

When an alpha particle enters the crystal near the cathode, $X = d$ and we may write

$$\varphi = \frac{w}{d}(1 - e^{-d/w}) \qquad \ldots\ldots\ldots\ldots(5)$$

A charge pulse Q due to an alpha particle, at saturation value of field strength, would be

$$Q = \frac{New}{d}(1 - e^{-d/w}), \qquad \ldots\ldots\ldots\ldots(6)$$

where N is the number of electrons released and e the electronic charge ($1{\cdot}6 \times 10^{-19}$ coulombs) as previously; d is in centimetres.

The voltage V induced on the grid of the preamplifying valve would therefore be

$$V = \frac{New}{Cd}(1 - e^{-d/w}) \qquad \ldots\ldots\ldots\ldots(7)$$

where C is the total capacitance of the crystal, holder and input circuit.

The above formulae hold good only for particles that enter the crystal near the cathode and expend all their energy in the crystal. With gamma rays this is not so and the calculations become more complex. In diamond the major number of secondary electrons produced by gamma rays is due to Compton effect and the following equation applies

$$V = \frac{New}{Cd}\left\{1 - \frac{w}{d}(1 - e^{-d/w})\right\} \ldots\ldots\ldots(8)$$

where all the symbols have the same meanings as before.

Having derived a number of equations theoretically, we will now show how they can be applied to practical measurements.

2.3. *Diamond Counter Measurements*

Diamond counter measurements may be expressed by drawing a graph, known as a Hecht curve and shown in Fig. 3, either from the equations already derived or experimentally.

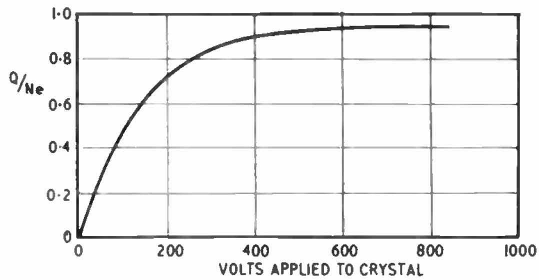

Fig. 3.—Hecht's curve. The plot of Q/Ne against V, where $V = Ed$. Q/Ne may be plotted against w/d instead of V since $w = vET$.

It should be noted that it is characteristic of all counting crystals to produce pulses of various amplitudes although the source be a homogeneous one. To avoid any confusion which may arise from this peculiarity, only pulses of maximum amplitude are selected for measurements because they alone give a reading which is truly proportional to the energy of the source.

When we draw a graph from experimental results we make use of the fact that when an alpha particle enters the crystal at the cathode, the charge pulse is a maximum when all the secondary electrons reach the anode (eqn. 6). Alpha particles from a polonium source are recommended for this experiment because they are mono-energetic and the exact energy of the particle entering the crystal may be calculated from the well-known Bragg curve. Furthermore, alpha particles only penetrate the crystal to a depth of about 10 microns, and as they enter the crystal near the cathode the movement of the positive holes may be neglected because they contribute practically nothing to the total charge. All the electrons that reach the anode will then have traversed the entire distance between the electrodes. In all experiments of this kind the most sensitive part of the counting diamond should be selected for bombardment. More accurate results will be obtained by using this method than by scanning a large portion of the crystal and taking an average of the resulting current. One assumes in the latter instance a homogeneous distribution of traps but this is rarely the case in practice.

In Fig. 3 the graph of the readings of the pulse heights obtained with a change in field strength is shown. V is the actual voltage across the crystal and is equal to Ed where E is in volts per centimetre and d is the thickness of the crystal in centimetres. In the graph only the maximum signal pulses are used. These are selected by means of a discriminator, the function of which will be described later.

By inducing a known voltage on the grid of the preamplifying valve to correspond in magnitude to the size of the signal pulse, it is possible knowing the capacitance of the circuit to calculate the charge Q. It is usually more convenient to inject a pulse from a pulse generator, calibrated directly in ion pairs, to obtain this value.

The Hecht curve is valuable for finding the most suitable value of e.h.t. to apply to the crystal. Taken in conjunction with the theoretical formulae, the Hecht curve also enables us to calculate the important measurements of mobility, rise time, energy per ion pair and proportionality. We shall consider each of these in turn. In all these measurements the diamond is assumed to be free from space charge effects and to have a homogeneous trap distribution.

2.4. *Mobility*

From equation (2) we know that the average distance travelled by a cluster of electrons is $w = vET$. If we obtain the number of secondary electrons released Q from the Hecht curve we are able to calculate the value of vT. We measure Q as a function of E/d to find vT (eqn. 6). All symbols have the same meanings as in previous equations.

Values for vT of the order of $3{\cdot}5 \times 10^{-5}$ cm^2/volt have been found for electrons[9], and for holes $3{\cdot}8 \times 10^{-5}$ cm^2/volt which is equivalent to a mobility of about 4,000 cm^2/sec-volt. This is considerably higher than the theoretical figure[10] of 156 cm^2/sec-volt.

2.5. *Rise Time Measurements*

When electrons traverse the full length of the crystal the rise time r of the pulse is

$$r = d/vE \qquad \text{...........(9)}$$

where d is the thickness of the crystal, v is the mobility and E the field strength. It is assumed in this equation that the time T an electron spends in the conduction band is long compared with the rise time r. The value of T has been determined for diamond by Pearlstein and Sutton[9] to be of the order of 10^{-8} sec, which is a very short time; consequently T is not large compared with r so we need a more accurate expression than that given in eqn. (9).

In general, r will be the ratio of the average distance moved by a cluster of electrons in a single pulse to the drift velocity of the cluster. From eqn. (3) we know that the mean distance drifted by an electron is

$$x = w\,(1 - e^{-d/w})$$

then

$$r = x/vE = T\,(1 - e^{-d/w}) \qquad \text{........(10)}$$

As T is very small this reduces to $r = T$.

2.6. *Trapping Sites*

Let T be the time an electron spends in the conduction band and let s be the trapping cross-section. If there are N traps present per cubic centimetre of the crystal, an electron covering a distance of 1 cm has a chance of being trapped equal to sN cm^{-1}. With a thermal velocity of u cm/sec, the distance covered in T seconds is uT cm. In this time the electron is trapped. So

$$sNuT = 1 \qquad \text{...........(11)}$$

This implies that the trapping is thought to limit T.

McKay[11] gives a figure of between 2×10^{-16}/cm^3 and 2×10^{-17}/cm^3 for the trap density. This corresponds to concentrations of one to ten traps per 10^7 atoms.

The traps lie at different electrical depths which Newton[12] estimates to be between 0·25 eV and 0·75 eV.

2.7. *Energy per Ion Pair*

Referring to the Hecht curve again (Fig. 3), it is possible to calculate the energy necessary to create an ion pair if one knows the energy of the incident particle at the moment of entering the crystal and the absolute number of electrons liberated by the particle.

The Hecht curve will enable us to find the number of electrons liberated—in other words, the number of ion pairs created—and we can then determine the experimentally measurable quantity z for the energy per ion pair.

$$\text{Thus} \quad z = H/N \qquad \text{...........(12)}$$

where H is the energy of the penetrating particle and N the number of ion pairs. The quantity z is given in electron-volts.

The energy gap between the conduction band and the highest occupied band in diamond has been computed theoretically by Kimball[13] to be about 7 eV. Most experimental results appear to be of the same order of magnitude and generally lie between 7 eV and 10 eV.

The significance of these results is that the energy needed to produce an ion pair in diamond is considerably lower than the 30 eV required to produce an ion pair in a gas.

2.8. *Proportionality of Pulses*

It may be deduced from eqn. (12) above that N is proportional to the energy expended by an ionizing particle in the crystal. In principle, therefore, the diamond conduction counter is linear or, to put it another way, the pulse height is proportional to the energy of the bombarding particle.

From the foregoing formulae we are able to calculate all the preliminary data necessary for us to proceed with the design of our instrument. The amount of energizing voltage required to make the crystal count efficiently is our first consideration, but if we use a diamond having a diameter of one or two millimetres, as is usually the case in practice, we find we have sufficient voltage available in a normal h.t. supply. Knowing the energy per ion pair and the characteristics of the amplifier, we can calculate the minimum energy an incident particle would have to possess at the time of entering the crystal to be detectable above the amplifier noise and the impedance of the circuit. The diamond we would choose would be the one with the lowest number of trapping sites, that is, the most sensitive.

3. Counting Diamonds

3.1. *Counting Property vs. Structure*

From earlier work on photoconductivity it was known that those diamonds which did not fluoresce under an ultra-violet lamp fitted with a Wood's filter had the best photo-electric properties and in 1934 Robertson, Fox and Martin,[14] while examining a parcel of non-fluorescent diamonds, discovered that the best photo-conductors were those which had a laminated structure and in which the absorption band, characteristic of most diamonds, at 8 microns was missing. In their nomenclature they classified these rare specimens as Type II diamonds as opposed to Type I, the common variety. Other differences noted were that the cut-off frequency in the ultra-violet region of the spectrum for Type I was 3,000 A.U. and Type II 2,250 A.U.

Although van Heerden knew these facts he was unsuccessful with a diamond in his early experiments and it was left to Wooldridge, Ahearn and Burton to find the first counting diamond and, incidentally, the first room-temperature crystal counter. Both this team and Curtiss and Brown complained about the difficulty they experienced in finding suitable counting diamonds, but the task was made easier when a definite correlation between counting diamonds and Type II's was established by Friedman, Birks and Gauvin.[15] Diamonds having a cut-off in the ultra-violet below 2,660 A.U. were reported to make the best gamma counters.

Champion[16] confirmed this and showed that there was an 80 per cent. probability that any small transparent diamond of good optical quality, whose absorption lay below 2,650 A.U. would be, at least, a moderately good counter, whereas if the absorption lay above 2,650 A.U. the chance that it would count appreciably was less than 10 per cent.

Continuing his investigations, Champion[17] reported further that counting diamonds were composed of layers of highly perfect crystalline material separated by relatively few, and much thinner, partial barriers of imperfect material. If the average thickness of the crystallite layers is about 10 microns then the specimen has the texture necessary for a good counter. The poorer counting shown by some specimens of still greater crystallites he attributed to the presence of wider barrier layers.

3.2. *Selection of Counters*

By definition a Type II diamond is one which has a cut-off in the ultra-violet of about 2,250 A.U. Since there are only one or two of these diamonds in every thousand, a less tedious method was required to find counting diamonds than the measuring of the cut-off frequency of each individual stone. When large parcels of diamonds have to be tested for the counting property, we have found that the quickest way is to sort them by hand under an ultra-violet lamp fitted with a Wood's filter and reject those which fluoresce. In this way, with practice, we were able to discard up to 90 per cent. of the parcel. The non-fluorescent diamonds were then tested singly for the counting property in an amplifier to be described. On checking the cut-off frequencies of our best gamma counters we found that they varied considerably, ranging from 2,250 A.U. to 2,900 A.U.[5] More curiously still a number of perfect Type II specimens failed to count at all. Probably Champion's findings about the different barrier layers in the crystal provide the reason for this.

Our own results indicate that there are fewer than 1 diamond in 1,000 which can be classified as excellent counters. The author defines an excellent counter as one which will count alpha, beta, gamma and x-rays equally well and besides this—and this is the important point—will function perfectly as a counter without any outside assistance being necessary to maintain a steady counting rate. This point is stressed because all workers in this field, especially those with a limited number of diamonds at their disposal, will sooner or later be faced with the problem of polarization.

3.3. *Polarization*

Although diamonds have been grouped into Types I and II, it is seldom that we get perfect specimens of either type, and crystals which are a mixture of both types are far more common. Similarly, the amount of counting material and its position in the diamond vary with every specimen, and so does the number of impurities present. When we speak of impurities in this sense we infer lattice defects like vacant sites and interstitial atoms, as well as chemical impurities, and so distinguish them from the ordinary cracks in the crystal.

These imperfections act as trapping sites for the electrons and positive holes. A space charge consequently builds up inside the counting crystal and opposes the externally-applied field, causing a reduction in the size and number of pulses. What actually happens is that the electron drifts into the region near the imperfection, exchanges energy with the surrounding atoms and finds that it does not have enough energy left to leave the region because of the peculiar electrical disturbance set up by the imperfection. The carrier has fallen to a level of lower energy and is trapped, and the longer it is free to wander through the crystal the more likely it is to be drawn into a trap.[11] To avoid this, the energizing field across the diamond must be kept at a high level, for when the applied field is weak carriers take longer to traverse the crystal, trapping becomes more effective, and conditions become very favourable for the formation of the space charge.

It is interesting to display the amplified pulses from a counting diamond on a cathode-ray tube screen and to watch the change which takes place as the crystal polarizes. At first, one gets a pulse pattern of varying amplitudes familiar to all nuclear physicists, but gradually as the space charge builds up, the efficiency of the counter diminishes and eventually the counting rate drops and may even cease altogether. Should the energizing voltage to the diamond be disconnected at any time during the experiment, the pulses will be seen to reverse their polarity showing that the current carriers are trying to return to their original sites.

Polarization is not peculiar only to radiation counting but was known to the early workers in photo-conductivity. They overcame its ill effects by providing a complementary wavelength to the one which polarized the diamonds. Treating counting diamonds in a similar manner with light of proper wavelength has been equally successful. Although this procedure is the most common, some investigators have neutralized the space charge by electronic means, while heat treatment has also proved helpful. The effectiveness of any method will naturally depend on its ability to reach the offending trapping sites.

The depth of most trapping sites lies between 0·25 eV and 0·75 eV (eqn. 11) and by making proper use of red or infra-red light, electrons may be freed from them. Once free they are at liberty to combine with the positive holes and, if the conditions of wavelength and light intensity are correct, the crystal will maintain a steady counting rate for an indefinite period.

The conditions will be satisfied if the light from a 60-watt projection lamp is passed through a 6,900 A.U. filter. As the trapping sites are situated at different energy levels, it is reasonable to expect the electrons trapped in shallow traps to be liberated by a light of longer wavelength than is required to release those which have fallen into deep seated trapping sites. The experiments of Freeman and van der Velden[18] have actually shown that light of insufficient intensity or too short a wavelength will only restore the counting rate temporarily.

Although polarization is fundamental to all crystal counters the better diamond counters are not completely paralysed by the internal space charge. In a series of experiments Trott[19] found that polarization could be overcome by "activating" diamonds with beta particles. In this process the energizing voltage to the diamond was maintained and the beta flux continued all the time the experiment was in progress. At first there was the usual falling off in the counting rate but after about half an hour, in which time the crystal had presumably become activated, the diamond commenced counting again and rapidly reached a high level which it maintained indefinitely.

The author has always believed that to be a practical success the diamond counter would have to be kept as simple as possible and for this reason we have concentrated on searching for diamonds which would count satisfactorily without needing the help of any special anti-polarizing devices. In the course of our research, we have so far examined over 100,000 diamonds for the counting property and we have found that diamonds do exist which have the natural ability to maintain a steady counting rate for several hours at a time, perhaps indefinitely. Non-polarizing diamonds like these are naturally a great asset for practical work and we have used nothing else for the last five or six years.

4. The Diamond Probe

4.1. *Method of Mounting*

When doing experiments which entail making an investigation of the interior of a counting diamond in the laboratory, the crystal is usually mounted rigidly and as close as possible to the grid of the first valve. On the other hand, when the diamond is to be used as a practical counting device it must, of necessity, be placed some distance from the valve and the probe designed to be movable. The length the probe may be, however, is limited by the amount of extra capacitance it brings to the circuit (since $V=Q/C$) and this calls for careful design in order to get the maximum amount of signal voltage to the first valve for amplification.

Besides the purely engineering problems involved some difficulty was encountered with the setting of the diamond at the end of the probe. This has now been overcome by mounting the diamond in an entirely new way.

A very simple method of mounting, that used originally by Curtiss and Brown, is shown in Fig. 4. The diamond is clamped between two metal electrodes, one on either side of the crystal; in this way the voltage is applied across the diamond. In some cases, it may be more convenient to have the electrodes closer together and to apply the voltage to one particular face of the crystal. In either case, to improve the diamond–electrode contact it is usual to paint the faces of the diamond with aquadag or to sputter on some silver or gold.

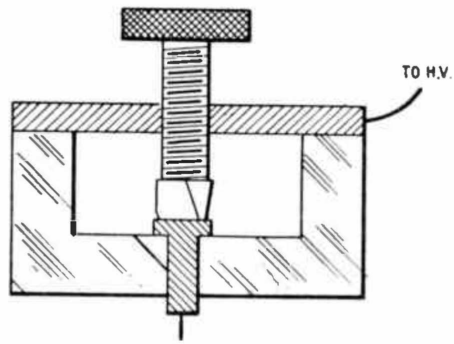

Fig. 4.—The original diamond holder of Curtiss and Brown.

For the practical applications of the counter that we had in mind, it was found better to mount the diamond at the end of a short probe. We have obtained our best results in this work by having the diamond drilled to take the inner wire of a coaxial cable and using the outer screen as a claw mounting. The hole is drilled about halfway through the diamond. The electrodes are consequently brought closer together and thus allow the applied voltage to be reduced by half. In order to prevent spurious pulses due to inadequate screening and photo-electric effect, a thin film of aquadag is painted on to the exposed parts of the diamond; this ensures that the diamond is perfectly screened electrostatically. This mounting is most desirable when the external diameter of the probe is strictly limited as would be the case if a diamond were to be mounted, for instance, inside a surgical needle (Fig. 5).

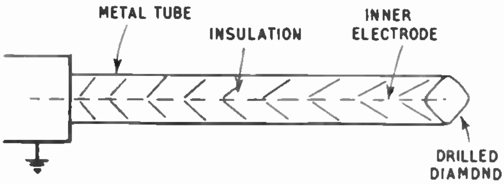

Fig. 5.—Drilled diamond mounted at the end of a rigid tube to form a probe.

4.2. *Design of the Input Circuit*

In designing a practical circuit for the probe, the diamond is coupled to the grid through a small capacitor C1 (Fig. 11). It should be remembered that the farther away from the valve we move the diamond the more capacitance we will add to the circuit and this will result in a reduction of the naturally weak diamond signals. It is estimated that the total input capacity is of the order of 30 pF, of which the diamond itself has contributed less than one-tenth. It should be remembered that, besides the actual capacitance of the holder itself, its very presence in the vicinity of the grid adds capacitance to the circuit.

5. The Design of the Associated Electrical Equipment

The pulses generated by a counting diamond are of the order of microvolts, so it is necessary to amplify them considerably before they are strong enough to operate an output recording device. Furthermore, in many cases it is not only necessary to amplify the signals but to reproduce them faithfully.

A block diagram of the equipment is shown in Fig. 6. The diamond pulses are amplified by the preamplifier before being transmitted through the coaxial cable to the main amplifier. On leaving the amplifier the signals are passed through a discriminator circuit and then on to the recording instruments.

The ionization chamber is the conventional counter most like the diamond counter in its mode of operation, indeed the diamond counter may be considered as a solid ion chamber. Therefore, for design purposes we may consider the electronic equipment associated with the ion chamber as equally applicable to the diamond counter. The most suitable amplifier for reproducing fast pulses faithfully is the linear pulse amplifier.

In the circuit to be described the main amplifier consists of two rings of three valves each, and a separate preamplifier. The principles of the discriminator circuit, whose function is to allow pulses above a predetermined level to pass and to suppress all others, will also be discussed.

5.1. *Gain of Amplifier*

The input voltage and hence the gain required to produce a given amplified output can be calculated if the total number of ion pairs collected per incident particle and the effective input capacitance are known.

Thus

$$V = \frac{ne}{C}$$

Where n equals the number of ion pairs created

e is the electronic charge

and C equals the total input capacitance.

The output voltage required to operate most discriminator circuits is between 5 and 100 volts and within these limits choice of amplitude is a matter of convenience.

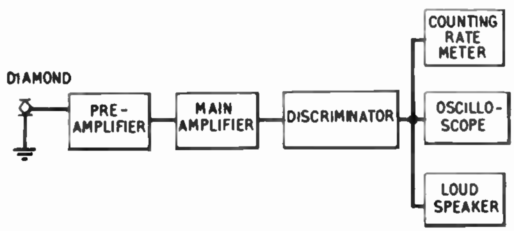

Fig. 6.—Block diagram of a diamond counter.

We know from eqn. (12) that a 1 MeV beta particle is capable of producing 10^5 ion pairs in diamonds, but this ideal would only be reached if the crystal were perfect. The total input capacitance is probably about 30 pF and the smallest detectable signal about 25 microvolts, so the amplifying circuits are designed for an overall amplification of the order of 10^6.

The gain that may be obtained from a linear amplifier is limited by the inherent noise of the input circuit of the first valve. When the gain control is set very high it is found that the output voltage of the amplifier is not zero but varies within limits in a purely random manner, in spite of there being no input voltage on the grid. This random variation arises primarily from the noise voltages associated with the components and valve of the first stage of the preamplifier and will be dealt with in detail when we discuss the design of that stage.

To achieve the results we desire, the amplifier must be stable as well as linear and in the next sub-section we will show how by using negative feedback in the design, the overall gain of the main amplifier is kept within limits of 10,000 to 20,000. The gain of the preamplifier is kept between 50 and 100.

5.2. *Stability*

The two causes most likely to affect the stability of the amplifier are, firstly, a variation in the power supply voltages and, secondly, variation in components and valves due to temperature and time. Regulated power supplies are used for the h.t. and to overcome variation in the components negative feedback is employed; the more feedback we use the more stable will the amplifier be.

A stabilized power supply with a regulation to within 0·5 per cent. for the h.t. voltages is therefore the first requirement. It is worth noting that a stabilized power supply accomplishes more than simply furnishing a direct voltage, it also affords a low source impedance, thus considerably reducing the interaction between the various parts of the circuit for which it furnishes power, while the stabilizing circuit itself acts as an excellent filter for ripple voltages.[20]

It may be a further advantage for the heaters of the valves in the early stages of the amplifier to be supplied from a direct current source instead of an alternating current one.

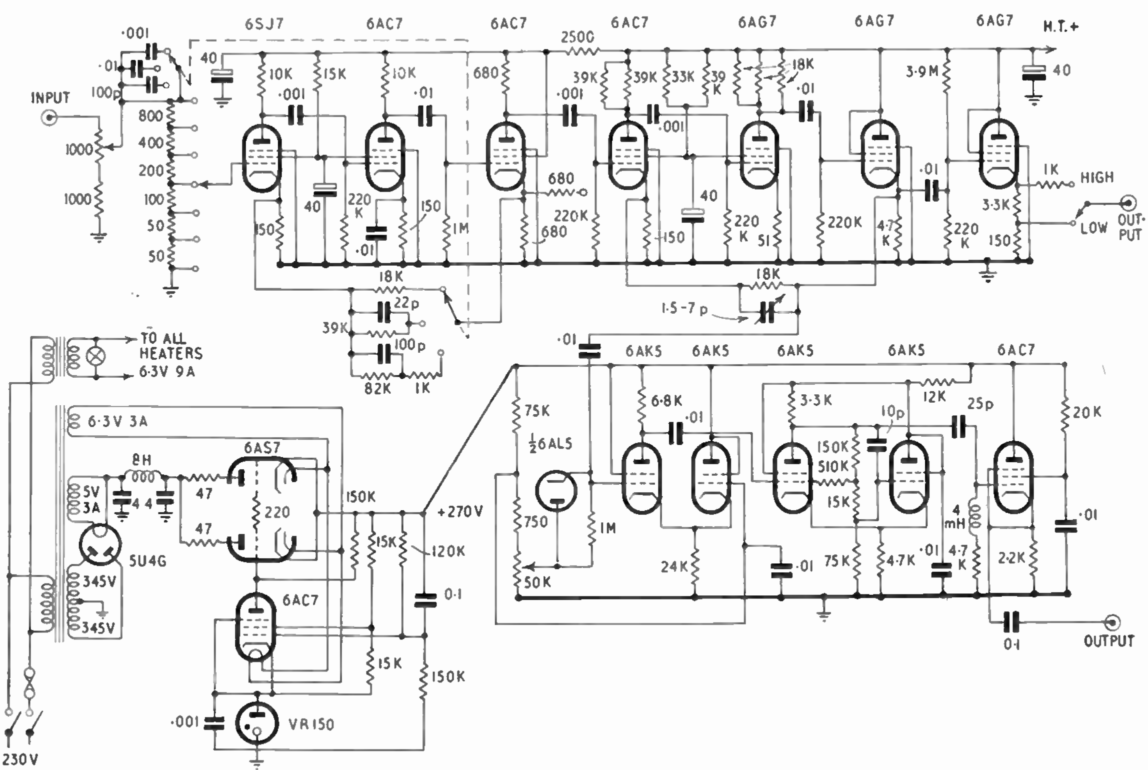

Fig. 7.—The Jordan and Bell linear amplifier.

We now turn our attention to another important factor in achieving stability, namely, negative feedback.

Negative feedback is a process whereby a fraction of the amplifier's output voltage is fed back to the input circuit in such a way as to oppose the input voltage. The two primary advantages of using negative feedback are to

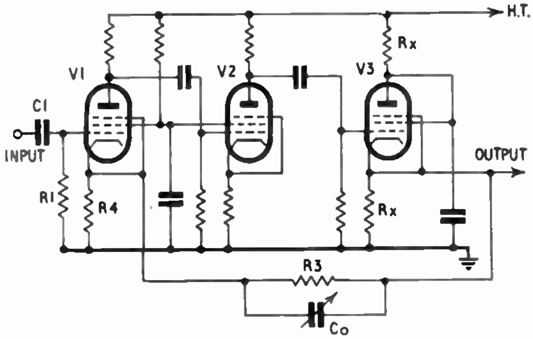

Fig. 8.—Ring of three, showing negative feedback loop from V3 to V1.

improve the stability of the amplifier and to overcome to a large extent the non-linearity between the input and output voltages due to the inherent curvature of the valve characteristics.

If g_m is the mutual conductance of a valve and R its anode resistance it can be shown that, if we increase the anode resistance to R_1 and feed back the extra gain so realized by returning a fraction B of the output voltage to the input, the amplifier gain becomes

$$\frac{g_m R_1}{1 + B g_m R_1}$$

which reduces to $1/B$ if $g_m R_1$ is very much greater than unity.

In the circuit in Fig. 8 the first two valves are the amplifiers, while V3 functions as a cathode follower and does not contribute to the gain of the stage. V3 isolates the feedback network from V2. Furthermore, it is more convenient to take the amplifier output from the cathode of V3 than from the anode of V2 because that would impose additional loading on V2 anode with consequent deterioration in the overall bandwidth. Another advantage may be gained by placing a resistor of the same value as the cathode resistor Rx in the anode circuit of V3, thereby making it possible to take an output from V3 anode of approximately the same magnitude as that on the cathode but of inverse phase.

The gain of the circuit depends on the amount of feedback from the cathode of V3 to the cathode of V1 and this is usually adjusted to be between 20 and 100. We have seen that the gain is determined by $1/B$ and for this type of circuit the gain may be accurately determined by the ratio (R3 + R4)/R4 in Fig. 8.

The capacitor C_0 is provided for adjustment of the circuit to give the best rise time with the smallest amount of overshoot.

The main amplifier is made up of two such rings of three and is shown in Fig. 7.[21]

5.3. *Bandwidth*

In a pulse amplifier it is customary to speak of the rise and decay times of the pulses instead of frequency response. This arises from the fact that the duration of the pulses may vary within wide limits and the amplifier circuits must be wide enough to pass frequencies from 30 c/s to 5 Mc/s. In practice, the time constants are made adjustable from 0·1 to 50 microseconds.

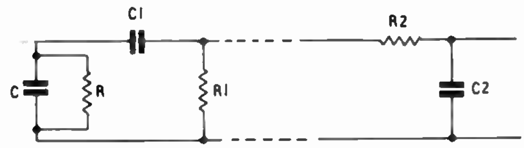

Fig. 9.—Bandwidth time constants.

Differentiating time constant is $C_1R_1 = T_1$
Integrating time constant is $C_2R_2 = T_2$
For best signal to noise ratio make $T_1 = T_2$

The frequency is made adjustable to enable the best signal-to-noise ratio to be obtained in conjunction with the resolution of pulses. If the amplifier suppresses the higher frequencies the pulse edges will be distorted, whereas suppression of the lower frequencies will cause distortion of the flat top.

In some designs the bandwidth is determined by two circuits, the differentiating ($T_1 = C_1R_1$), that is the low frequency limit, and the integrating circuit ($T_2 = C_2R_2$), the high frequency limit or rise time of the amplifier (Fig. 9). In the diagram the input diamond circuit is shown as CR, and T is the resolving time of the crystal. C represents the total capacitance of the input circuit and R the feed resistor in shunt with C. The dotted lines imply that the circuits do not impose electrical loading on each other. A differentiating stage is necessary to shape the pulse and in this way it prevents pulses piling up on each other in later stages of the amplifier. These circuits are most conveniently placed in the grid circuit of the first valve of the main amplifier.

It is assumed that the diamond generates a step pulse of value $V=Q/C$ in a time T. The shape of the pulse will be determined by the time constant of the first circuit. By making the value of the grid leak large the pulse is given a long exponential tail; on the other hand, if we reduce the value of R we lose certain other advantages. This point will be discussed further when we consider the design of the preamplifier. The long tail of the pulse is immediately shortened by passing the pulse through the differentiating circuit having a time constant equal to or slightly longer than the collection time T. A time constant of this order will not distort the leading edge of the pulse. Gillespie[22] has pointed out that by making the integrating time constant T_2 equal to the differentiating time constant T_1, the best signal-to-noise ratio can be obtained. As already mentioned, an adjustable control is provided on the instrument so that these time constants can be varied to suit individual requirements.

If the amplitude of the output pulse is to be correctly preserved any one pulse must decay to 1 per cent. before the next pulse arrives, so the next shortest time constant in the amplifier is made 100 times longer than the time constant of the differentiating circuit; all other time constants are made at least 10 times this last value or 1,000 times that of the differentiating circuit.

5.4. *Linearity*

The linearity of the amplifier's amplitude response is important, particularly if the pulses at the output are to be separated into a number of energy channels. In the previous sub-section we showed how the amplitude of the pulse depends on the short time constant of a subsequent coupling. If we do not take this precaution the amplifier will distort the shape of the pulse and upset the amplitude/energy characteristic of the discriminator. A type of distortion known as "ringing" will occur every time a large pulse overshoots the base line and the resulting hump will be counted as an extra pulse, indeed, the number of humps produced will correspond to the number of coupling circuits in the amplifier.

To retain the shape of the pulse the valves throughout the amplifier should be operated on the linear portion of their characteristic curves. The main source of non-linearity is the last valve where the signal is largest. If we are to maintain a linearity of 1 per cent. or better, the grid excursion must be kept low by running the grid on the linear portion of the curve and as much negative feedback introduced as is consistent with reasonable gain. A blocked grid will suppress those small pulses which normally would have to be counted. We may remedy this fault by considerably reducing the value of the grid resistor in the final stage and adjusting the value of the coupling capacitor so that the time constant of the circuit remains appreciably the same.

5.5. *The Output Stage*

The output pulses from a pulse amplifier are always chosen to have a positive polarity so that a low impedance cathode follower stage can be used. This type of circuit finds its chief application as an impedance matching device and is particularly useful in cases where fast signals must be faithfully transmitted through a coaxial cable from one instrument to another. The important feature of the cathode follower is its high input impedance and its low output impedance. Furthermore, it is capable of maintaining linearity within a tolerance of 1 per cent. and giving high output voltages.

The output impedance of the circuit is approximately $R_k/(1+g_mR_k)$ where R_k is the cathode load resistance. This impedance is usually between 50 and 100 ohms. Where it is desired to feed more than one type of circuit from the output of the cathode follower the load resistance may be tapped to give a low impedance output of, say, 5 V at 50 Ω and a second of 100 V at an impedance of 1,000 Ω.

5.6. *Discriminator*

The output from the output stage of the amplifier is fed into the discriminator, which passes signals above a discrete amplitude.

The discriminator should be able to discriminate reliably between pulses which differ in amplitude by only a fraction of a volt, and at the same time, it should not overload on pulses much greater than the critical amplitude. The discriminator bias voltage is made adjustable to facilitate selection of any amplitude.

In some cases, the output pulses of various amplitudes from the discriminator are passed direct to an oscillograph for observation.

When the output from the discriminator is to be used solely to operate an output recording device, such as a rate meter, better results will be obtained if all the pulses fed to the recording instrument are of the same amplitude. Very often modern discriminators are designed to serve the dual purpose of suppressing signals below a predetermined level and making all those that trigger it a uniform shape and size.

Figure 10 shows the circuit of such a discriminator. In principle, it is a two valve flip-flop circuit with a variable cut-off bias applied to the grid of V1. The circuit is triggered to a new state when the voltage at point A goes above some critical value and

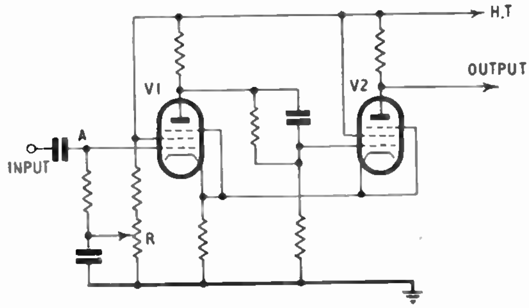

Fig. 10.—Discriminator circuit.

returns to its original state when the voltage drops below that level. The magnitude of the signal that will just trigger the discriminator is determined by the setting of the potentiometer R. The output signal is a square wave and, in our case, has an amplitude of 40 volts.

5.7. *Preamplifier Considerations*

The minimum signal that can be usefully amplified by any amplifier is limited by the magnitude of the noise generated in the amplifier itself, since a signal which is small compared with this noise will not be discernible or measurable at the amplifier output. All valves and resistors in the amplifier generate noise but in general the limiting sensitivity is determined by the noise voltages arising from the first valve and its associated components because such voltages are subject to the full amplification of the amplifier. In an amplifier, such as we are considering, the thermal and shot noise are the most important.

Thermal noise arises from the thermal agitation velocities of the free electrons in the conductor. The grid leak R in the pre-amplifier (Fig. 11) will function as a source of thermal noise but the circuits are modified somewhat by the stray capacitance C across the grid leak.

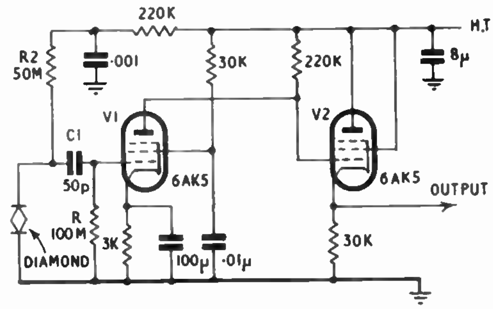

Fig. 11.—Preamplifier circuit.

Gillespie[23] has shown that the noise is proportional to $(1/R)^{\frac{1}{2}}$ and its contribution to the total amplifier noise is so small that it may be neglected. There are thus two good reasons for making R large, namely, (i) so that no charge leaks away during collection time and (ii) to reduce thermal noise to a minimum. R is always made very much larger than $1/\omega C$ and its value may be further increased if it appears to be noisy. For the same reason the load resistor R2 is made high. In actual fact an increase in R does not alter the total noise in the spectrum but simply concentrates the noise energy in the lower frequencies of the spectrum and effects a reduction in the noise passed by the amplifier having a restricted low frequency cut-off. This cut-off is determined by the differentiating time constant of the amplifier.

Shot noise arises from the fact that the emsision from the cathode is a large number of electrons which are random in character. The anode current of the valve will therefore have superimposed on it a random noise. Shot noise increases with bandwidth as the amplifier pass-band moves higher in the frequency spectrum.

To get the best results, then, one would choose a valve with a high mutual conductance and a low grid current and use it under the best operating conditions. We have the choice of a pentode or a triode. In a pentode we have to take into account the additional noise contributed by the screen grid current. This is one reason for preferring a triode as the first preamplifier valve but against this is the disadvantage that the capacitance of the input circuit of a triode is increased by Miller effect.

If necessary, we can improve the signal-to-noise ratio by reducing the bandwidth of the main amplifier but, if this procedure is carried too far, it will distort the shape of the incoming signal. Here again, it is a compromise and as a first consideration we fix the band-width of the main amplifier to give us the desired resolving time.

It has been found that in most experiments spurious pulses due to noise will not seriously interfere with the results if the signal-to-noise ratio exceeds five to one. In our particular case this means a minimum signal of 25 μV.

V1 is a pentode amplifier directly coupled to a cathode follower V2 and the values of the time constants of the circuit must be carefully considered. In many instances the preamplifier forms part of the probe and must be made small enough to be held comfortably in the hand. To achieve this we have had to reduce the size of the components repeatedly and have now stripped the circuit of everything but the bare essentials. Negative feedback has therefore been omitted. The design of the circuit is such that the h.t. on the anode and screen grid of V1 is only about 70 volts and the circuit itself appears to be fairly stable (Fig. 11).

5.8. *Complete Instrument*

The complete instrument based on the above principles has been constructed and is illustrated in Fig. 12. Whereas most commercial instruments are built in sections and rack mounted, our instrument was designed to be portable. The main amplifier, regulated power supply and output recording devices are all mounted on one chassis. Adequate screening prevents one circuit from interfering with another. The photograph shows the main amplifier with built-in loudspeaker, cathode ray tube and counting rate meter. The preamplifier is a separate unit and is connected to the main amplifier by standard cables. Attached to the preamplifier is the diamond probe which is about a foot long. Two spare probes can be seen in the foreground of the photograph.

6. Discussion

Work on the diamond counter has been confined to a few research workers whose preliminary investigations have done little more than reveal its advantages and disadvantages and our practical experience has been too short to add to these.

For our experiments we have used only the most sensitive diamonds, those not unduly affected by polarization, and the results we have obtained have been encouraging. Such diamonds will count all types of radiation (α, β, γ, x) and, provided the stones are no larger than 2 or 3 mm^3 in size, they will be suitable for pinpointing the source, particularly so if the diamond has been specially shaped.

From a practical point of view, the diamond's ability to count at room temperature and its density are attractive qualities, and the fact that the gamma ray absorption of both diamond (carbon) and human tissue are of the same magnitude, suggests a useful role in radiotherapy.

Fig. 12.—The author's equipment.

The small size of the diamonds, although advantageous for pinpointing a source, is, in another sense, a handicap because the pulse it generates is so small that a great deal of amplification is necessary to make it operate a recording device. Under present conditions, diamond pulses smaller than 25 microvolts are detectable only with great difficulty. To increase the sensitivity we should have to reduce the threshold noise of the input circuit and this will not be easy to achieve until we can get "noiseless" valves and components. Again, as we gain more experience in the polishing of counting diamonds it may eventually be possible to remove the poor counting material and expose the most sensitive layers.

The most serious disadvantage of the conduction counter is polarization and until this bugbear is entirely eliminated progress is likely to be restrained. The outlook, however, is much brighter than it was some time ago and as more "non-polarizing" counters become available, future progress should be more rapid. When it is proved that these particular diamonds will give consistently accurate readings, then electronic properties, like its fast resolving time of approximately 10^{-8} sec. and its linear response to sources of different energy, will attain even greater importance.

The diamond counter is only now emerging from the research stage and it is too early to compare its performance with that of other counters. Indeed, it was never intended that the diamond counter should be anything but complementary to counters in common use and up to now the policy has been to develop them for special purposes only.

7. Acknowledgments

The author gratefully acknowledges the help given by the Board of Management and the London Committee of The Diamond Producers Association. He wishes, too, to express his thanks for the assistance rendered by Dr. F. C. Champion, London University, and Dr. J. F. H. Custers, Diamond Research Laboratory, Johannesburg; Mr. R. E. Leeds, General Electric Company, London, for drilling the diamonds; Mr. A. B. Gillespie, Atomic Energy Research Establishment, Harwell, for technical advice; Dr. T. A. Chalmers, The Radium Institute, Liverpool, for many interesting discussions, and to Sir Francis Simon, F.R.S., Oxford University, for his interest in this work. Finally, it is a pleasure to acknowledge that Fig. 8 is the A.1 amplifier designed by Messrs. W. H. Jordan and P. R. Bell of the Los Alamos Laboratory, U.S.A.

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THE 40th PHYSICAL SOCIETY EXHIBITION

The number of stands at this year's Physical Society Exhibition of Instruments and Apparatus, held in London during May, was in actual fact rather fewer than in 1955; far more space was however made available for the items, since the Royal Horticultural Society's Old and New Halls were both used.

Mention was made last year of the number of exhibits showing the application of transistors: this year a notable feature was the the number of instruments and other pieces of apparatus making use of transistors in order to achieve compactness, or power economy, or both. Circuit techniques involved included oscillators and amplifiers, as well as switching devices; probably the most ambitious project was an experimental oscilloscope developed by Mullard in which transistor circuits were used for sweep generation, amplification and power conversion. Normally the instrument operates from a 12-volt accumulator but a mains unit can be employed. A 3-in. diameter c.r. tube displays the linear time base which produces sweeps of from 100 milliseconds to 100 micro-seconds; the amplifier, which has a bandwidth of 50 kc/s, is suitable for a.f. use, and for higher frequency inputs direct connection can be made to the plates.

Other "transistorized" devices included a counter-chronometer (also by Mullard), a time-base calibrator (B.T.-H.), an oscillator and amplifier for use with a.c. bridges (Tinsley) and a noise-cancelling microphone with transistor amplifier for telephone handsets (Siemens).

Developments in the semi-conductor devices themselves were numerous, and included a water-cooled germanium diode suitable for operation at mean rectified d.c. currents up to 300 amperes.

A prototype post-deflection acceleration c.r. tube shown by Twentieth Century Electronics makes use of a spiral p.d.a. electrode of resistive material deposited on the glass across which a 10-kV accelerating potential is applied. Although first put forward many years ago by von Ardenne, this array has only recently been exploited. It gives a more even potential gradient and does not call for a series of separate accelerating voltages.

Much research and development in different fields is being carried out by the manufacturers of electronic computers and items shown included a magnetic tape handling unit developed by Ferranti as an input/output device; a tape speed of 100 in./sec is used and the time taken to accelerate to full speed in either direction or to stop is claimed to be 10 milliseconds. With a digit density of 100 per inch the handling rate is 10,000 digits per second for each of the four information channels, giving a storage capacity of $1{\cdot}25 \times 10^6$ digits per channel (allowing for spacing of blocks of information). Plessey showed a new storage device using a barium titanate crystal which has a "read out" time of 0·2 micro-seconds. In this, charges can be established within the crystal at the intersections of strip electrodes mounted on either side; thus two sets of three electrodes give a matrix having nine intersections. Application of a higher voltage pulse than that which established the charge at a certain intersection gives an output pulse if it is of opposite polarity—if it is the same polarity no output is given. Although higher voltages are required than in comparable two-state magnetic stores, this electrostatic type is smaller and draws less power.

Decca Radar have developed a high speed serial digital computer in which the use of two-state magnetic cores is stated to have reduced the number of valves to about 200, thus reducing its cost, size and power consumption and increasing reliability. Input is by tape for data and instructions, and decimal data words are processed directly, obviating conversion programmes. The main store is a magnetic drum rotating at 6,000 rev/min and having a capacity of 4,032 words and an average access time of 5 milliseconds.

An aerial plotter designed for investigation of the amplitude and phase of the near field of a microwave radiator was shown in operation by Elliott Brothers. An r.f. signal is radiated from the aerial under test and picked up by a search aerial which moves across and in front of it. The search aerial output is mixed with a signal differing by 220 c/s from the test frequency, thus giving a signal whose amplitude and phase are recorded graphically. Near-field plotting presents a number of advantages over radiation plotting—less space required, less trouble from reflections and more definite indication of the cause of undesired features.

BRIDGE STABILIZED OSCILLATORS AND THEIR DERIVATIVES *

by

ir E. J. Post † and J. W. A. van der Scheer †

SUMMARY

The bridge stabilized oscillator with unspecified amplifying section can be regarded as a source of a wide class of feedback oscillators. Simple operations such as crosswise interchange of bridge elements, as well as suitable unbalancing of the bridge, will lead to most of the available feedback oscillator circuits as the (sometimes degenerate) derivatives of the bridge device. The principles involved are studied in some detail in this paper.

1. Introduction

The bridge stabilized oscillator as proposed by Meacham deserves the qualification "stabilized" for more than one reason. In Meacham's device the bridge has a two-fold function, i.e. frequency stabilization as well as amplitude control. The stabilization we mean in this paper is in particular the frequency controlling function of the bridge which is related to the phase discriminating properties of the circuit. The question of amplitude control which in itself is essential (but not sufficient) to generate constant frequencies will be left out of discussion, as technically it is not essential to unite the two functions in a single bridge device.

For feedback oscillators it is usually common practice to discriminate between a frequency determining part and an amplifying part of the oscillatory loop. A more conscientious application of this principle, however, immediately leads to the problem whether an unambiguous intersection can be made which corresponds to the above mentioned passive and active parts of the loop. The answer to this question is in general that no unique distinction can be made separating the two sections.

A particularly interesting aspect of this ambiguity has been discussed in a previous paper.[1] Let us assume that the bridge has an asymmetrically grounded input, then this leaves us with two floating output terminals.‡ The connection of the bridge output and the amplifier input can be interpreted in two distinct ways. Either we regard the amplifier as having a floating input, forming a single loop with the bridge circuit, or we regard the amplifier as having a differential input (i.e. two asymmetric input terminals with phase difference π). In the latter case the two wires leading from the bridge output to the amplifier input, together with the ground connections, can be interpreted as closing loops of positive and negative feedback, respectively, the amount of positive feedback being slightly in excess of the negative feedback in accordance with the available gain of the amplifier.

It is customary to include the selective frequency determining element in the loop of positive feedback. The aperiodic loop of negative feedback is then regarded as a means to improve the (phase) stability of the amplifier. At first sight, it seems illogical from this point of view to do it the other way around, i.e. aperiodic positive feedback and selective negative feedback. However, looking at it from the point of view of floating bridge output and floating amplifier input, the two aspects are perfectly identical. In this case there is only a single feedback loop, the properties of which are determined by the bridge circuit. The interchange of selective positive feedback and aperiodic negative feedback now corresponds to a crosswise interchange of bridge elements. The last operation leaves the properties of the bridge circuit unaffected. Hence an exchange of the

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U.D.C. No. 621.373.421.

‡ Similar arguments apply if we take a floating input of the bridge.

selective properties of the two feedback loops does not affect the performance of the oscillator.

Summarizing these introductory remarks we arrive at the following sets of equivalences to be used for practical oscillator design.

I { double feedback loop; positive loop is selective; negative loop is aperiodic

II { double feedback loop; positive loop is aperiodic; negative loop is selective

III { single feedback loop, e.g., original Meacham oscillator with floating bridge.

2. Analytical proof of the equivalences

The analytical proof is simple and hardly reveals the circuit-technical aspects of the equivalences. The latter will be dealt with in the last paragraph.

First we may establish the equivalence between the situations I and II. In both cases the oscillating loop should be considered to consist of an amplifier with single output and double antiphase input together with a bridge circuit with single input and double output or in reverse order; amplifier with double antiphase output and bridge circuit with double input.

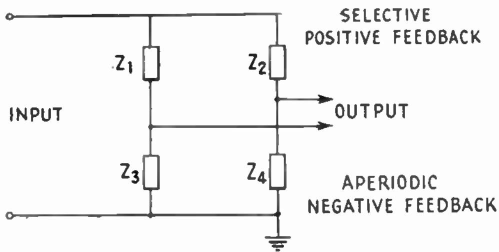

Fig. 1

The bridge circuit corresponding to the first configuration is drawn in Fig. 1. As a rule only one of the impedances, say Z_2, has phase discriminating properties, the others are supposed to be real and constant. The impedance Z_2 may be thought of as a quartz crystal operating near its low impedance resonance. The transmissions along the output loops (for infinite input impedances of the amplifier), are:

$$f = \frac{Z_4}{Z_2 + Z_4} \quad \text{positive feedback; selective} \qquad \ldots(1)$$

$$a = \frac{Z_3}{Z_1 + Z_3} \quad \text{negative feedback; aperiodic} \qquad \ldots(2)$$

A crosswise exchange of the bridge elements is expressed by the transformation:

$$\begin{matrix} Z_1 \to Z_4 & Z_4 \to Z_1 \\ Z_2 \to Z_3 & Z_3 \to Z_2 \end{matrix} \qquad \ldots\ldots\ldots(3)$$

The transformation (3) affects the loop transmissions in the following way:

$$\frac{Z_1}{Z_1 + Z_3} = 1 - a \quad \text{positive feedback; aperiodic} \qquad \ldots\ldots\ldots(4)$$

$$\frac{Z_2}{Z_2 + Z_4} = 1 - f \quad \text{negative feedback; selective} \qquad \ldots\ldots\ldots(5)$$

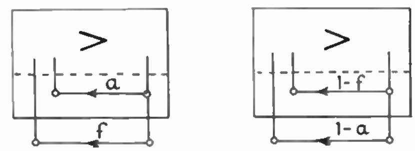

Fig. 2

If we now turn to the complete loop, the situations I and II corresponding to formulae (1), (2) and (4), (5) are represented in Fig. 2. The negative feedback is considered enclosed in the amplifier.

The implicit information about the frequency of oscillation is given by the roots of the frequency equation. Bode's return difference $D(\omega)$ in the loop of positive feedback may be taken as an appropriate network function for our general case. The complex frequency of oscillation is determined by the equation:

$$D(\omega) = 0 \qquad \ldots\ldots\ldots(6)$$

The return differences corresponding to the situations I and II, if A is the complex gain of the amplifier without feedback, are:

$$D_{\mathrm{I}}(\omega) = 1 - \frac{Af}{1 + aA} = \frac{1 + aA - fA}{1 + aA}, \qquad \ldots\ldots\ldots(7)$$

$$D_{\mathrm{II}}(\omega) = 1 - \frac{A(1-a)}{1 + A(1-f)} = \frac{1 + aA - fA}{1 + A(1-f)} \qquad \ldots\ldots\ldots(8)$$

The roots of (7) and (8) are determined by the same equation, the common enumerator of (7) and (8), i.e.

$$D(\omega)=1+aA-fA=0, \quad \ldots\ldots\ldots(9)$$

as the divisors $1+aA$ and $1+A(1-f)$ are different from zero, if (9) holds. This argument, as a matter of fact, excludes the unrealistic cases of $1+aA$ or $1+A(1-f)$ tending to infinity.

The practical importance of (7), (8) and (9) is contained in the fact that the frequencies of oscillation in situations I and II are determined by the parameters of the same equation, namely equation (9). Hence if situation I in principle yields a highly stable oscillator, the same will hold for situation II, proving the equivalence between the two arrangements I and II.

As a next step we may prove that (9) corresponds to situation III. The difference between the output of the positive and the negative loop of feedback is equal to:

$$(f-a) \quad \ldots\ldots\ldots(10)$$

After crosswise exchange of the bridge elements (the third transformation) it is still:

$$(1-a)-(1-f)=(f-a) \quad \ldots\ldots\ldots(11)$$

The floating single output of the bridge is led to a single floating input of an amplifier with gain A. Hence the return difference for either configuration of situation III becomes:

$$D_{\mathrm{III}}(\omega)=1-A(f-a)=D(\omega), \quad \ldots\ldots(9a)$$

which is identical to (9).

Putting $D_{\mathrm{III}}(\omega)=0$ yields the same roots. This completes the analytic proof of the equivalences I, II and III.

The equivalence between I and III had already been established in the former article, using methods of network geometry (see corresponding note in ref. 1 for the suggestion of the analytic proof).

3. Circuit possibilities

The "source" circuit used to obtain the derived circuits is, for the time being, an unspecified amplifier with a bridge circuit as a general feedback element as shown in Fig. 3. The steps to be accomplished to produce a family of oscillator circuits are the following:

(*a*) A specification of the amplifier section should give us information about the nature of the output and input terminals of the active section. Let us assume that we have an asymmetric grounded output, terminal B having ground potential. This leads to a twin input. We will suppose A' to be the terminal of positive feedback and B' the terminal of negative feedback, i.e. A and A' are in phase whereas A and B' are supposed to be in antiphase. Using a quartz crystal near its minimum impedance frequency requires that Z_2 should be the selective element for the positive feedback to be maximum near resonance. The impedances Z_1, Z_3 and Z_4 are taken as real and constant. Using a quartz crystal near its maximum impedance frequency, requires that Z_4 should be the selective element, whereas Z_1, Z_2 and Z_3 are taken as constant and real.

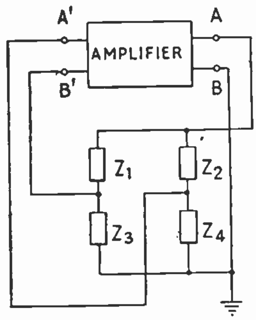

Fig. 3

(*b*) Once we have our amplifier specified, a crosswise exchange of bridge-elements will lead to another equivalent oscillator circuit with a selective negative feedback loop. The negative feedback will tend to be a minimum near the frequency of oscillation as such enabling the positive feedback to do its work for one specified frequency only.

(*c*) The two equivalent oscillator circuits are submitted to a simplification in the form of shorting the impedance Z_3, the diagonal element of the selective quartz crystal in the case of minimum impedance resonance. This procedure is either equivalent to a reduction of the aperiodic negative feedback to zero or, after crosswise exchange, equivalent to an increase in the positive aperiodic feedback to a maximum value. The shorting of Z_3, which should be regarded as an operation leading to degenerate circuits is interesting, because it produces most of the existing oscillator circuits.

Summarizing the operations we have: Selection of a specified amplifier yields an oscillator circuit. Crosswise exchange of the bridge elements leads to another equivalent oscillator circuit. Shorting Z_3 gives us in addition two degenerate circuits, making a total of at least four circuits belonging to the one amplifier section which has been chosen as a starting point.

Example A

The amplifier chosen for this case is shown in Fig. 4a; it has a cathode follower output (asymmetric) and a twin input using cathode and grid of the other tube as terminals of positive and negative feedback respectively. The derived circuits have been drawn in Figs. 4b-e.

The oscillator shown in Fig. 4b is characterized by a selective positive feedback and an aperiodic negative feedback. (A description of this oscillator will be found in Ref. 1.) Shorting Z_3 yields the simplified version (Fig. 4c) which has been discussed in Refs. 2 and 3.

Application of transformation 3 to the oscillator shown in Fig. 4b will yield a new circuit (Fig. 4d) which probably has not previously been discussed. Except for the advantageous feature that one of the crystal terminals is grounded, the performance of the circuits of Figs. 4b and d can be expected to be equivalent. Shorting Z_3 in the present situation leads to the circuit shown in Fig. 4e. This oscillator is discussed in Ref. 4.

Example B

The amplifier section to be used in example B is the same as that in example A, except that instead of a cathode output, an anode output has been used. The relatively high output impedance of the amplifier is advantageously applied for bridges using crystals with relatively high (minimum) impedance. The amplifier is shown in Fig. 5a.

Connecting the bridge leads at once to a circuit shown in Fig. 5b. Shorting Z_3 gives the original Heegner oscillator (Fig. 5c); a description of this circuit is found in Ref. 5.

Applying the rule of crosswise exchange in the bridge yields the oscillator shown in Fig. 5d. Shorting Z_3 plus a slight modification of the basic circuit, produces the circuit of Fig. 5e.

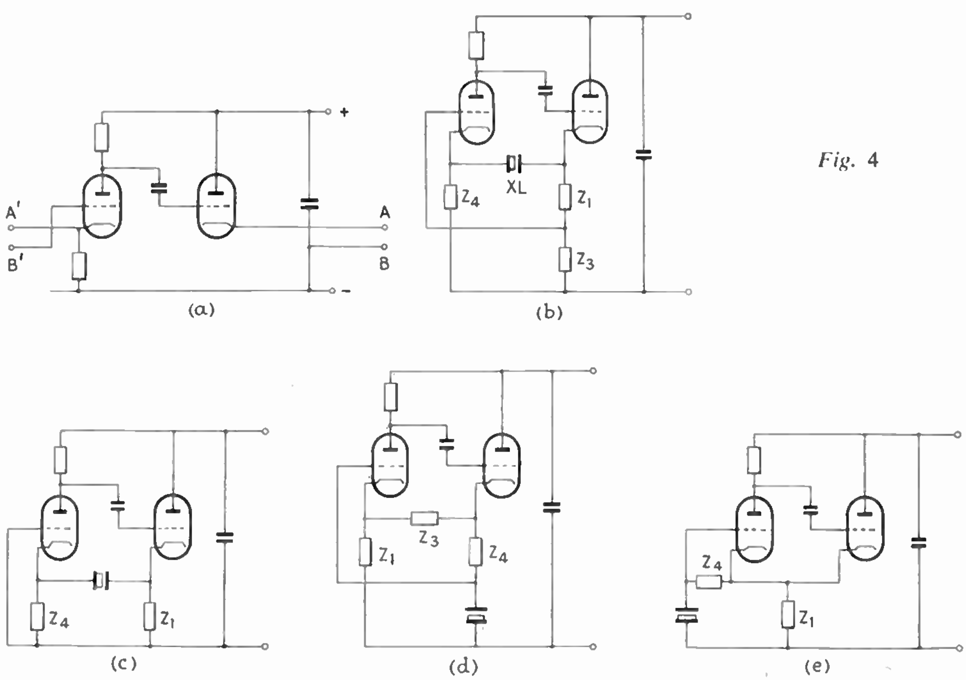

Fig. 4

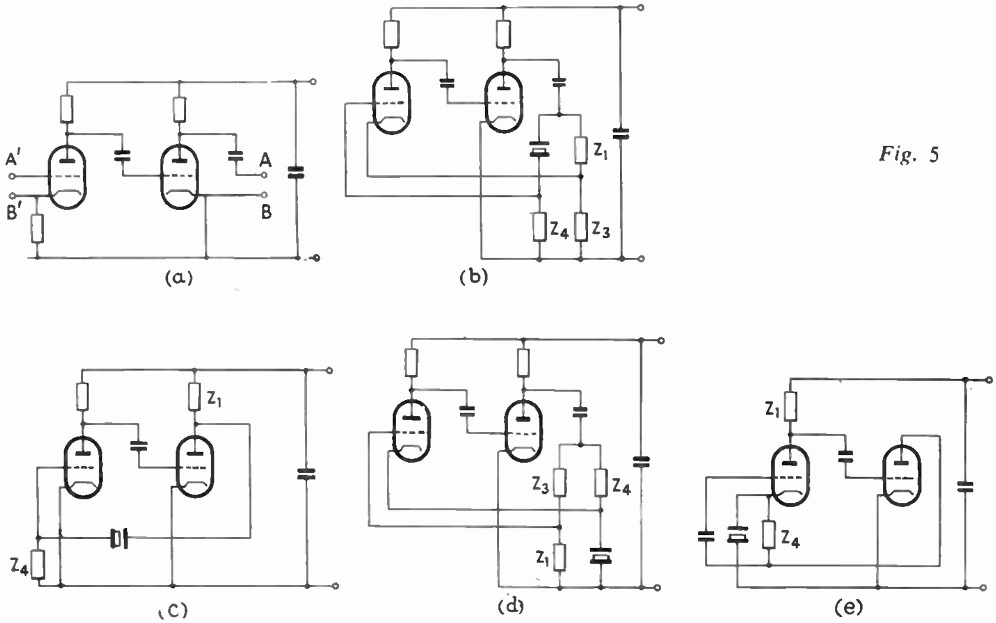

Fig. 5

Example C

Another amplifier section which is suggested by the well-known CI-meter is shown in Fig. 6a. Insertion of small resistances in the capacitive as well as in the inductive branch of the tuned circuit provides for the possibilities of positive and negative feedback.

The circuit drawn in Fig. 6b is a practical realization with the crystal in the positive feedback loop. Shorting Z_3 then leads to the

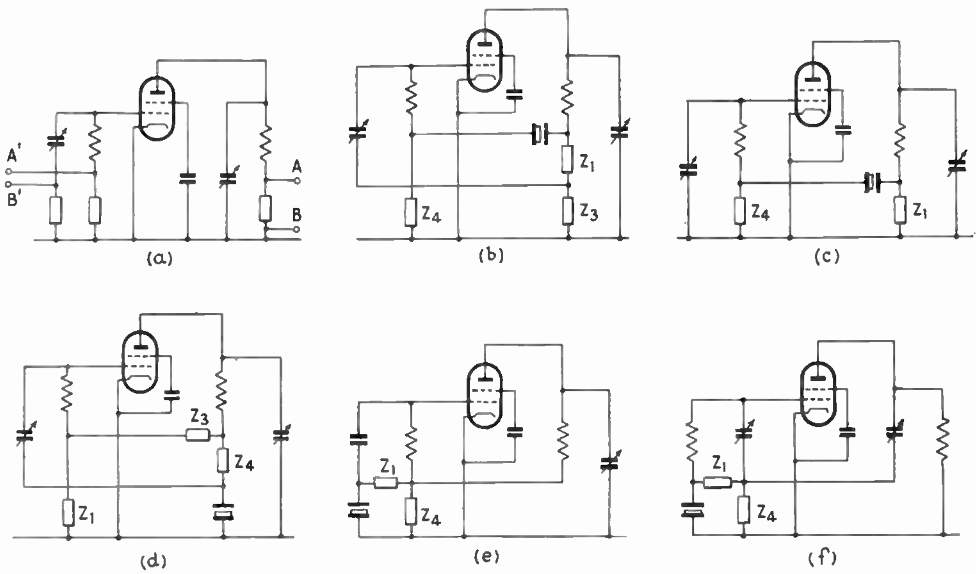

Fig. 6

conventional CI-meter[6] shown in Fig. 6c. The bridge exchange operation yields the circuit dual to Fig. 6b with one of the crystal terminals grounded. Shorting of Z_3 again gives the simplified arrangement, which is indeed an interesting new version of the conventional CI-meter. The oscillator shown in Fig. 6f is identical to Fig. 6e, except for the fact that a split stator capacitor can be used. This feature is brought about by an obvious, and minor, change of the original amplifier circuit in Fig. 6a.

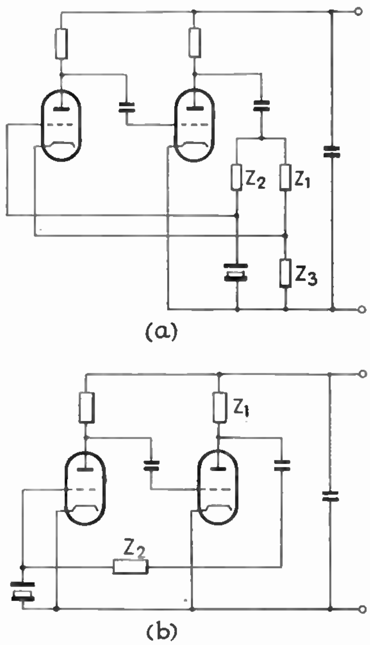

Fig. 7

Example D

As a last example we may consider the possibility of a high impedance bridge using a crystal near its high impedance resonance. Instead of Z_2 we now regard Z_4 as the selective element. Taking the amplifier section of Fig. 5 leads to Fig. 7a and b. The second (Z_3 shorted) is known as the Heegner oscillator using the crystal in anti-resonance. The other two members of this oscillator family will be left out of discussion for practical reasons.

4. Conclusion

Using a set of three circuit equivalences, it has been shown that many existing oscillator circuits are related to a correspondingly more general and fundamental bridge oscillator.

Theoretically, every appropriate amplifier will lead to a family of at least four circuits. At least two members of the family are degenerate in the sense that the bridge is unbalanced, which may lead to a mismatch between amplifier gain and insertion loss of the frequency determining four-terminal network.

Henceforward the performance of many existing oscillator circuits might be improved by completing and balancing the bridge, whereas other circuits are obtained by means of the rule of crosswise exchange of bridge elements.

The examples given in the last section can be extended by the choice of other amplifiers. In particular amplifiers using a twin output, instead of a twin input, may be considered. Another possibility is the application to transistor amplifiers.

5. References

1. E. J. Post and H. F. Pit, "Alternate ways in the analysis of a feedback oscillator and its application," *Het PTT-Bedrijf*, **3**, p. 39, 1950, and *Proc. Inst. Radio Engrs*, **39**, pp. 169-171, February 1951.
2. F. Butler, "Cathode-coupled oscillators," *Wireless Engineer*, **21**, pp. 521-6, November 1944.
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6. A. C. Prichard and M. Bernstein, "Crystal impedance meters replace test sets," *Electronics*, **26**, pp. 176-80, May 1953.

. . . Radio Engineering Overseas

A selection of abstracts from European and Commonwealth journals received in the Library of the Institution. All papers are in the language of the country of origin of the journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

519.272:621.372.54

Correlation functions and their uses—a review.—R. SETHURAMAN. *Journal of the Institution of Telecommunication Engineers*, **2**, pp. 87-92, March 1956.

Limitations of Fourier analysis are examined and the merits of correlational analysis are discussed. The definition and properties of correlation analysis are studied. The use of correlation functions in the study of (a) the transient response of filters, (b) the spectra of delay modulated signals, (c) the signals in the presence of noise, and (d) the design of linear filters are studied. A typical correlator is described. Limitations of this method of analysis are examined.

522.6:621.396.9

A receiver for the radio waves from interstellar hydrogen. II. Design of the receiver.—C. A. MULLER. *Philips Technical Review*, **17**, pp. 351-361, June 1956.

This article deals with the design of a receiver for radio waves with a frequency of 1,420 Mc/s which are emitted by interstellar hydrogen. In order to be able to distinguish the very weak signal from the noise inherent in the receiver, a method of frequency comparison is employed. The receiver is switched to and fro at a rate of 400 times per second between two frequencies, one lying on the spectral line and the other adjacent to it. The spectral line is traversed by slowly varying both frequencies simultaneously while maintaining a constant difference of 1,080 kc/s between them. Two separate i.f. channels are used in order to keep the receiver tuned to the spectral line all the time (instead of only half the time, as would be the case if only one i.f. channel were used). Each channel has its own comparison frequency on opposite sides of the measuring frequency, and each is tuned to the line in turn. After further mixing and subsequent detection the two signals are fed to a push-pull amplifier, where they are added together and where unwanted voltages are eliminated. The fundamental frequency of the 400 c/s square-wave voltage, which is a measure of the intensity of the interstellar radiation, is rectified in a tuned detector circuit and, after emerging from a low-pass filter of 1/200th c/s bandwidth, it is fed to a strip-chart recorder. The article deals finally with the noise characteristics of the h.f. section and of the first mixer, and discusses the factors which determine the sensitivity of the receiver. The sensitivity achieved is better than 0·1 per cent. of the inherent noise. In terms of antenna temperature, this means that it is possible to detect temperature differences down to about 1°K.

534.2

Determination of amplitude distribution on planar surfaces from the directional distribution of their radiation fields.—K. FEHER. *Archiv der Elektrischen Übertragung*, **10**, pp. 125-131, 163-173, March and April 1956.

The amplitude distribution of an oscillating surface is related by a Fourier integral to its distant field directional radiation pattern. If one of these two quantities is known by magnitude and phase, the other can be determined from it. This paper gives firstly the basic theory for the determination of the amplitude distribution by a Fourier analysis of its directional radiation pattern, secondly a description of an experimental method, and thirdly the checking of the measuring method with the aid of oscillating plane plates and electrostatically excited diaphragms. The method finds use in ultrasonics, where direct measurement of the amplitude distribution with a detector for solid-borne sound is difficult.

538.123

End-fire arrays of magnetic line sources mounted on a conducting half-plane.—R. A. HURD. *Canadian Journal of Physics*, **34**, pp. 370-7, 1956.

Formulae are developed for the far-field radiation patterns of end-fire arrays of magnetic line sources mounted on the surface of a half-plane. Patterns have been plotted for a number of interesting cases. It is found that the Hansen-Woodyard optimum end-fire condition no longer holds when the array is near the edge of the half-plane. The theory seems to describe reasonably well the behaviour of corrugated surface radiators embedded in a finite ground plane, providing the array is removed from the edge by a distance about equal to the array length.

621.317.76

The frequency microscope; a frequency measuring instrument of very high sensitivity.—G. OHL. *Archiv der Elektrischen Ubertragung*, **10**, pp. 145-50, April 1956.

The "frequency microscope" permits of recording, with high sensitivity, the relative deviation $\Delta_r f$ of two frequencies in the neighbourhood of discrete, very closely spaced values $\Delta_r f_0$. The two frequencies are heterodyned to give pulsations. A sequence of time markers produced by these pulsations is recorded on a drum type chronograph whose drum frequency is so chosen that a family of parallel curves results. From the direction of their tangents the "instantaneous" $\Delta_r f$ values can be determined with a very high accuracy by a simple graphical or metering method. Averages over any desired intervals of time can also be read in the same manner from the direction of the chords of the curves. Since as a rule a continuously variable oscillator is not stable enough for driving the drum, a large number of drum frequencies is provided in fine steps by dividing the fixed frequency of a quartz crystal clock on the electronic counting principle. The calculation of a favourable frequency division ratio is the principal problem in preparing a recording, and the method adopted by the author is given for this purpose.

621.372.413

Some investigations at microwave frequencies.—S. K. CHATTERJEE. *Journal of the Institution of Telecommunication Engineers*, **2**, pp. 67-86, March 1956.

A brief review of the investigations at 3 cm wavelength carried out at the Indian Institute of Science is given. The work includes theoretical investigations on the circuit aspects of microwave cavities, perturbation effects in a cavity resonator and interaction of modes. The paper also gives a report on the results of investigations on the propagation characteristics of microwaves through a cylindrical metallic guide filled completely with two concentric dielectrics. A method of calculating the attenuation and phase constant of a hybrid $(EH)_{11}$ mode propagating through an imperfectly conducting guide filled with an imperfect dielectric is also discussed. A new method for the determination of conductivity of metals at microwave frequencies is outlined. A survey of the theoretical investigations on a parallel metal plate dielectric and their experimental verifications with the help of a microwave interferometer constructed for the purpose is also given. The paper concludes with a discussion of theoretical and experimental investigations on the radial field spread, in the propagation of microwaves on a single wire.

621.373.423

Modern reflex klystrons.—R. HECHTEL. *Archiv der Elektrischen Ubertragung*, **10**, pp. 133-8, April 1956.

After a short survey of the merits and demerits of the reflex klystron, an outline is given of the design considerations leading to an improvement of its characteristics. Three modern klystrons are presented as examples of the latest development in this field. At frequencies around 4,000 Mc/s efficiencies of 5 per cent. have been obtained with bandwidths of 30 Mc/s, and of 2 per cent. with bandwidths of 45 Mc/s. In the 7,000 Mc/s band the attained efficiency is 3·5 per cent. with a bandwidth of 25 Mc/s.

621.375.122

Problems related to zero drift of d.c. amplifiers.—T. ALEKSIC. *Elektrotehniski Vestnik*, **23**, pp. 396-402, November-December 1955.

Causes of the zero drift of d.c. amplifiers are discussed, the most serious being the random character of the cathode drift. An analysis is given of errors due to drift fo the d.c. amplifier being part of a feedback system loop. The error has finite limits or increases steadily with time, depending on the system of parameters. A survey of known methods of zero drift compensation is given.

621.375.3:621.318.435

Behaviour of saturable reactors in magnetic amplifiers.—P. N. DAS. *Indian Journal of Physics*, **30**, pp. 129-142, March 1956.

The exact behaviour of saturable reactors when used in a magnetic amplifier depends not only on the nature of the material of the core and on the magnitudes of d.c. and a.c. excitations used, but also on the external circuit conditions, and on such factors as the number of cores used, the nature of the d.c. source and the way in which a.c. and d.c. windings are connected. A number of different cases have been studied, for the following conditions: a single core with d.c. source of infinite or low impedance; two cores series connected with d.c. source of infinite or low impedance; two cores parallel connected with d.c. source as above. The differences observed in the shape of the *B-H* loop and the a.c. wave form, and the variations in current and flux density, are explained from fundamental considerations.

621.385.3:621.375.122

The selection of triode valves and circuits for direct coupled amplifiers.—R. E. AITCHISON. *Journal of the Institution of Engineers, Australia*, **27**, pp. 339-41, December 1955.

The amplification factors of commercial triode valves lie within narrow limits, and circuits which depend primarily upon this property show greatest uniformity and stability. The use of twin triodes gives pairs of valves which are better matched than triodes in separate envelopes. Further selection for equality of anode currents and their mutual conductances can give pairs of valves with parameters nearly equal. Of the various d.c. amplifier circuits the Artzt circuit or the balance version of this circuit gives the amplifier with the most uniform and stable characteristics.

621.396.931.029.62

Some particular aspects of mobile v.h.f. equipments. H. RENARD and M. BRULEY. *Onde Electrique*, **36**, pp. 349-52, April 1956.

The design of mobile radio telephone equipments, particularly for urban areas with traffic congestion, presents difficult problems of propagation and interference between a number of channels, tightly pressed into a narrow frequency band. Such problems can only be solved by a close collaboration between equipment manufacturers and the authority allotting the frequencies. In particular, the use of duplex working, although it appears to facilitate the allocation of frequencies, presents more difficult problems to the manufacturer.

621.396.965.8

Automatic tracking radar systems.—P. BOUVIER. *Onde Electrique*, **36**, pp. 336-47, April 1956.

After a brief historical survey, the author outlines some of the basic principles used in an automatic tracking radar system. Parameters particular to this type of radar are then analysed in the case of a classical tracking radar Finally, a French radar in large-scale production is briefly described.

Extended summaries of the following papers from *Archiwum Elektrotechniki, Warsaw*, **5**, No. 1, January 1956, are available in English:

Frequency stability of LC oscillators with large grid and anode capacitances.—J. GROSZKOWSKI. (pp. 35-66.)

The design of pulse systems based on the method of moments.—R. KULIKOWSKI. (pp. 69-79.)

The magnitude method of evaluating the characteristic equations of automatic control systems.—P. NOWACKI. (pp. 107-132)

The excitation of a cavity resonator by a density-modulated electron beam passing through the entire resonator cross-section.—P. SZULKIN. (pp. 149-206.)