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(FOUNDED IN 1925 - INCORPORATED IN 1932)

" To promote the general advancement of and to facilitate the exchange of information and ideas on Radio Science."

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THE PRESIDENTIAL ADDRESS

of

L. H. Bedford, O.B.E., M.A.

(Delivered at the 23rd Annual General Meeting of the Institution in London, on October 21st, 1948, and repeated before the West Midlands Section on December 15th.)

I wish first of all to express my thanks to the Membership for the honour conferred upon me in electing me President of the Intitution. I am particularly conscious of this in view of the distinguished Presidency that I am following.

It is an established custom that a presidential address should be of a broad and general character; I find it however much more within my scope to be specific and quantitative and thus find myself somewhat in the position of the mathematics student, who, confronted with a compulsory paper on General Physics, found himself unable to answer a single question. He therefore selected one question at random, which was: " Describe with diagrams the action of any one type of vacuum pump," and wrote as follows:— " Before describing the action of a vacuum pump, it is necessary to prove the Binomial Theorem . . . ."

Notwithstanding that limitation I am well aware that my election comes at a time of crisis in many directions. The radio industry, and with it therefore the radio profession, faces one of its own at this time; neither the first, of course, nor the last. I feel that in these circumstances it will certainly be of interest, and possibly even useful, to endeavour to take stock of the position of the radio art, looking for the most part into the past by way of assisting us to imagine the future.

At the age of 75 years radio is one of the youngest of the sciences, but it is still a formidable task to attempt a review within the period of an hour. More especially is this so, as it is not my intention to limit this review to a mere process of cataloguing.

The science of radio traces its origin to the year 1873 with Maxwell's theory. What was this Maxwell's theory? As I found myself unable to answer this question to my satisfaction without revision, and as it is one which lies at the very roots of our whole subject, I feel that it will be proper to commence with just this process of revision.

Prior to Maxwell, the subject of electromagnetism had not been free from the attention of able mathematicians and inspired physicists. It was probably Maxwell's combination of great mathematical ability and amazing physical insight that enabled him to make his extraordinary contribution. The mathematicians, headed perhaps by Weber, had mostly concerned themselves with the conception of " action at a distance." Faraday, on the other hand, thinking physically, emphasized the opposite point of view, regarding action at a distance as unphysical and explaining all forces in terms of a physically existant medium, the æther. Maxwell gave mathematical expression and clarity to Faraday's ideas, but took them much further. In particular, he introduced the idea of " electric displacement." This is a conception which is clear enough in the case

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of polarized matter, but Maxwell extended it also to the hypothetical æther, which he regarded as capable of electric polarization or strain. The conception of electric displacement led to that of displacement current, this occurring both in matter and free space whenever there occurs a rate of change of electric force.

Maxwell's theory of light can now be sketched in a few lines of algebra. It may perhaps be considered somewhat out of order to include algebra in a Presidential Address, but, apart from implications of my introduction, I submit that Maxwell's theory cannot be exhibited in any other way, and that, in view of what follows, any attempt to express it physically rather than mathematically is futile.

The experimental relations between the previously quite separate subjects of electricity and magnetism had been determined by Ampère and Faraday in terms of the so-called "circuital relations." That of Ampère states broadly that the magnetomotive force round a loop is proportional to the current through it. That of Faraday states that the electromotive force round a loop is proportional to the rate of change of magnetic flux through it.

To formulate a field theory, Maxwell reduced these relations to differential form :

$$\text{curl } \mathbf{H} = k_6 \mathbf{i} \text{ (i being the current density)}$$

$$\text{curl } \mathbf{E} = -k_7 \dot{\mathbf{B}}$$

What do these equations become in free space? For the second equation we merely substitute **H** for **B** and change the constant to take account of units :

$$\text{curl } \mathbf{E} = -k_8 \dot{\mathbf{H}}$$

The first equation contains the new and startling contribution. Prior to Maxwell, one would be inclined to write it

$$\text{curl } \mathbf{H} = 0,$$

expressing the fact that in free space there could be no current density. But Maxwell states that we have to regard $\dot{\mathbf{E}}$ as contributing a current density, whereupon the Ampere equation becomes

$$\text{curl } \mathbf{H} = k_9 \dot{\mathbf{E}}$$

These two equations lead by a few lines of vector algebra to

$$\nabla^2 \mathbf{H} = -k_8 k_9 \frac{\partial^2 \mathbf{H}}{\partial t^2}$$

and similar equation for **E**.

This equation is a well-known one in mathematical physics and tells us that **H** and **E** are propagated through space with a velocity $1/\sqrt{k_8 k_9}$. The identification of this constant with the velocity of light leads to Maxwell's theory of light.

Maxwell had no means of experimentally demonstrating the existence of electro-magnetic waves as such. This achievement fell to Hertz in 1887. I quote from a previous Presidential Address by Dr. Garrard ; he writes, "I well remember as a student some few years back how excited we were to see small electric sparks passing across an air gap in a ring of wire induced by discharge from an induction coil at the other end of the lecture theatre."

But was this necessarily wireless transmission in the Maxwell sense? The conception of coupled circuits does not involve Maxwell's theory and the whole question turns on the quantitative aspect. To illustrate this let us consider the field of a Hertzian oscillator or dipole calculated according to Maxwell's theory. This may be written

$$\left[E_r = 2A \cos\theta \left\{ \frac{1}{r^3} \sin p\left(t - \frac{r}{c}\right) + \frac{1}{r^2}\frac{p}{c} \cos p\left(t - \frac{r}{c}\right) \right\} \right]$$

$$E_\theta = A \sin\theta \left\{ \frac{1}{r^3} \sin p\left(t - \frac{r}{c}\right) + \frac{1}{r^2}\frac{p}{c} \cos p\left(t-\frac{r}{c}\right) - \frac{1}{r}\frac{p^2}{c^2} \sin p\left(t-\frac{r}{c}\right) \right\}$$

$$H_\phi = A' \sin\theta \left\{ \frac{1}{r^3} \cos p\left(t - \frac{r}{c}\right) - \frac{1}{r}\frac{p}{c} \sin p\left(t - \frac{r}{c}\right) \right\}$$

The first term in each expression is what one would erroneously calculate by a quasi-static method. The later terms are the Maxwell contribution and the important thing to notice is that they are of a lower order in $\frac{1}{r}$ than the quasi-static terms and therefore become predominant as the distance is increased.

The other quantity of interest is the Poynting vector indicating the rate of energy flow in space.

The radial component of this has the magnitude $E_\theta H_\phi$. If we form this product from the above expressions it will be noticed that the first terms in the two expressions multiply together to give a wholely oscillatory expression, that is to say, there is no net loss of energy to the system. The product of the last two terms, however, contains a D.C. term indicating a uniform outflow of energy. This underlies the idea of " radiation resistance."

Teaching or learning, I have found this problem in radiation theory basically explicit.

We come now to the appearance of Marconi in 1896. What in fact were the principal Marconi contributions ? They would seem to be first an appreciation of the necessity of an aerial showing a relatively high proportion of radiation resistance. Secondly, appreciation of the significance of resonant tuning, an aspect also emphasized by Lodge. Thirdly, and perhaps most important of all, that he approached the subject always as a communication problem rather than as an abstract piece of scientific research.

The above three names, Maxwell, Hertz and Marconi can perhaps be considered to define the foundational phase of the subject. Before leaving the basic theory, which I have treated in more detail than I will be able to devote to the subsequent developmental phases, I wish to refer to a fundamental element of paradox.

In 1905, Einstein formulated the special theory relativity, in 1915 the general theory, and in 1919 the solar eclipse expedition completed the experimental confirmation of the predictions of Einstein's theory which has since become a fully accepted doctrine of theoretical physics.

From the radio point of view, the principal repercussion of Einstein's theory is the abolition of the æther. What then happens to Maxwell's theory ? The answer to this is that Maxwell's theory reduces to Maxwell's equations. These stand unaffected and fully expressive of the laws of nature in electro-magnetism. When correctly applied within their proper domain, they always give the correct answer.

You may wish to describe this situation as the building standing after the scaffolding has been removed. I think it is something much more startling ; it is as though someone kicked away my chair and I remained seated in space ; or it is like Lewis Carroll's grin on the face of the Cheshire Cat ; the Cat faded but the grin remained.

The idea that a theory should express the facts of nature without regard to any comprehensible mechanism is one which is now well accepted in physics, and radio engineers must align themselves with this view point.

With Marconi we come to the romantic phase of radio development ; it is well mile-stoned by certain of Marconi's own achievements, viz.:—

1900 First cross-channel radio, Wimereux to Dover.

1902 First transatlantic radio, Poldhu to Newfoundland.

I read that the demonstration of this achievement consisted in Marconi " snatching a letter S from the æther." I have myself frequently snatched letter S's from the æther without the co-operation of any specific transmitter, but any suspicion that the Marconi result might have been spurious was dispelled when Marconi returned on S.S. *Philadelphia* and presented recorded morse signals received from Poldhu at various stages of the crossing.

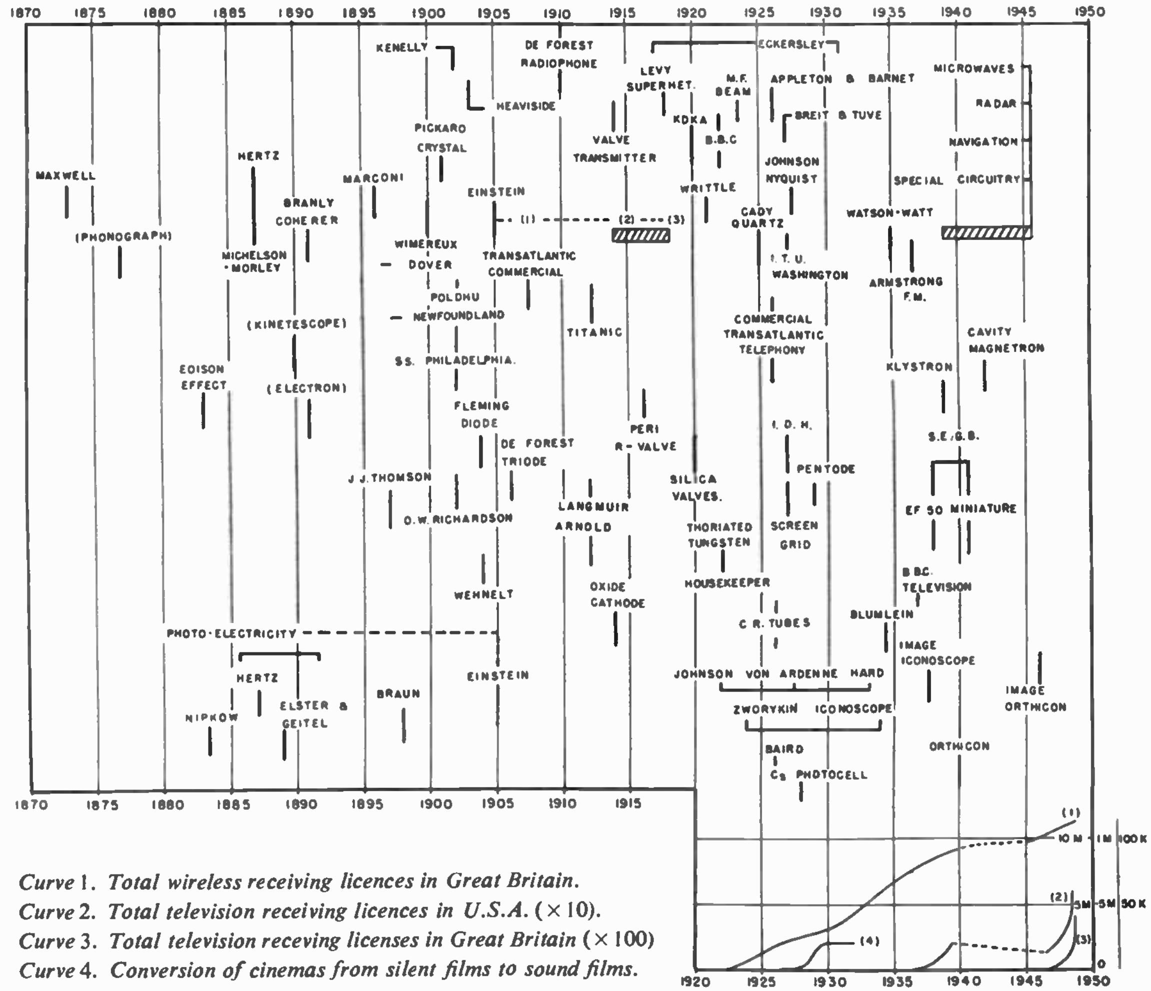
In 1907 commercial transatlantic telegraphy was established.

In 1912 the contribution of radio to the rescue work in the *Titanic* disaster brought the significance of radio at sea dramatically before the public eye and marine radio application can be reckoned to have entered on its intensive phase from that time.

We may here pause to observe how the achievements of these earlier radio giants went beyond the technical equipment of the time.

Maxwell's equations, for instance, look very simple in vector notation and better still in a good system of units. But these things hardly existed, and Maxwell hewed his equations out of the formidable Cartesian stone. One has only to read this work to understand Silberstein's remarks (1913) on the subject of Cartesianification :—

" Very often such a procedure gives rise to a hopeless complication of the scalar formulæ, a complication which does not arise from the intrinsic peculiarities of the phenomena in question, but is wholly artificial. . . . Now Nature is herself wonderfully complicated, so



Curve 1. *Total wireless receiving licences in Great Britain.*
Curve 2. *Total television receiving licences in U.S.A.* (× 10).
Curve 3. *Total television receving licenses in Great Britain* (× 100)
Curve 4. *Conversion of cinemas from silent films to sound films.*

that supplementary complication is not wanted."

On the practical side, Hertz performed his original experiments without anything that could normally be called a detector. When Marconi entered the field he had only the coherer of Branly (1891), which he greatly improved and, indeed, commercialized in so far as a device of this character can be called commercial. The crystal detector came only with Pickard in 1906.

We now reach a point when the whole art becomes revolutionized by the introduction of the thermionic valve. A separate line of the chart will serve to correlate this branch of the subject. The principal events are :—

1883 Edison effect.

1891 Johnstone Stoney christened the electron.

1897 J. J. Thomson "liberated" the electron.

1902 O. W. Richardson's basic work on thermionic emission.

1904 The Fleming diode.

1906 The De Forest triode (audion).

The last two names certainly bring their own romance to this phase of radio development. De Forest in particular had something of Marconi's flair for dramatic demonstration. Thus within three years of the appearance of the audion De Forest achieved the world's first radio broadcast, 1910, a transmission of Caruso from the Metropolitan Opera House.

(Diverting our attention for one moment to the social implications of radio, one may here lament that in later days of American broadcasting the Metropolitan Opera was due to give place to the Soap Opera.)

The De Forest radiophone used an audion amplifier to modulate an arc transmitter. Valve transmitters as such appeared in 1914. This date seems to correlate naturally with the work of Langmuir and Arnold in 1912 which taught the value of high vacuum in thermionic devices and showed how to get it.

We come now to the war period with its intensification effort on special projects such as direction finding, communication with aircraft, etc. Out of this period I select only two items for inclusion in the chart :—

1916 The French R valve due to Peri, Ferrié and others.

1918 The Levy superheterodyne patent.

This immensely important superheterodyne principle was also worked on independently by Armstrong in U.S.A.

The next phase in radio is that of broadcasting commencing in 1920 with station KDKA. In the same year the Marconi Company obtained an experimental licence to broadcast from Chelmsford ; not to be outdone in showmanship by De Forest, on June 15th, 1920, the Marconi Company staged a broadcast of Melba, for which she received a fee of 1,000 gns. In 1921 they obtained an M.W. licence for Writtle (200 watts). In 1922 the British Broadcasting Company was formed.

From this time on the broadcasting side of radio began to become the major industry. Although progress was immense, much of it was by way of cost reduction by simplification and standardization, and few items of basically new character suggest themselves for inclusion in the chart.

One may record the following :—

1922 Thoriated tungsten (dull emitter) valves.

1926 The screen grid valve.

1926 Moving-coil loudspeaker.

1927 Indirectly heated cathode valves.

1929 The pentode.

Other events not directly concerned with broadcasting were :—

1922 Housekeeper metal-glass seals.

1923 Beam transmission, Marconi-Franklin.

1925 Cady, the quartz crystal. Poldhu.

1926 Commercial transatlantic telephony.

1927 First International Telecommunications Union, Washington.

A word should now be said on an important aspect of radio propagation, the ionosphere. Although a rigorous (Maxwellian) solution of wave propagation over a sphere appeared only with Watson in 1919, it was already evident that the distances achieved by Marconi and successors needed explaining. Kenelly (1902) and Heaviside (1903) postulated an ionized layer in the upper atmosphere. In 1917, Eckersley made indirect measurements on the ionosphere (published 1921); and in 1926, Appleton and Barnet, using a frequency-modulation method, and in 1927, Breit and Tuve using a pulse method, made direct measurements on it.

Watson's analysis, ignoring the upper layer, showed that only the long waves were capable of useful diffraction round the earth's curvature. This point of view brought about the assignment of the shorter waves to the amateurs who, not having read Watson, proceeded (1923) to get them to work and soon established communication records which soon out-distanced the long waves very easily.*

The 1933 International Telecommunications Union, Madrid, records the assignment of the short-wave broadcast bands.

We turn now to the essentially related subject of Television. This appears to date back to 1884 with Nipkow who conceived the basic process of scanning. The chart shows that this was a

* In this connection the author has noticed the following remark displayed on the wall of a Government Research Establishment :—

"It has recently been established by aero-dynamic theory that the humble-bee is incapable of flight as it does not satisfy conditions of aero-dynamic stability. However the humble-bee does not know this and continues to fly just the same."

conception which was ahead of its time, since none of the necessary associated technique was available.

In photo electricity we record :—

1882 Hertz.
1884 Elster and Geitel.
1905 Einstein.
1928 Caesium photo-cathode.

On cathode ray tubes we record :—

1898 Braun.
1922 J. B. Johnson.
1928 Von Ardenne.
1933 High vacuum CR.O. tubes.

In 1926, Baird commenced his series of television experiments in England. In 1934, Zworykin published information on his "iconoscope," an invention conceived in 1924. Parallel work in this country led to the Emitron.

In 1936, the B.B.C. established the first serious television service in the world. This service is still unexcelled. The work underlying this system, technically brilliant and much ahead of its time, is largely associated with the name of Blumlein.

In this connection, I would particularly mention the name of a Past-President and one of the most outstanding figures in the history of the Institution—Sir Louis Sterling—whose control of policy did so much toward developing television service in this country.

Descendants of the iconoscope family are :—

1937 The Image iconoscope (Super Emitron).
1939 The orthicon (C.P.S. Emitron).
1946 The image orthicon.

In 1935, R. A. Watson-Watt invented radar. A statement of this kind is usually considered controversial. It is true that most of the essentials of the radar process are found in Breit and Tuve and that there are various relevant publications of much earlier date including one by Marconi in 1922,* and also a much later Italian paper which proved that the radar process could not work.

It remained, however, to Watson-Watt to conceive the radar process as we know it, with special regard to its military application, to calculate the possibilities and organize the work which culminated in the demonstrations of Orford Ness, and thereafter to convince the authorities of the necessity of the big-scale development, which he himself initially organized, and which assumed such vital significance throughout the war.

Thus, in addition to the normal intensification of radio effort involved by war, we had this new and special branch, radar. What emerges from all this effort? These results are still too close to our view to allow a proper perspective, but the following seem to stand out :—

The opening up of the microwave bands.
Various systems of radio navigation.
Radar processes for military and civil application.

Associated with these on the tube side we record :—

1939 The Klystron (Varian brothers).
1942 The cavity magnetron.

There also appeared a very great number of new and advanced tube types, together with an immense amount of ancillary circuit technique, not excluding the "invention of the diode by F. C. Williams."

The fourth line of the chart expresses some statistical data on the development of the broadcast industry, with which is included television.

As a convenient yardstick on sound broadcasting, curve (1) gives the number of wireless receiving licences in Great Britain. Curve (3), plotted on 100 times the scale, shows corre-

* Extract from an address by G. Marconi to the Joint Meeting of the Institute of Radio Engineers and the American Institute of Electrical Engineers on June 20th, 1922.

"As was first shown by Hertz, electric waves can be completely reflected by conducting bodies. In some of my tests I have noticed the effects of reflection and deflection of these waves by metallic objects miles away.

"It seems to me that it should be possible to design apparatus by means of which a ship could radiate or project a divergent beam of these rays in any desired direction, which rays, if coming across a metallic object, such as another steamer or ship, would be reflected back to a receiver screened from the local transmitter on the sending ship, and thereby immediately reveal the presence and bearing of the other ship in fog or thick weather.

"One further great advantage of such an arrangement would be that it would be able to give warning of the presence and bearing of ships, even should these ships be unprovided with any kind of radio.

"I have brought these results and ideas to your notice as I feel—and perhaps you will agree with me—that the study of short electric waves, although sadly neglected practically all through the history of wireless, is still likely to develop in many unexpected directions, and open up new fields of profitable research."

sponding data for British television. Curve (2), on ten times the original scale, gives equivalent data for American television.

I consider that deductions from the slope and curvature of these curves must be made with the utmost reserve. Nevertheless the curves do indicate very much what one expects to happen, namely, an approaching saturation in sound broadcasting with an accelerating spread of television. Saturation in the case of sound broadcasting can be expected, not from the curvature of curve (1), but from the fact that the present figure of 11·2 million licences represents a " licence per family " figure of approximately 75 per cent. A similar saturation tendency was evident in U.S.A. before the war. Counter measures were push-buttons and F.M., neither entirely successful in this particular role.

In reflecting upon the possibilities of an ultimate change over from aural to visual broadcasting, one may usefully consider history in certain related arts. Thus, again referring to the chart, we note Edison's invention of the phonograph, 1889, followed by his invention of the " Kinetoscope" in 1891. In looking up these dates I was particularly interested to read the reason for Edison's invention of the Kinetoscope; it was that he wanted a visual accompaniment to his improved phonograph. The outcome, however, was the quite separate cinema industry.

In 1926 appeared the Vitaphone, the first reasonably successful of a number of attempts to bring sound to the film. Curve (4) indicates the sequel ; over two years at the most the intrusion of sound was complete.

In weighing the possibilities of a corresponding event in broadcasting, one finds that many of the factors are similar in kind but different in quantitative aspects, being in this respect less favourable to television. However, there are certain additional factors in favour of television, amongst which is its educational and political significance.

When radio broadcasting began, many saw in it the possible role of an instrument of international peace. We have certainly been grievously disappointed ; from propaganda to radar, radio has been outstanding as an instrument of war. Can television do any better? This is a matter of speculation. On the one hand the visual art has basically a better chance of international dissemination than has the aural one, where questions of language are predominant. On the other hand, international distribution of television presents serious problems, not least among these being international standardization. If television could, in time, make even the slightest contribution towards the avoidance of war, it would not only serve the interests of mankind, but would be in a fair way to solve its own economic difficulties. This may be expressed by the fact that the annual cost of the B.B.C. television service is equivalent to about two hours of war.

My belief is that, having regard to the evidence of the past and the implications for the future, the introduction of television into broadcasting on an overwhelming scale is an ultimate inevitability, and is the future and hope of the radio industry.

I believe that this review will be of interest in planning ahead. Apart from that, one of my objects has been to remind us of the priceless heritage of scientific achievement that lies behind our art and industry. It is, in my opinion, one of the finest records of human achievement and an outstanding testimony to the glory of God. The duty of this Institution is to be worthy of that possession.

NOTICES

Obituary

Council regrets to record the death of Mr. E. A. Kirtley (Associate Member).

Mr. Kirtley, who joined the Institution in 1934, was engaged in teaching at the Bournemouth Technical College and will be remembered by many members who attended the 1947 Convention.

Huddersfield Site for N. England Television

The B.B.C. has recently been granted permission to use a site on Dean Head Hill, in the Digley Reservoir catchment area, for the erection of a television transmitting station. The site is over 1,500-ft. high, and with the addition of a 500-ft. mast a large area should be covered, including parts of the Lancashire and Yorkshire coasts.

R.C.M.F. Components Exhibition

The Radio Component Manufacturers' Federation announce officially that their sixth annual private exhibition of British components, valves and test gear for the radio, television, electronic and telecommunication industries will be held in the Great Hall, Grosvenor House, Park Lane, London, W.1, during the period Tuesday, March 1st to Thursday, March 3rd, 1949, daily, from 10 a.m. to 6 p.m.

Approximately one hundred exhibitors will be participating in the display, which is promoted to acquaint radio and electronic manufacturers, engineers and research technicians with the latest advances in design and development.

Admission will be by invitation only, as in previous years. Applications for tickets and further information can be obtained from the secretary of the Federation at 22 Surrey Street, Strand, London, W.C.2.

Yorkshire Section Meeting

A meeting will be held at the Y.W.C.A., Cookridge Street, Leeds, at 7 p.m. on January 13th, when a paper will be given by Mr. L. Grinstead (Member) and Dr. H. P. Zade on "Radio Frequency Welding."

Second Annual Amateur Radio Exhibition

The Second Annual Amateur Radio Exhibition, organized by the Incorporated Radio Society of Great Britain, was held at the Royal Hotel, Woburn Place, London, from November 17th to 20th. After an informal luncheon, at which the Institution was represented, the exhibition was opened by Dr. R. L. Smith-Rose, Director of Radio Research, D.S.I.R.

The 27 stands in the exhibition carried a large variety of apparatus and literature of interest to the radio amateur. Greater interest was evident in V.H.F. equipment, and there was an increase in the number of prefabricated units, such as coil-turrets, for inclusion in complete apparatus. It is interesting to note that since the similar exhibition a year ago no fewer than four new communication receivers have made their appearance on the home market.

R.T.E.B. Radio Servicing Certificate Examination

The closing date for the May, 1949, examination is February 1st. The written examination will be held on May 3rd and 5th at centres throughout the British Isles. The practical examination will be held on May 21st, at a smaller number of centres, at London, Birmingham, Manchester, Glasgow and Bristol, and at such additional centres as are warranted by the number of applicants.

Journal of the Institution

Increased paper allocation now enables the Institution to revert to monthly publications of the Journal. In future, members should receive their Journal about the 10th of each month.

Although the number of pages per issue will still have to be severely restricted, the increased allocation will facilitate the publication of a number of papers which have been held up for some time. In the case of papers submitted and approved for reading before section(s), publication should now be arranged within two or three months of the date of reading the paper ; papers accepted for publication only should be published within two months of final acceptance by the Papers Committee.

Members are therefore urged to offer papers at least five months before they are due to be read before any section. In this connection it may be pointed out that the Papers Committee is now making arrangements for the 1949/50 session of meetings throughout the sections. Offers of papers, accompanied by the synopsis at least, would therefore be very welcome.

THE PHYSICAL APPLICATIONS OF MICROWAVES*

by

J. B. Birks, B.A., Lecturer in Natural Philosophy, University of Glasgow

(Read before the Merseyside Section on March 31st, the Scottish Section on April 21st and the London Section on December 16th.)

I. Introduction

A major part of the physicist's knowledge of the structure of matter has been derived from the study of the interactions between electromagnetic radiation and matter. These investigations have ranged from one extreme end of the electromagnetic spectrum to the other—from long radio waves of several kilometres wavelength to the high-energy gamma rays, produced by disintegration of the atomic nucleus, having wavelengths of 10^{-11} cm. and less. In the present Paper we shall be concerned only with the high-frequency radio waves, known as microwaves, whose wavelengths lie in the region from 50 cm. down to 5 mm.

Prior to 1939 comparatively little work had been done on the properties of materials at centimetre wavelengths, due largely to the primitive nature of the existing experimental techniques, though limited use was made of such sources as the spark-gap generator, the split-anode magnetron and the Barkhausen-Kurz oscillator. During the period from 1939 onwards very rapid advances took place in the whole microwave field, due to the concentrated effort of teams of scientists engaged on the development of centimetric radar. Vastly improved methods for the generation, transmission, reception, detection and measurement of microwaves were devised for use in such radar systems as H_2S, ASV, AI, Oboe, GL and GCI. The technical applications of microwaves for radar have been described elsewhere[1] and will not be reconsidered here. Apart from radar however, these major wartime advances in experimental technique and apparatus have provided the physicist with important new research tools, and have encouraged fresh investigations of the properties of matter in the centimetre wave region. The aim of this paper is to review the progress made in such physical applications of microwaves.

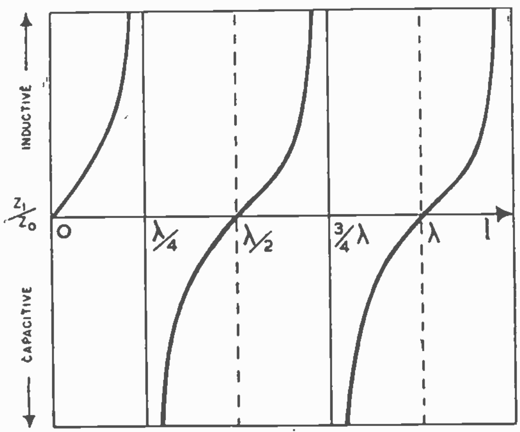


Fig. 1.—Input impedance of short-circuited transmission line. The arrow represents the length l.

2. Principles of Microwave Technique

Microwave experimental techniques differ fundamentally from those used at longer radio wavelengths. These differences arise from the fact that micro-wavelengths are of the same order of magnitude as the dimensions of the circuit components.

2.1. *Circuits*

Lumped circuits employing conventional coils and condensers are quite unsuitable at microwaves because of excessive radiation losses. Instead non-radiative circuits with distributed constants have to be used. These take the form of coaxial transmission lines, waveguides and cavity resonators.

For example, a short-circuited lossless transmission line of length l, characteristic impedance Z_0, functions as a reactance, presenting at its input a purely imaginary impedance of

$$Z_1 = jZ_0 \tan(2\pi l/\lambda) \quad \text{.................. (1)}$$

at a line wavelength of λ. The nature and magnitude of this reactance depend on both the dimensions and the operative wavelength (Fig. 1). In the range $l = 0$ to $\lambda/4$, Z_1 is positive, i.e., inductive, increasing from 0 (short-circuit) to $+\infty$ (open-circuit), while from $l = \lambda/4$ to $\lambda/2$, Z_1 is negative, i.e., capacitive, decreasing from $-\infty$ down to 0 again. The

* U.D.C. No. 621.3.029.64 : 53.
Manuscript received February 1948.

identical inductive-capacitive sequence is repeated after each value of $l = n\lambda/2$ (n integral), the multiple $\lambda/2$ line functioning as a 1 : 1 transformer.

Just as in normal radio practice, an inductive reactance may be linked with an equal capacitive reactance to form a resonant LC circuit, so transmission line reactances may be combined to form closed resonant circuits. Thus, if a shorted line, length $\lambda/2 - l$, is joined to a similar line, length l, the two sections form a closed half wavelength line. The input impedance of the first section, is from (1)

$$= j\,Z_o \tan 2\pi(\lambda/2 - l)/\lambda$$
$$= -\,j Z_o \tan 2\pi l/\lambda$$

and is hence equal and opposite to that of the second section, so that the combined circuit resonates at a wavelength of λ. The efficiency or Q-factor of the resonance is determined by the small residual resistance of the circuit.

Cavity resonators and transmission lines will be considered in more detail later.

2.2. *Oscillators*

The period of microwave oscillations is comparable with the transit-time of the electron beam in a normal radio valve, and different methods have therefore to be used to generate microwaves. Although by reducing valve size and electrode clearances to a minimum, triode oscillators have been constructed for use down to less than 20 cm., the shorter microwaves can only be adequately produced by special types of valves, such as the klystron and cavity magnetron. In microwave oscillators the transit-time of the electrons is designed, by means of the valve geometry and the applied electric and/or magnetic fields, to synchronize with the natural period of a cavity resonator which forms an integral part of the valve, and thus stimulate electromagnetic oscillations in the resonator.[2]

In the reflex klystron (Fig. 2) which is most widely used as a microwave source, the electron beam from the cathode K passes through the gap G in the centre of a resonant cavity, and is velocity-modulated by the oscillating cavity field. The electrons emerge into the drift-space D, travel towards the reflector R, where they are repelled by a negative field and are returned towards G. During the transit-time through the drift-space, the electrons " bunch " due to their velocity-modulation, and they arrive at G at the right instant to stimulate the cavity oscillations. Thus energy is transferred from the electron beam into electromagnetic oscillations in the cavity. Power is taken from the oscillator by means of the transmission line T coupled to the resonator.

The frequency of the oscillations is varied mechanically by changing the shape of the resonator by pressure, or by screwing in the tuning plunger P. Alternatively, the frequency may be swept over a small range by varying the voltage applied to the

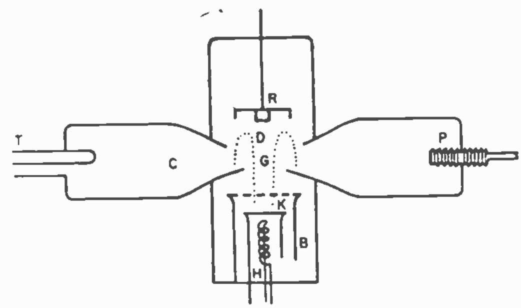


Fig. 2.—Reflex klystron oscillator. H—*heater ;* B—*beam forming electrode ;* C—*cavity ;* P—*tuning plunger ;* T—*output transmission line.*

reflector R. The amplitude of the oscillations can be modulated by means of a modulation voltage applied to the grid of the reflex klystron.

2.3. *Detectors*

Diode detectors have been designed for the longer centimetre wavelengths, but the general technique is to use crystal rectifiers of the silicon-tungsten type. When adequate power is available, less sensitive detectors, such as the bolometer or thermocouple, may be used. These thermal detectors are suitable for absolute power measurements since they may be calibrated with direct current.

Two alternative methods of crystal detection are generally employed.

(*a*) *Direct detection.*—Below a few microamps, the rectified crystal current is proportional to the square of the H.F. voltage detected, and by working in this " square-law " region accurate measurements may be made with a sensitive galvanometer. The crystal noise is less at audio-frequencies than at D.C., so when extreme sensitivity is required a low-frequency amplitude modulation is applied to the oscillator grid, and the rectified crystal current amplified by an audio-amplifier.

(*b*) *Superheterodyne detection.*—Here the signal to be measured is mixed in the crystal rectifier with a steady signal from a klystron local oscillator, and the resultant beat frequency current amplified by means of an I.F. amplifier usually operating in the 10-60 Mc/s region. In this method, which is similar to that used in radar, the I.F. crystal current is directly proportional to the H.F. voltage detected.

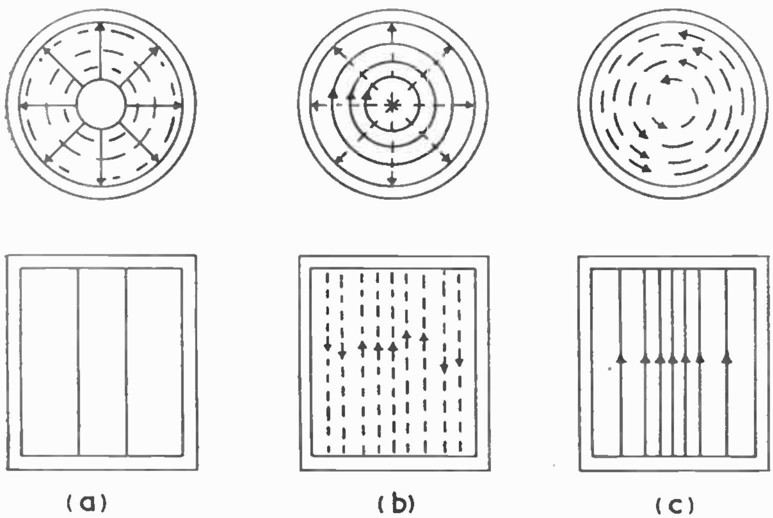


Fig. 3.—Cavity resonator fields. Electric field ——— Magnetic field - - - - -. (a) Coaxial half-wave resonator ; (b) H_{011} resonator ; (c) E_{010} resonator.

3. The Expression of the Electromagnetic Properties of Materials

The electric and magnetic properties of a material are most simply expressed in terms of two parameters, the permittivity ε and the permeability μ, either or both of which may be complex. The values of ε and μ will be taken relative to the constants for free-space, as in the C.G.S. system. Various alternative methods of expressing these parameters occur in practice, and these will be briefly considered.

3.1. *Electrical Properties*

The complex permittivity ε may be written

$$\varepsilon = \varepsilon' - j\varepsilon'' \quad \text{.................(2)}$$

where ε', ε'' are called the *dielectric constant* and *dielectric loss factor* respectively. Dielectric loss is expressed in terms of the *dielectric loss tangent*

$$\tan \delta_\varepsilon = \varepsilon''/\varepsilon' \quad \text{...................(3)}$$

or the *equivalent conductivity*

$$\sigma = \omega\varepsilon'' \quad \text{...................(4)}$$

where ω is the angular frequency. (This second mode of expression, though common, is most misleading, since H.F. dielectric loss arises from causes other than conductivity. By analogy, H.F. magnetic loss would be expressed as equivalent to a fictitious magnetic conductivity. It originates in the conventional formulation of Maxwell's equations in terms of three *real* parameters ε' μ', and σ.)

The optical properties of a material are stated in terms of a *complex index of refraction* $\bar{n}$, which for non-magnetic media is given by

$$\varepsilon^{\frac{1}{2}} = \bar{n} = n'(1 - jk) \quad \text{.....(5)}$$

where n' is *the refractive index*, k the *index* of *absorption*. Hence

$$\varepsilon' = n'^2(1 - k^2) \quad \text{........(6)}$$

$$\varepsilon'' = 2n'^2k \quad \text{..............(7)}$$

Since the microwave region forms the bridge between short-wave radio and long-wave infra-red both electrical and optical terminology are commonly used.

3.2. *Magnetic Properties*

At high frequencies ferromagnetic and paramagnetic materials show magnetic dispersion and absorption, which may be described in terms of a *complex permeability*

$$\mu = \mu' - j\mu'' \quad \text{................(8)}$$

where μ', μ'' may be called the *real permeability* and *magnetic loss factor* respectively. By analogy we define the *magnetic loss tangent*

$$\tan \delta_\mu = \mu''/\mu' \quad \text{.................(9)}$$

Alternatively for paramagnetics, the properties can be expressed in terms of a *complex susceptibility*

$$\chi = \chi' - j\chi'' \quad \text{.............(10)}$$

where

$$\mu = 1 + 4\pi\chi \quad \text{.............(11)}$$

and hence

$$\mu' = 1 + 4\pi\chi' \quad \text{.............(12)}$$

$$\mu'' = 4\pi\chi'' \quad \text{.................(13)}$$

3.3. *Electromagnetic Properties*

In practice at micro-wavelengths the measurements are not made in a purely electric or magnetic field, but in a combined electromagnetic field. For a plane electromagnetic wave, the properties of a medium are expressible in terms of two parameters, the intrinsic impedance, relative to free space, and the propagation coefficient. The *relative intrinsic impedance*

$$Z = (\mu/\varepsilon)^{\frac{1}{2}} \qquad \ldots\ldots\ldots\ldots\ldots\ldots (14)$$

in conjunction with Fresnel's equations, determines the reflection and transmission coefficients of the wave at the boundary of the medium. *The propagation coefficient*

$$\gamma = j\omega(\mu\varepsilon)^{\frac{1}{2}} \qquad \ldots\ldots\ldots\ldots\ldots (15)$$

may be written

$$\gamma = \alpha - j\beta \qquad \ldots\ldots\ldots\ldots\ldots\ldots (16)$$

where α is the *attenuation coefficient* and β is the *phase constant* of the wave propagated through the medium.

4. Experimental Methods

4.1. *Cavity Resonator Methods*

Cavity resonators, which are the microwave analogues of resonant LC circuits, have been widely used for the measurement of low-loss dielectric materials at centimetre wavelengths. They usually consist of closed sections of coaxial transmission lines or waveguides, oscillating in one of the many possible characteristic modes of resonance. In the principal transmission mode of a coaxial line, both electric and magnetic field vectors are purely transverse to the direction of propagation. The simplest form of coaxial resonator thus consists of a closed half-wave section of coaxial line having the field configuration shown in Fig. 3*a*.

In waveguides, on the other hand, no purely transverse modes exist, and the various transmission modes are classified into two main types, E and H, depending on whether there is a longitudinal component of the electric or magnetic field.[3] Each waveguide mode is characterized by a critical wavelength λ_c, above which the mode cannot be propagated. The value of this critical wavelength depends on the shape and dimensions of the guide, and on the order of the mode. For example, the critical wavelength of the simplest mode in a rectangular guide, the H_{01} type, is equal to $2b$, where b is the tube width, while for the H_{01} mode in a circular guide, $\lambda_c = 1{\cdot}64a$, where a is the tube radius. λ_c determines the wavelength λ_g in the empty guide, according to the relation

$$1/\lambda_g = \left[1/\lambda^2 - 1/\lambda_c^2\right]^{\frac{1}{2}} \qquad \ldots\ldots\ldots\ldots (17)$$

Normally the simplest transmission modes, having the longest critical wavelengths, are used to avoid a multiplicity of possible modes. Resonators may be constructed from any shape of guide, but the highest Q-factors are obtained with spherical or cylindrical resonators, and because of their simple construction, the latter are almost exclusively used. By analogy with the half-wave coaxial resonator, we derive the H_{011} cylindrical resonator consisting of a closed $\lambda_g/2$ section of H_{01} circular guide (Fig. 3*b*). A section $n\lambda_g/2$ in length, n integral, gives the H_{01n} mode. Similar resonator modes can be derived from $n\lambda_g/2$ lengths of the other waveguide modes.

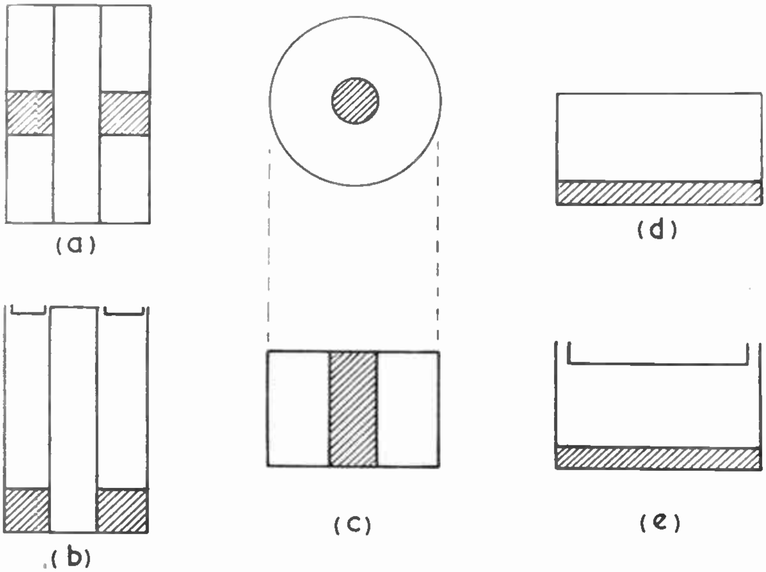


Fig. 4.—Location of dielectric specimens in cavity resonators ; (a) Coaxial—Variable frequency ; (b) Coaxial—Variable length ; (c) E_{010}—Variable frequency ; (d) H_{01n}—Variable frequency ; (e) H_{01n}—Variable length.

Some cylindrical resonator modes, however, do not correspond to sectional waveguide modes, since the addition of terminating end-plates introduces new boundary conditions, for which additional solutions of the electromagnetic wave equations are possible. The simplest of these is the E_{010} mode in which the electric field is purely longitudinal, and the magnetic field is purely circumferential (Fig. 3*c*). The resonance condition of this mode is governed, not by the length of the cavity, but by its radius *a*, resonance occurring at a free-space wavelength of 2·61 *a*.

The Q-factor of a cavity resonator is extremely high and this property makes it a very sensitive instrument for the measurement of dielectric loss. Comparative figures computed for the Q-factors of typical 10-cm. resonators[4, 5] are :

Coaxial, 6,100 ; E_{010}, 17,500 ; H_{011}, 47,500.

These values may be compared with the much lower figure of 500 obtained with a typical lumped resonant circuit at its shortest wavelength of operation (300 cm.). The resonator type to be used at a particular wavelength is chosen primarily for convenience of dimensions, which are restricted by the resonance conditions, and by the critical wavelengths of the dominant mode and any possible higher order ones. Coaxial resonators have been used from 50 cm. to 10 cm., E_{010} resonators from 50 cm. to 3 cm. and the H_{01n} type from 3 cm. downwards.

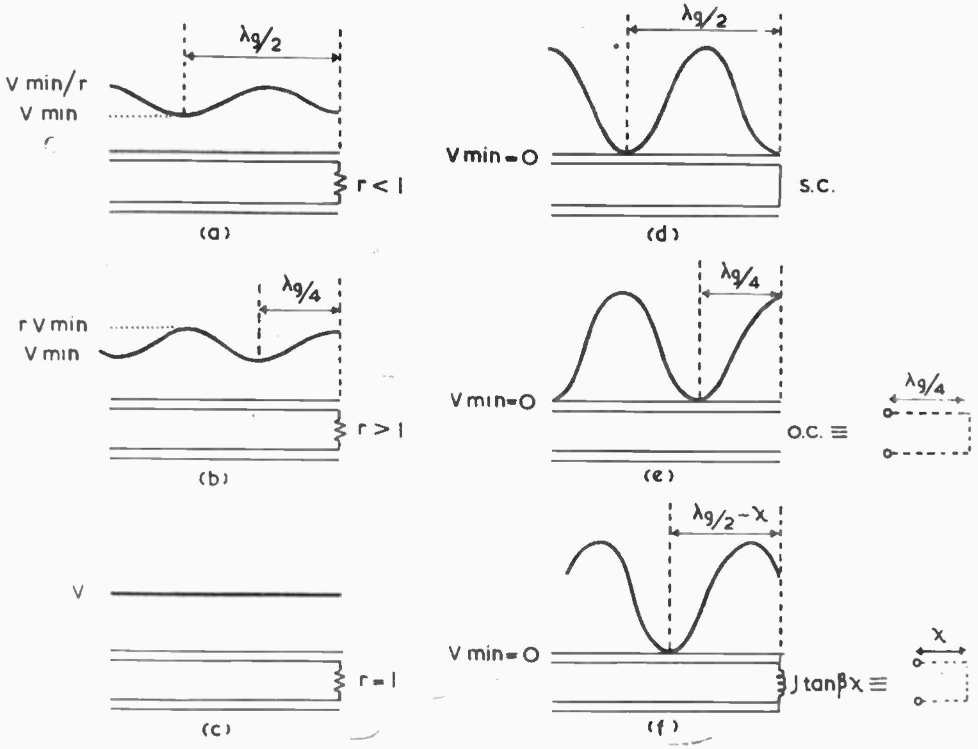


Fig. 5.—Voltage standing waves on terminated lines. (a) (b) (c) Resistive terminations ; (d) (e) (f) Reactive terminations.

The dielectric properties of a material may be measured by introducing it into a resonator and observing the change produced in the resonance condition and the Q-factor. For example, if an empty coaxial resonator, length l_o, resonant frequency ν_o, of Q-factor Q_o, is filled with a low-loss dielectric, resonance can be restored either by reducing the length to l, or by increasing the excitation frequency to ν. The dielectric constant of the material is then given by

$$\varepsilon' = l_0^2/l^2 \quad \ldots\ldots (18)$$

or by

$$\varepsilon' = \nu^2/\nu_0^2 \quad \ldots\ldots (19)$$

The dielectric loss tangent is given by

$$\tan \delta_\varepsilon = 1/Q - 1/Q_0 \quad \ldots\ldots (20)$$

where Q is the reduced Q-factor of the system, determined from the width of the resonance curve, obtained by incremental frequency or length variations. Similar, though more complicated relationships involving λ_c, are valid for the H_{01n} and E_{010} filled resonators[4] except that in the E_{010} case, as the resonance condition is independent of the length, only the frequency-variation method analogous to (19) is practicable.

The Q-factor of a cavity is so sensitive to the introduction of a small loss that only for gases is it necessary or desirable to fill the resonator completely with the material under investigation. For solids and liquids measurements are made on small cylindrical or disc-shaped samples suitably located in the resonator field (Fig. 4). The effect of the thin containing vessels used for liquids may be eliminated by preliminary measurements. The dielectric properties are evaluated from similar relations to (18), (19) and (20) derived from the field and boundary conditions within the resonator.[4,6,10]

4.2. *Transmission Line Methods*

Coaxial transmission lines and rectangular H_{01} waveguides have also been widely used for the measurement of the microwave properties of materials. They have the advantage over cavity resonators of not being restricted by resonance conditions to a narrow frequency band. They can also be used for a much wider range of materials, including high-permittivity ceramics, lossy dielectrics and semi-conducting ferromagnetics, for which cavity resonator methods are unsuitable. For measurements on small samples of very low loss materials, transmission lines are rather less sensitive than resonators because of their reduced Q-factor. This limitation disappears, however, if there are no restrictions on the volume of the material, and in fact the most sensitive measurements in microwave spectroscopy have been made on gases at very low pressures in waveguide absorption cells.

Transmission line methods may be divided into two types: (*a*) impedance measurements; (*b*) propagation measurements. The first method is used primarily for solids, and the second method for gases and liquids.

4.2.1. *Impedance Measurements*

The basis of such methods is the determination of the input impedance of a line section containing the material under investigation, from measurements of the standing-wave pattern on a transmission line terminated by the section. The standing-wave pattern on a terminated line may be defined and measured in magnitude by the voltage standing-wave ratio n, which is the ratio of the voltage maximum to minimum, and in phase by the electrical distance βl of the first voltage minimum from the termination. Figure 5 illustrates the voltage standing wave patterns produced by various impedance terminations on a loss-less line of phase velocity $\beta = 2\pi/\lambda_g$, where λ_g is the line wavelength. As in experimental practice, all impedance values are taken relative to the characteristic impedance of the empty transmission line.

If the line is terminated in a *pure resistance r*, the *s.w.r.* $n = 1/r$ or r, depending on whether $r <$ or > 1 and the phase βl is equal to 0 or $\pi/2$ respectively (Fig. 5*a*, *b*). For the particular case when the line is terminated in its own *characteristic impedance*, it is matched, i.e., $r = n = 1$. (Fig. 5*c*). The limiting cases $r = 0$ and $r = \infty$ give the *short-circuit* and *open-circuit* conditions respectively (Fig. 5*d*, *e*). A *pure reactance* $j \tan \beta x$ is equivalent from (1) to a short-circuited line, electrical length βx, and hence the standing-wave pattern produced is similar to the S.C. case, but shifted in phase towards the termination by βx (Fig. 5*f*). Hence we see that a pure resistance can be determined from n, while a pure reactance can be determined from βl. A *complex impedance* Z_l depends on both n and βl, and is given by

$$Z_l = \frac{1/n - j \tan \beta l}{1 - j(\tan \beta l)/n} \quad \ldots\ldots (21)$$

Thus the relative input impedance of any terminating line section may be determined from standing-wave measurements. Such measurements are usually taken by means of a tuned crystal detector, which can be moved along the transmission line. A probe from the detector projects a small distance into the line through a narrow slot cut parallel to its axis, and the voltage induced in the probe is rectified by the crystal.[13]

In a coaxial line, since the wave mode is purely transverse, the characteristic impedance Z_o relative to that of an empty line, of a section filled with material of constants ε, μ, is equal to the relative intrinsic impedance of the material

$$Z_o = Z = (\mu/\varepsilon)^{\frac{1}{2}} \quad \text{(14a)}$$

and the propagation coefficient γ_o in the line section is equal to that for a plane wave in the unbounded material,

$$\gamma_o = \gamma = j\omega(\mu\varepsilon)^{\frac{1}{2}} \quad \text{(15a)}$$

For the H_{01} waveguide, similar but more complex expressions for Z_o and γ_o are valid, involving μ, ε and the experimentally known quantities λ and λ_c.[13]

The simplest method of determining Z_o is to measure the input impedance of Z_{inf} of a line-section of semi-infinite length (i.e., of sufficient length for the reflection from the back face to be negligible) for which

$$Z_{inf} = Z_o \quad \text{(22)}$$

This method is only suitable for very lossy materials, and a line-section terminated in a short circuit is more generally employed.[16] For such a section of length d, the input impedance

$$Z_{sc} = Z_o \tanh \gamma_o d \quad \text{(23)}$$

Alternatively, if the section is terminated in an empty closed quarter-wave line, which is equivalent to an open-circuit

$$Z_{oc} = Z_o \coth \gamma_o d \quad \text{(24)}$$

For dielectric materials, μ = 1, and hence the input impedance is in each case a function involving only ε as an unknown factor. For ferromagnetic materials, where both μ and ε are required, a single input impedance measurement is insufficient. The method used by the author[12,13] involves the measurement of both Z_{sc} and Z_{oc} for the same sample. Then from (23) and (24)

$$Z_o = (Z_{sc}Z_{oc})^{\frac{1}{2}} \quad \text{(25)}$$

$$\tanh \gamma_o d = (Z_{sc}/Z_{oc})^{\frac{1}{2}} \quad \text{(26)}$$

from which both Z_o and γ_o, and hence ε and μ, may be evaluated. This dual impedance method is also applicable to non-magnetic materials.

4.2.2. *Propagation Measurements*

Other transmission line methods, which do not involve impedance determinations, have been used for absorption measurements on gases and liquids. The line, usually an H_{01} waveguide, is filled with the material under investigation and direct measurements are made of the attenuation of the filled guide. The method is particularly suited to the microwave absorption spectroscopy of gases at low pressures.

Such measurements give the loss, but not the dielectric constant of the material, but usually the latter is unimportant. An extension of the absorption method[7] which overcomes this limitation, uses two H_{01} waveguide cells of different guide widths, so that they have different propagation coefficients. Both ε′ and ε″ may be derived from the two absorption measurements.

5. The Dielectric Properties of Solids and Liquids

5.1. *Origin of Dielectric Properties*

When an electric field is applied to a dielectric, the positive and negative charges in its atoms and molecules are displaced in opposite directions, producing a field in opposition to the applied field. The magnitude of this field, or polarization, determines the permittivity of the material. The polarization takes a finite time, known as the relaxation time τ, in which to form. The relaxation frequency of a polarization is defined by

$$\nu_R = 1/2\pi\tau \quad \text{(27)}$$

With alternating fields of frequencies much less than ν_R the polarization is able to form completely before the field is reversed, but as the frequency is increased through ν_R the polarization begins to lag, and consequently the dielectric constant decreases. At the same time, the dielectric loss increases to a maximum, and decreases as ν_R is exceeded. Three main types of polarization are possible (a fourth type, interfacial polarization, which is restricted to heterogeneous materials at low frequencies, need not concern us here) and every dielectric exhibits one or more in its frequency spectrum (Fig. 6).

The different types of polarization are distinguished by their relaxation frequencies and by their sources of origin within the material. *Electronic polarization* is due to the displacement of locally-bound electrons within the atom, and has characteristic frequencies in the ultra-violet and optical region. It contributes a component of permittivity ε_E, equal to the square of the optical refractive index. *Atomic polarization*, which is due to the displacement of atoms within the molecule, has resonance frequencies in the infra-red, and is responsible for a permittivity component ε_A, which is usually small.

In addition to these induced polarizations, in many molecules there exist permanent electric dipoles, due to assymmetry of charge distribution, produced by differences in electron affinity of the constituent atoms. Such permanent dipoles give rise to *dipole polarization* by orientation of the molecules in the applied field. The molecules of gases, being free to rotate, follow the field more or less instantaneously, and the oscillation frequencies occur in the infra-red, extending into the microwave region. In solids, however, the rotational freedom is greatly restricted by the interaction of adjacent molecules, and the orientation phenomenon degenerates into aperiodic oscillation under high friction, occurring at radio frequencies. Similar damping occurs in liquids, but due to the reduced intermolecular forces, the relaxation frequencies are higher, often within the microwave region. Dipole polarization contributes a component ε_P to the permittivity, and is responsible for most of the H.F. dielectric loss in a polar material. Thus the static dielectric constant is in general given by

$$\varepsilon_0 = \varepsilon_E + \varepsilon_A + \varepsilon_P \quad \ldots\ldots(27a)$$

For non-polar materials, $\varepsilon_P = 0$, and hence the dielectric constant retains its static value up to microwave frequencies, while the dielectric loss is very small, and attributable to impurities. For polar materials however, as the frequency is increased, the permittivity decreases and becomes complex. According to theory[14] the permittivity at a frequency ν is given by

$$\varepsilon = \varepsilon_E + \varepsilon_A + \frac{\varepsilon_P}{1 + j\nu/\nu_R} \quad \ldots\ldots\ldots(28)$$

for a material having a single dipolar relaxation frequency ν_R. This gives a dielectric spectrum due to dipolar relaxation of the form shown in Fig. 6, with

$$\varepsilon' = \varepsilon_E + \varepsilon_A + \frac{\varepsilon_P}{1 + (\nu/\nu_R)^2} \quad \ldots\ldots(29)$$

$$\varepsilon'' = \frac{\varepsilon_P (\nu/\nu_R)}{1 + (\nu/\nu_R)^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots(30)$$

The dielectric loss factor ε'' has a maximum when $\nu = \nu_R$.

5.2. *Dielectric Properties of Solids*

The microwave properties of solid dielectrics are of great technical importance in the design of centimetre wave equipment, though they are of little direct scientific interest. Non-polar materials, such as polythene and polystyrene, are widely used for microwave insulation, and since dielectric loss may be responsible for nearly half the attenuation in a typical 10-cm. transmission system, the reduction of the loss tangent of polythene from around 0·0005 to 0·0002 represents an increase of up to 30 per cent. in efficiency. A detailed investigation of the factors determining the residual dielectric loss of polythene, including the effect of slight impurities and oxidation due to prolonged heating during processing, has led to an improvement of this order in the final product.[15] The work was carried out at 9-cm. wavelength, using thin cylindrical specimens in an E_{010} cavity resonator.

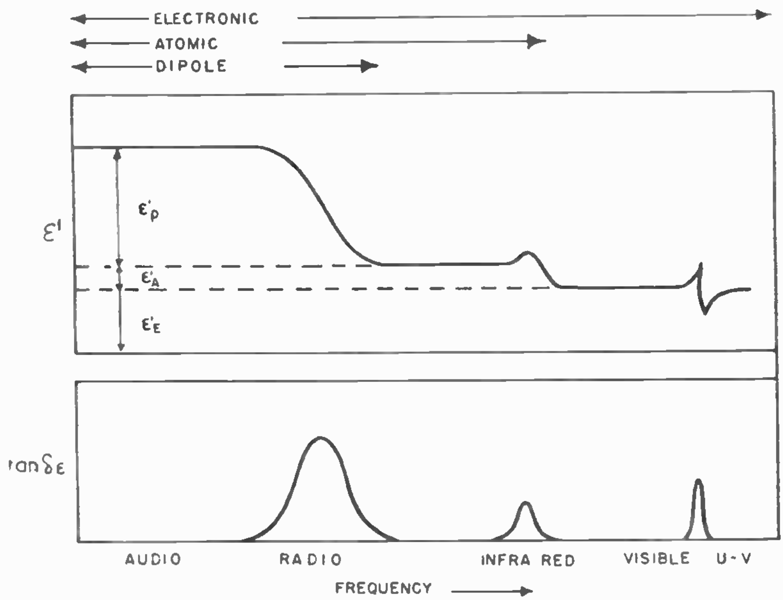


Fig. 6.—General dielectric spectrum.

Measurements on other solid dielectrics have been chiefly of a cataloguing nature. The numerous investigations include :

(*a*) Coaxial line measurements down to 6 cm. on commercial dielectrics.[16]

(*b*) H_{01} waveguide measurements at 9 cm. and 3 cm. on potential antenna housing materials.[11]

(c) Coaxial, E_{010} and H_{01n} resonator measurements at 40 cm., 9 cm., and 3 cm., respectively.[4]

(d) H_{01n} resonator measurements at 1·25 cm.[6]

The dielectric spectra of two typical polar solids are shown in Figs. 7 and 8. The first, one of the vinyl polymers, has a slowly decreasing microwave dielectric constant, associated with a decreasing loss-tangent, due to radio-frequency dipolar relaxation (Fig. 7). The second material, a silicate glass, has a decreasing dielectric constant, and an increasing loss, attributable to atomic polarization effects in the infra-red (Fig. 8). It will be noted by comparison with Fig. 6 that these spectra are much broader and flatter than those predicted by the simple theory (28). This may be accounted for by supposing that the inter-molecular forces in the solid give rise to a broad distribution of relaxation frequencies, instead of the single one assumed in deriving (28).

5.3. *Dielectric Properties of Liquids*

The microwave properties of many polar liquids are of interest, because they have relaxation frequencies in this region. Water, for example, which has a high static permittivity ~80 and an optical one of only 1·8, has a dipolar relaxation around 30,000 Mc/s. Hence the microwave dielectric constant of water falls rapidly with increasing frequency, and its loss rises to a peak near 1 cm. wavelength (Fig. 9). This high dielectric loss affects the design and operation of equipment at the shorter micro-wavelengths. Hermetically sealed waveguides, waterproofed circuits and non-hygroscopic insulation have to be used, while

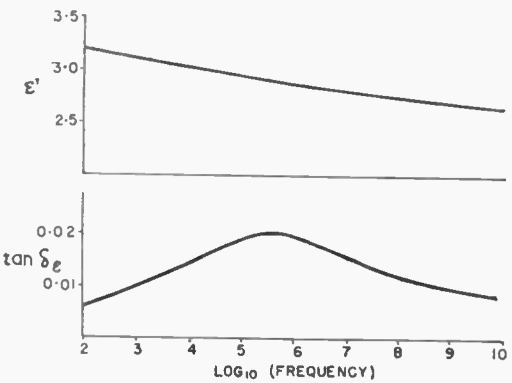


Fig. 7.—Dielectric spectrum of polyvinyl formal.

atmospheric propagation is adversely affected by rain and clouds.

The microwave properties of water have been studied by several methods, including :

(a) Coaxial line impedance measurements.[16]

(b) The double H_{01} absorption cell technique.[7]

(c) An H_{01n} resonator with cylindrical specimens placed axially.[10]

(d) A free space transmission and reflection method.[17]

The experimental results are in fair agreement with the simple theory (28) of a single dipolar relaxation frequency. Since a decrease in temperature reduces the rotational freedom of the

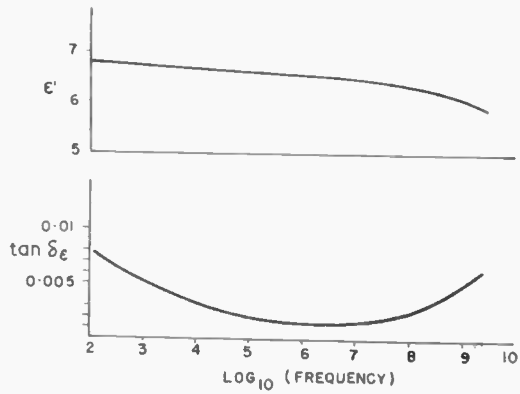


Fig. 8.—Dielectric spectrum of typical lead silicate glass.

dipoles, it also reduces the relaxation frequency (Fig. 9). The transition to the solid state causes an abrupt " freezing-in " of the dipoles, so that the relaxation frequency for ice at −5°C. occurs at only 35 kc/s, a drop of nearly a million times. Ice is thus a low-loss dielectric at microwaves.[9]

Other liquids, whose microwave properties have been investigated, include the alcohols. These form a series—methyl (CH_3OH), ethyl (C_2H_5OH), etc., whose molecules are similar to water (HOH) but which are progressively larger and less polar. This causes a progressive decrease in ν_R with increasing molecular weight, and a similar progressive decrease in ε_0 with decreasing polarity.[17]

Symmetrical molecules, such as benzene and the paraffins, are non-polar and have a low microwave loss tangent.[8]

5.4. *Dielectric Properties of Polar Solutions*

Since the dipole moment of a molecule depends on the relative positions of its component atoms, the measurement of dipole moments is an important

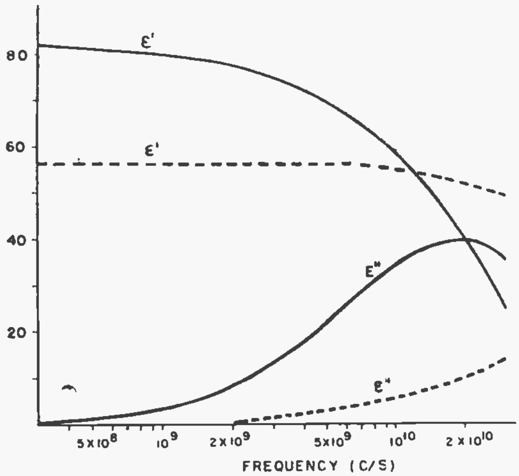


Fig. 9.—Dielectric spectrum of water
Temperature : 20°C ——
85°C - - - -

method for the study of molecular structure. The theory of hindered molecular rotation mentioned in 5.1 forms the basis of such measurements.

The dipolar permittivity component ε_P depends on the square of the dipole moment ρ, and on the reciprocal of the absolute temperature T, i.e.,

$$\varepsilon_P = F(\rho^2/T) \ldots\ldots\ldots\ldots\ldots\ldots (31)$$

The induced permittivity components, ε_E and ε_A, are on the other hand independent of T. The form of the function (31) is not generally known for solids and liquids, but it has been derived theoretically for polar gases, and for dilute solutions of polar materials in non-polar liquids.[14] The dipole moment ρ can therefore be obtained from measurements of ε_P in the gaseous or solute state. Two experimental methods are commonly used.

(*a*) ε_P is obtained from the difference between the static or low-frequency permittivity ε_0 and the optical permittivity ε_E (27a). This method involves assuming ε_A to be negligible.

(*b*) Measurements are made of the variation of ε_0 with temperature, and the temperature dependent ε_P is separated from the temperature-invariant ε_E and ε_A.

Discrepancies are often found between dipole moment values derived by the two methods, and consequently the use of a third independent method is desirable.

Such a method has become practicable with the development of the microwave region, in which many polar solutions have relaxation frequencies. As we have seen, the dielectric loss tangent passes through a maximum, and for a solution of dielectric constant ε', containing a concentration c of a polar solute, dipole moment ρ, and having a single relaxation frequency ν_R, this maximum occurs at ν_R and is given theoretically by

$$(\tan \delta_\epsilon)_{max} = A \frac{(\varepsilon' + 2)^2}{\varepsilon'} c \,.\, \frac{\rho^2}{T} \ldots\ldots (32)$$

where A is a known numerical constant. Thus measurements of $(\tan \delta_\epsilon)_{max}$ provide an independent method for the evaluation of ρ.

This method has been used for measurements on various organic polar materials dissolved in benzene.[8] Cavity resonators operating at various spot wavelengths were used to measure the $\tan \delta_\epsilon - \nu$ curve and hence derive $(\tan \delta_\epsilon)_{max}$ and ρ. The experimental curve for the solution of benzophenone, shown in Fig. 10, agrees closely with that predicted theoretically for a single relaxation frequency (30). Measurements on liquid paraffin solutions however give broader curves, attributable to a distribution of relaxation frequencies, and (32) has to be modified to allow for this.

6. The Magnetic Properties of Solids

6.1. *Origin of Magnetic Properties*

The property of magnetism is associated fundamentally with the angular momenta of charged

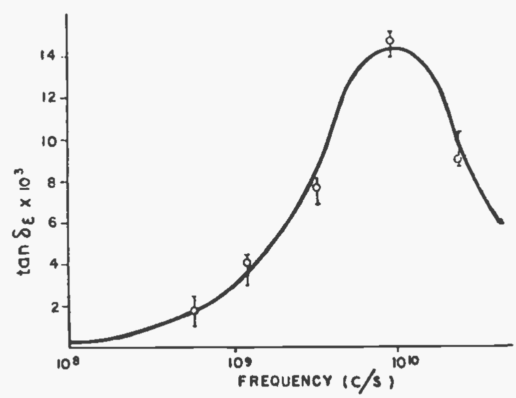


Fig. 10.—Dielectric loss of benzophenone-benzene solution concentration : 1 gm./100 cc., at 19°C.

elementary particles. The bulk magnetic properties of a material are determined by the angular momenta of the electrons in its constituent atoms. An electron has a magnetic moment, due to its spin or to its orbital motion, of

$$\mu_B = \frac{eh}{4\pi mc} \quad \dots\dots\dots\dots\dots\dots (33)$$

where e, m, h, c are the fundamental constants (electronic charge, electronic mass, Planck's constant, and the velocity of light) and μ_B is the electronic unit of magnetic moment, known as the Bohr magnetron.[18] The magnetic moments of the orbital electrons in an atom or ion add up vectorially to give the resultant magnetic moment of the ion, and these ionic moments in turn summate to give the resultant magnetic moment of the molecule.

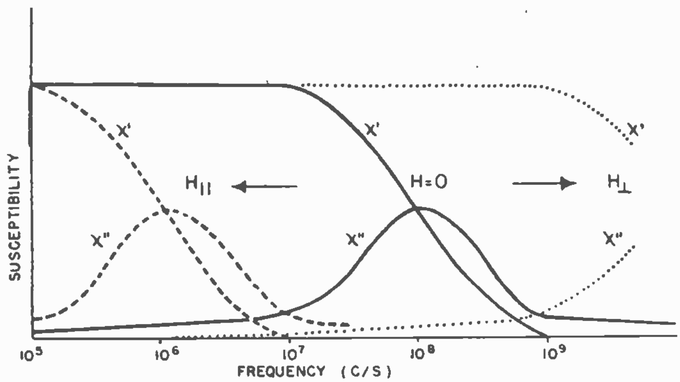


Fig. 11.—Effect of static magnetic field H on paramagnetic dispersion and absorption.

In a material, whose ions all have closed electron groups, of the inert gas-like type (e.g., $NaCl$, consisting of Na^+ and Cl^- ions) the resultant ionic moments are zero, and consequently the molecules have no magnetic moment. The only effect of a magnetic field applied to such materials is to induce a very small field in opposition. Consequently they exhibit the property of *diamagnetism* with a very small negative susceptibility of less than 10^{-6}. Ions of the transition elements (e.g., Fe, Ni, Co) and the rare earth elements have incomplete electron groups, and consequently their resultant ionic magnetic moment is not zero. Molecules containing such ions therefore act as magnetic dipoles, and a magnetic field applied to the bulk material tends to align these dipoles in the direction of the field. Such materials exhibit the property of *paramagnetism*, with a positive susceptibility, in addition to the small induced diamagnetism.

Ferromagnetism is a particular form of paramagnetism where, due to an exchange interaction between the electrons, the molecular magnetic dipoles tend to align themselves parallel to each other, forming molecular aggregates of large magnetic moment, known as domains. Ferromagnetic materials have a very high susceptibility and they display the familiar properties, known commonly as "magnetic," e.g., permanent magnetism, alignment in the earth's magnetic field, magnetic hysteresis and saturation.

Although the high-frequency magnetic properties of materials have not been so extensively studied as their dielectric properties, and the theory is in many cases incomplete, it is possible to predict the broad nature of the effects to be expected. Diamagnetics, since they are magnetically non-polar, should behave in a similar manner to non-polar dielectrics, i.e., the susceptibility should remain constant up to microwave frequencies; though to the author's knowledge, no experimental data is yet available on this. Paramagnetics, on the other hand, have a permanent magnetic dipole moment, and similar thermodynamic relaxation phenomena to those observed in polar dielectrics are to be expected in the radio-frequency region. Moreover, since paramagnetism is derived from the angular momentum of the electron, any interaction between the high-frequency field and the electron spin should be observed in the properties of the paramagnetic. In ferromagnetics, similar phenomena are to be expected, with additional effects due to the domain structure. The static magnetization process is describable in terms of domain-boundary motions and domain-rotations. In an oscillating field such movements will be subject to similar frictional constraints as those of the permanent dipoles, though the effects will be rather more complicated.

6.2. *Paramagnetic Relaxation and Resonance*

The high-frequency properties of a paramagnetic exhibit similar relaxation effects to those observed in polar dielectrics. Measurements made on paramagnetic salts (iron ammonium alum, etc.) at ~ 10-20 Mc/s show a magnetic absorption increasing with frequency[19] so that the properties may be expressed in terms of a complex susceptibility

$$\chi = \chi' - j\chi'' \quad \ldots\ldots\ldots\ldots\ldots\ldots(10)$$

Although no dispersion (i.e., variation of χ' with frequency) is observed in this frequency region, the values of χ'' correspond to a relaxation frequency ~300 Mc/s.

If a static magnetic field $H_{\parallel}$ is applied *parallel* to the H.F. magnetic field, the whole paramagnetic dispersion and absorption curve is shifted to lower frequencies, the relaxation frequency decreasing as $H_{\parallel}$ is increased.[20] The observed magnetic spectra agree well with a modified form of the theoretical relaxation formula (28) for dielectric media. If a static magnetic field $H_{\perp}$ is applied *perpendicular* to the H.F. magnetic field, the absorption χ'' is found to decrease, corresponding to a shift of the dispersion region towards higher frequencies, i.e. microwaves (Fig. 11).

Investigations of the microwave properties of paramagnetic salts in the presence of a static perpendicular field $H_{\perp}$ have disclosed a resonance effect, quite distinct from the relaxation described.[21, 22] This resonance arises from an interaction between the static field and the electron spins in the paramagnetic ions, known as Larmor precession. Theoretically an electron spin in a static magnetic field $H_{\perp}$ precesses about the direction of the field, with a frequency ν_L given by[18]

$$h\nu_L = 2\mu_B H_{\perp} \quad \ldots\ldots\ldots\ldots\ldots\ldots(34)$$

Numerically this relation becomes

$$H_{\perp}\lambda_L = 10{\cdot}7 \times 10^3 \text{ gauss-cm.} \quad \ldots\ldots(35)$$

for the wavelength λ_L, at which Larmor precession resonance may occur in a field $H_{\perp}$.

Such resonances have been observed by placing

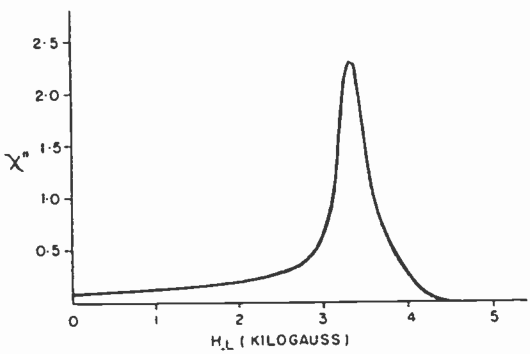


Fig. 12.—Paramagnetic resonance in $MnSO_4.4H_2O$ at $\lambda = 3{\cdot}20$ cm.

the paramagnetic salt in a resonant cavity, between the poles of an electromagnet in such a way that the lines of force are everywhere perpendicular to the H.F. magnetic field.[22] The power absorbed in the

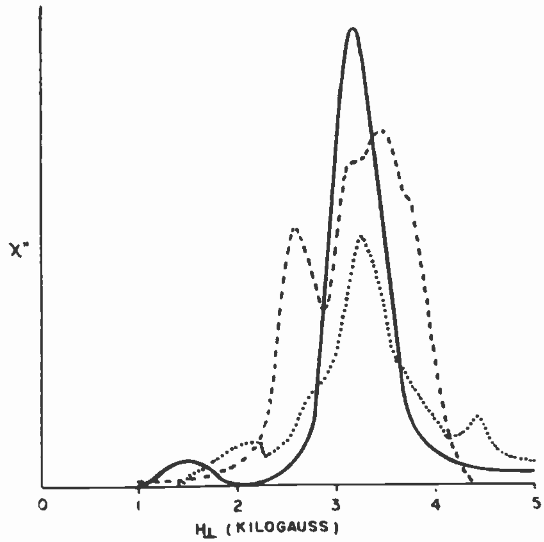


Fig. 13.—Paramagnetic resonance in chrome alum. ($\lambda = 3{\cdot}18$ cm.)

——— *$H_{\perp}$ to 100 crystal plane*
- - - *$H_{\perp}$ to 110 crystal plane*
..... *$H_{\perp}$ to 111 crystal plane*
(Ordinates in arbitrary units)

cavity at a particular wavelength is measured as a function of $H_{\perp}$, and curves similar to Fig. 12 are obtained, showing an absorption peak when $H_{\perp}\lambda$ satisfies relation (35).

In certain salts that have been investigated in this manner (e.g., chromium ammonium alum) several resonances occur in place of the single one corresponding to free electron spin.[23] These are due to a splitting of the magnetic energy levels, produced by the large local electric field in the crystal, caused by water molecules. This splitting of energy levels means that several values of the magnetic field are able to satisfy the resonant absorption condition (Fig. 13). Paramagnetic resonance thus provides a new method for investigating internal crystalline fields.

6.3. *Ferromagnetic Resonance*

Similar spin resonances have been observed in ferromagnetic metals. The metal under investigation forms part of the conducting surface of a cavity resonator, and a static field is applied perpendicular to the H.F. magnetic lines of force entering the metal. The variation of μ with $H_\perp$ is determined from the change in the Q-factor of the resonator. Measurements made on iron, nickel and cobalt at $\lambda = 1{\cdot}2$ cm. and $3{\cdot}2$ cm. each show a single spin resonance.[24] The sharpest resonance has been obtained with supermalloy, which is probably the most easily saturable material known.[25] The variation of $\mu = \mu' - j\mu''$ with $H_\perp$ for supermalloy at $\lambda = 1{\cdot}25$ cm. is shown in Fig. 14.

The ferromagnetic resonance occurs at a much higher value of $H_\perp$, than would be expected from (35). A theoretical explanation of this anomaly has been advanced[26] which indicates that the effective Larmor precession field in a saturated ferromagnetic is not $H_\perp$ but $(B_\perp H_\perp)^{\frac{1}{2}}$ where $B_\perp$ is the magnetic induction. This theory agrees within about 10 per cent. with the experimental values.

6.4. *Ferromagnetic Dispersion*

The properties of ferromagnetics are much more complicated than those of other materials. The permeability μ of a ferromagnetic is, unlike the susceptibility χ of a paramagnetic and the

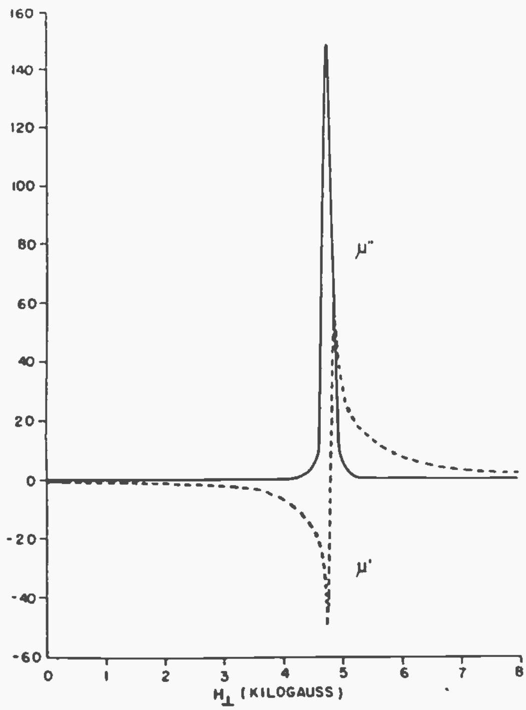


Fig. 14.—Ferromagnetic resonance in Supermalloy at $\lambda = 1{\cdot}25$ cm.

permittivity ε of a dielectric, not a constant, but it is dependent on the strength of the applied field, and also on the previous magnetic history of the material. These properties arise from the domain structure, which is peculiar to ferromagnetics. However, in very weak fields a demagnetized specimen has reversible magnetic properties which may be represented by a constant static initial permeability. By confining all measurements to such weak fields, the variation of initial permeability with frequency may be studied, in a manner analogous to dielectrics and paramagnetics.

The high-frequency properties of ferromagnetic metals into the microwave region, have been recently reviewed elsewhere[27] and will not be reconsidered in any detail here. Due to various inherent difficulties, there is as yet little correlation between the mass of experimental data and the numerous alternative theories of ferromagnetic dispersion that have been proposed, except on a purely empirical basis. The collected measurements[27] on iron are shown in Fig. 15. The experimental values of permeability fall into two distinct groups, dependent on the method of measurement employed. The permeability μ_R derived from measurements on the resistive losses of the H.F. circuit differs from the permeability μ_L obtained from measurements of the circuit reactance. This anomaly arises from the implicit assumption that μ is a real (rather than a complex) quantity at high frequencies. By confining the measurements to either the resistive or the reactive component of the circuit impedance, different values of the *apparent* real permeability μ_R and μ_L are obtained, neither of which is equal to or simply related to, the true permeability μ. The measurement of both impedance components of the same circuit element, and the consequent determination of μ, is made difficult in ferromagnetic metals by the presence of a high conductivity, associated with the permeability. These difficulties may be readily overcome, however, when the material has a low conductivity.

The microwave properties of the ferromagnetic semi-conductors, γ-ferric oxide and magnetite (Fig. 16) have been investigated, using the dual impedance method described in 4.2.1.[12, 13] The magnetic dispersion and absorption observed differs from that of a conventional relaxation, the sharpness of the $\mu'' - \nu$ curve being characteristic of a damped resonance. This resonance is similar to those described previously, which were produced by Larmor spin precession in an applied static field. In this case, however, the field about which

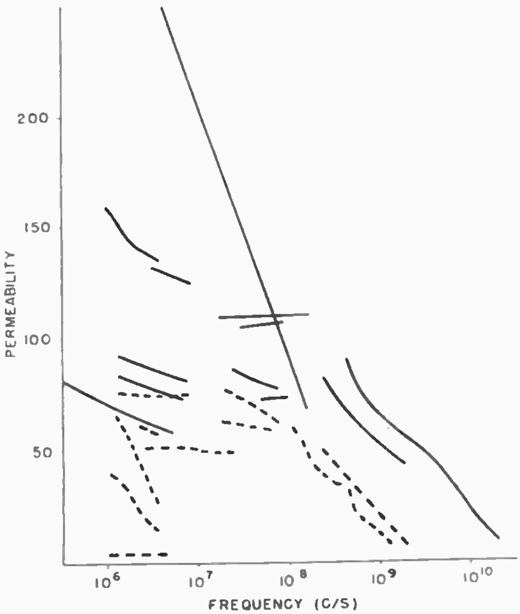


Fig. 15.—Permeability measurements on iron.
——— μR
- - - - - μL

the spins precess is the internal anisotropy field of the domains. The ferromagnetic dispersion is therefore attributed to resonance between the high-frequency field and the natural Larmor precession of the electron spins within the material.

7. Absorption Spectra of Gases

7.1. *Origin of Absorption Spectra*

Spectra, the emission or absorption of radiation by matter, are caused by the transitions of molecules, atoms or nuclei from one state of energy to another. A transition from an energy level E_2 to a lower energy level E_1 is accompanied by the *emission* of the excess energy, in the form of radiation of frequency ν given by the quantum relation

$$h\nu = E_2 - E_1 \quad \text{..................(36)}$$

Similarly, a transition from E_1 to E_2 may be accomplished by the *absorption* of radiation, according to the same relation. The magnitude of the differences in energy-level increases as we penetrate deeper into the molecular structure, and the frequency of the characteristic radiation increases accordingly. The principal types of radiation associated with the constituent parts of the molecule are shown in the table below.

Spectra	Radiation	Origin
Nuclear	γ-rays	Atomic nucleus
Atomic	X-rays	Inner electron shells of atoms
	Optical	Outer valency electrons
Molecular	Near infra-red	Vibration of atoms within molecule
	Far infra-red and microwave	Rotation of molecule

The absorption spectra of gases observed in the microwave region are mainly due to transitions between neighbouring rotational energy states of the molecules. These energy states are limited by quantum conditions to certain discrete values.

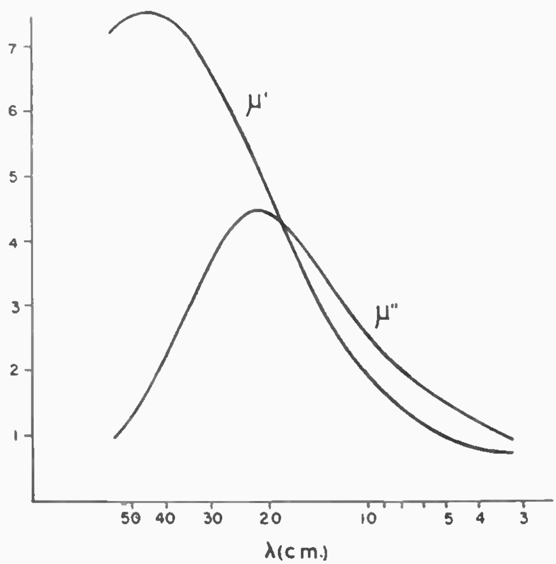


Fig. 16.—Magnetic spectrum of magnetite. (Fe_3O_4)

According to quantum theory, the angular momentum of the system must be an integral multiple of $h/2\pi$. If we consider, for example, a simple diatomic molecule, moment of inertia I, rotating with angular velocity ω, then the angular momentum

$$I\omega = Jh/2\pi \quad \ldots\ldots (37)$$

where J is an integer, known as the rotational quantum number. The kinetic energy of the molecule is consequently

$$E_J = \tfrac{1}{2}I\omega^2 = \frac{h^2}{8\pi^2 I} J^2 \quad \ldots\ldots (38a)$$

The more rigorous quantum-mechanical form is obtained by writing J(J + 1) for J^2. Consequently the rotational energy-states of a diatomic molecule are given by

$$E_J = \frac{h^2}{8\pi^2 I} J(J+1) \quad \ldots\ldots (38b)$$

Transitions can only take place between neighbouring states. Hence J can only increase by 1 (corresponding to absorption) or decrease by 1 (corresponding to emission). The frequencies of the rotational absorption lines are, from (38*b*) and (36), given by

$$\nu_{J,\,J+1} = \frac{E_{J+1} - E_J}{h} = \frac{h(J+1)}{4\pi^2 I} \quad \ldots (39)$$

where the rotational quantum-number J may take the values 0, 1, 2 . . ., etc.

Alternatively, J may remain unchanged, while some other type of energy transition (e.g., vibrational) is taking place.

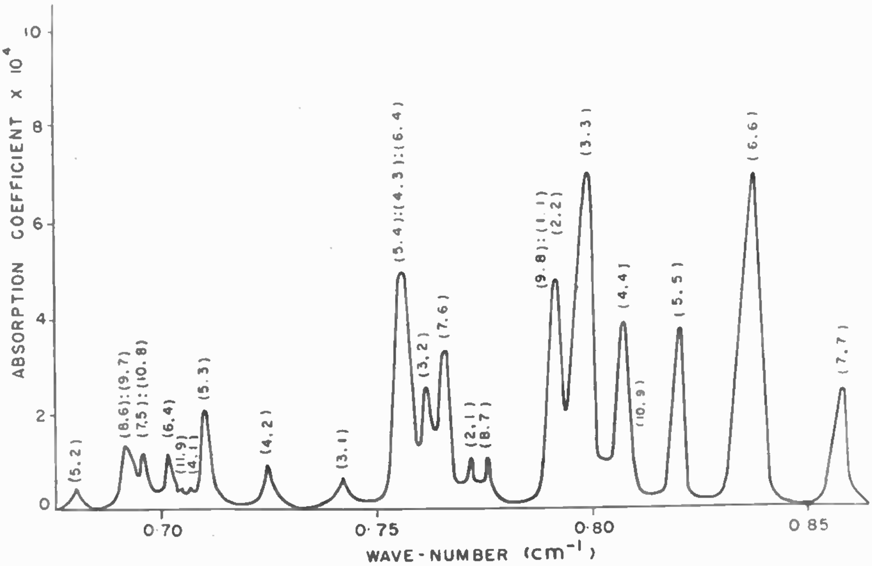


Fig. 17.—Fine structure of Ammonia Inversion Spectrum. (J, K) rotational states are indicated above lines. Pressure = 1·2 mm.

7.2. *Ammonia*

Ammonia has a strong absorption in the 1·25 cm. wavelength region which has been investigated in detail, using microwave techniques of high precision. This absorption is not due to a rotational energy transition, but to a structural property of the NH_3 molecule known as inversion. The NH_3 molecule has the form of a pyramid, with the three H atoms forming a regular triangle at the base, and the N atom at the vertex. The N atom thus has two equilibrium positions, one on either side of the base, and it is the transition between these two positions which gives rise to the microwave absorption.

The energy change in this inversional transition is that required for the N atom to penetrate the potential barrier in the plane of the H atoms. The shape of this potential barrier depends on the rotational energy-state of the molecule, since the molecular rotation causes centrifugal distortion of the H atoms. Hence a distinct inversion absorption line is associated with each of the possible rotational energy-states of the NH_3 molecule. In a non-diatomic molecule like NH_3, two quantum numbers are required to specify the rotational energy-states, J—the total number of $h/2\pi$ units of angular momentum, and K—the number of $h/2\pi$ units of angular momentum about the symmetry axis. In general $J > K$ and for purely symmetrical rotations J = K. The states are referred to by their (J, K) numbers, e.g., the (3, 2) state has J = 3, K = 2.

The inversion spectrum of ammonia has been investigated experimentally using a tunable H_o resonant cavity[28] and using an H_{01} waveguide cell, a few metres long, filled with the gas.[29, 30] Although the tunable cavity measurements have been made with great precision, the waveguide cell method is much more convenient experimentally because of its frequency flexibility, and it has hence been adopted for most of the subsequent work on microwave spectra.

The absorption spectrum of ammonia displays many interesting features.

7.2.1. *Line Breadth*

At atmospheric pressure, only a single broad absorption band is observed. As the pressure is decreased however, the breadth of the spectral lines narrows, and the fine structure corresponding to the various (J, K) rotational states becomes resolved (Fig. 17). The broadening of microwave spectral lines is due to molecular collision, and is much more pronounced than at optical frequencies, because the collision frequency of the molecules at atmospheric pressure is of the order of 10^{10}, comparable with that of the radiation. The collision frequency decreases with pressure, and consequently all measurements on the fine structure of microwave spectra are made at very low pressures. Down to pressures of 10^{-3} cm. Hg the spectral line breadth is determined solely by collision broadening, and it has been found experimentally to be directly proportional to pressure.[31]

7.2.2. *Fine Structure*

Over 30 separate lines corresponding to different (J, K) rotational states have been resolved, and identified, and an accurate empirical formula derived for their frequencies.[28, 29, 30] The intensity of each line is a function of the number of molecules in each rotational state and their inversion transition probabilities. The observed intensities agree within 5 per cent. of those calculated theoretically. The lines observed include :

$J = K$ from (1, 1) to (7, 7)
$J = K + 1$ from (2, 1) to (10, 9)
$J = K + 2$ from (3, 1) to (11, 9)
$J = K + 3$ from (4, 1) to (7, 4)

No inversion lines corresponding to $K = 0$ states occur. The spectrum stretches over the region from 19,000 to 26,000 Mc/s. Precision measurements of the line frequencies to an accuracy of ± 50 kc/s have been made using oscillators calibrated from a 50 kc/s standard crystal.[32]

7.2.3. *Isotope Effect*

The common isotope of nitrogen (abundance 99·6 per cent.) is N^{14} (i.e., atomic mass number = 14) and the spectrum so far discussed is that produced by $N^{14}H_3$ molecules. A similar spectrum has also been observed for $N^{15}H_3$, in which the line frequencies are shifted due to the increased mass of the nitrogen atom.[32, 33, 34]

7.2.4. *Hyperfine Structure*

When the pressure of the gas is reduced below 10^{-3} cm. Hg the hyperfine structure of the individual (J, K) lines can be resolved for the more intense $N^{14}H_3$ lines.[29, 34, 35] Four small satellite lines spaced symmetrically about the main line, and separated by less than 1 Mc/s, are observed (Fig. 18). This hyperfine structure results from quadrupole coupling of the N^{14} nucleus with the electric field of the molecule. The coupling depends on the orientation of the nuclear spin with respect to the rest of the molecule, and hence a given rotational level is split into a series of levels corresponding to the different possible values of the total (nuclear + rotational) angular momentum. From the spacing and intensity of the satellites it is possible to derive the nuclear spin, as well as the quadrupole coupling. The $N^{15}H_3$ lines do not show hyperfine structure, because the N^{15} nucleus has a spin of $\frac{1}{2}$.[33, 34]

The high resolution necessary to observe the hyperfine structure is obtained by using large electrical path lengths, and waveguide cells up to 100 ft. long have been employed for resolving lines up to $J = 7$.[36] The spectra are recorded on a cathode-ray oscillograph using the sweep frequency method.[29] As described in 2.2, above, the frequency of a klystron oscillator may be varied over a small range, by varying the reflector voltage. In the sweep frequency method, the klystron is tuned manually to the frequency of the line to be observed and a low-frequency sawtooth voltage is applied to the reflector which sweeps the klystron frequency over a range of a few Mc/s. The same sawtooth voltage is applied to the X-plates of an oscillograph, thus making its horizontal scale a frequency scale, since the oscillator frequency and the oscillograph spot are being swept in synchronism. The amplified voltage from the detector is applied to the Y-plates, producing a deflection proportional to the absorption in the waveguide cell, and thereby plotting the structure of the line on the oscillograph screen. The sweep frequency method has become the standard method of observing microwave absorption spectra in detail.

7.2.5. *Stark and Zeeman Effect*

The splitting of a spectral line into separate components by the application of an electric field is known in optics as the Stark effect. The similar phenomenon caused by a magnetic field is called the Zeeman effect. Both these effects have been observed with lines in the ammonia inversion spectrum.[33] For the Stark effect, the static electric field was applied between the waveguide wall and a thin insulated metal strip running down the centre of the waveguide. The Zeeman effect was observed in a perpendicular magnetic field of 6,600 gauss, which caused a splitting of each inversion line into a doublet, with a separation of about 6 Mc/s.

7.3. *Water Vapour and Oxygen*

Both water vapour and oxygen have absorption frequencies in the microwave region. The intensities of their absorption lines are small compared with ammonia, being due to more infrequent energy-transitions, but because of the presence of these two materials as major constituents of the atmosphere, their absorption characteristics have a pronounced effect on the propagation of the shorter microwaves.

7.3.1. *Water Vapour*

The absorption of microwaves by water vapour is due to two effects[37] :

(*a*) a single absorption line at $\lambda = 1{\cdot}348$ cm. due to a transition between two rotational energy states known as the 5_{-1} and the 6_{-5} states.

(*b*) The combined residual effect of all other absorption lines whose wavelengths are too short for resonance.

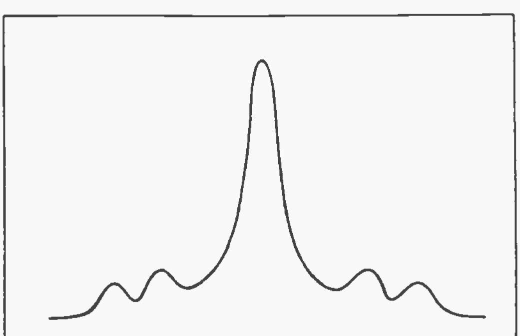


Fig. 18.—Hyperfine structure of NH_3 (2, 2) inversion line, resolved by sweep-frequency method.

There is a sharp peak in absorption due to (*a*) at $\lambda = 1{\cdot}35$ cm. of about $0{\cdot}25$ dB/km. per gram of H_2O per cubic metre. The absorption due to (*b*) is proportional to the square of the frequency (Fig. 19).

A precise measurement of the wavelength of the absorption line has been made on pure H_2O vapour at 10^{-2} cm. Hg pressure, using the sweep frequency method.[38] Extensive measurements on absorption from $\lambda = 0{\cdot}7$ cm. to $1{\cdot}7$ cm. have also been carried out under atmospheric conditions, for a wide range of humidities, using an ingenious multi-mode cavity method.[39] Radiation from a pulsed magnetron is fed into an air-filled cubical copper box, 8 ft. high. Strings of thermocouples, 360 in all, coated with " lossy " material, are placed at random in the box, which due to its large dimensions resonates simultaneously in many high-order modes. The e.m.f. of the thermocouples is found experimentally to be proportional to the Q-factor of the cavity and its contents. The Q of the cavity is determined from additional measurements taken when an aperture is opened in the side of the box. The total Q is then measured when the box is filled with circulating atmospheres of different humidities, and is given by

$$1/Q_{total} = 1/Q_{box} + 1/Q_{vapour} + 1/Q_{aperture}$$

from which the water vapour absorption may be determined. Several magnetron oscillators operating at spot wavelengths are used to cover the range of the measurements.

7.3.2. *Oxygen*

Oxygen has an absorption line just within the microwave region at a wavelength of $0{\cdot}502$ cm. This absorption is of theoretical interest, because electrically non-polar molecules like O_2 do not normally interact with radiation at such relatively long wavelengths. Oxygen, however, is paramagnetic, and it is the magnetic moment of the O_2 molecule which interacts with the electromagnetic field. The line at $0{\cdot}502$ cm. corresponds to an energy transition between different fine-structural states in the O_2 molecule.[40]

The wavelength of the oxygen absorption line lies below the range of existing klystron oscillators. The source of $\frac{1}{2}$ cm. radiation, used for the experimental investigations[41] was a silicon-tungsten crystal detector, operated as a generator of the second harmonic of signals produced by a 1-cm. klystron oscillator. These 1-cm. signals were intro-

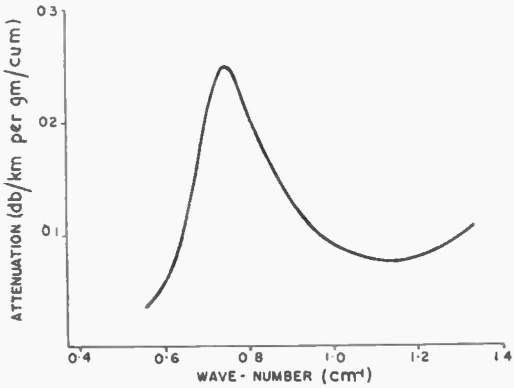


Fig. 19.—Microwave absorption of water vapour.

duced into the crystal via a coaxial line, and the ½ cm. harmonic extracted into a waveguide circuit. The efficiency of the harmonic generator (available ½ cm. power : available 1 cm. power) was less than 10 per cent. The absorption path used in the measurements was a 6-metre length of 0·18″ × 0·086″ H_{01} waveguide. Straight crystal detection was employed with the source modulated at 1,000 c/s to reduce detector noise. The maximum absorption observed at $\lambda = 0{\cdot}502$ cm. corresponds to an attenuation of 14 dB/km. in air, 70 dB/km. in pure oxygen, at atmospheric pressure and a temperature of 27°C.

7.4. *Other Gases*

The techniques of microwave absorption spectrography represent a considerable advance over those available in the neighbouring far infra-red region. They are consequently being increasingly applied to the investigation of pure rotational spectra, and to the resolution of their hyperfine structure. The principal molecular rotational lines studied to date are as follows :

Molecule	Rotational Transition
OCS	$J = 1 \rightarrow 2$
N_2O	$J = 0 \rightarrow 1$
CH_3Cl	$J = 0 \rightarrow 1$
CH_3Br	$J = 1 \rightarrow 2$
CH_3I	$J = 1 \rightarrow 2$
ClCN	$J = 1 \rightarrow 2$
BrCN	$J = 2 \rightarrow 3$
	$J = 3 \rightarrow 4$
ICN	$J = 4 \rightarrow 5$
CH_3CN	$J = 0 \rightarrow 1$
	$J = 1 \rightarrow 2$
ICl	$J = 0 \rightarrow 1$
SO_2	Various

From these measurements important physical data such as dipole moments, molecular moments of inertia, bond distances, nuclear quadrupole moments and nuclear spins have been derived.

8. Conclusion

In concluding this review, the author is aware of several important physical applications of microwaves—superconductivity effects, molecular beam resonances, microwave electron accelerators—that have not received mention. It is hoped, however, that this paper has succeeded in its principal object—that of demonstrating the scope and interest of the new field of physical research, opened up by the wartime development of microwaves.

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TRANSFERS AND ELECTIONS TO MEMBERSHIP

Subsequent to the publication of the list of elections to membership which appeared in the November/December issue, there has been a further meeting of the Membership Committee on December 7th, 1948. Twenty-three proposals for Direct Election to Graduate or higher grade of membership were considered, and 16 proposals for Transfer to Graduate or higher grade of membership.

The following list of elections was approved by the General Council : 22 to Direct Election to Graduate or higher grade of membership and 12 for Transfer to Graduate or higher grade of membership.

Direct Election to Full Member

DENHAM, Humphrey John, C.B.E., M.A., D.Sc. (Oxon),	Kidlington, Oxon

Direct Election to Associate-Member

BROWNE, de Courcy, Arthur William, Major	Strathfield, Australia
MITCHELL, Eric	Westcliff-on-Sea, Essex
MORGAN, Ronald William	Barbados, B.W.I.
PROPERJOHNS, Frederick James Frank	Purley, Surrey
*RANSON, Humphry Bohun	London, N.5
SWAINSON, David Robert Wyer	Enfield, Middlesex
TYRRELL, Sydney James	Esher, Surrey

Direct Election to Companion

COX, Herbert William	Mombasa, Kenya

Direct Election to Associate

BOLTON, Ronald,	Blackpool, Lancs.
*COMERFORD, Gerard Alphonsus	Edgware, Middlesex
DIAPER, Eric Thomas	Gosport, Hants.
LILLEY, Robert William	Dar-es-Salaam, Tanganyika
RAMANATH, K. R., Lieut., B.Eng.	Poona, India
*SQUIRES, Terence Leighton	Whitby, Yorks.
WILKES, Samuel John Herbert, B.Sc. (Hons.)	Birmingham, 25

Direct Election to Graduate

ALEXANDER, James Ronald, B.Sc.	Leigh-on-Sea, Essex
DEWEY, Thomas Norman, B.A. (Cantab).	Bromley, Kent
HODGSON, William Cliffe, Lieut.(L)	By Stirling, Scotland
RIDDELL, Matthew Kent, B.A.Sc.	Rugby, Warwicks.
SRIVASTAVA, Shamblu Laran, M.Sc.	Southampton, Hants.
STEPHENS, William George Sinclair	Cupar, Fife

* Reinstatement.

Transfer from Associate-Member to Full Member

HEIGHTMAN, Denis William	Clacton, Essex

Transfer from Associate to Associate-Member

GRAHAM, William Frederick	Brentford, Middlesex
HELSDON, Peter Bennett	Chelmsford, Essex
HODGKINSON, John Thomas	Nairobi
MAY, Bernard Calverley, Lieut.(L)	Chatham, Kent

Transfer from Graduate to Associate

BUMSTEAD, Maurice Charles,	Hythe, Kent

Transfer from Student to Associate

GILL, Owen John	Tanganyika, E. Africa
MATHEWS, Leonard Frederick	Manchester

Transfer from Student to Graduate

DRUMMOND, Gilbert Park	Glasgow, S.1
FLAUM, Ronald Raphael	Ilford, Essex
GILMOUR, George	Douglas, Isle of Man
LAVERICK, Mrs. Elizabeth, B.Sc.(Hons.)	Fencehouses, Co. Durham

V.H.F. RADIO EQUIPMENT FOR MOBILE SERVICES*

by

D. H. Hughes†

Read before the West-Midlands Section on October 27th, the South-Midlands Section on October 28th, the North-Western Section on November 5th and the London Section on November 18th.

SUMMARY

The first two Sections are introductory: the first explains the advantages of the V.H.F. band for local radio-telephone services to mobile vehicles, while the second briefly reviews some of the points which must be considered in planning such services.

Most of the Paper is concerned with the design of suitable equipment. Sections 3 and 4 deal with the principal factors in the design of the mobile and fixed stations, respectively, and these are illustrated in Section 5 by a description of a particular range of equipment which has been produced commercially. In conclusion, there is a short discussion of future trends in development.

1. Introduction

In the last few years, there has been a very large increase in the use of radio-telephone systems for communication with mobile vehicles, particularly in the United States of America,[1] but also in many other parts of the world. In addition to such well established services as police communications and others of an essential nature, similar systems are now being used by purely commercial interests, such as taxi and tug companies. At the same time, there has been an upward movement in the radio frequencies employed, from the M.F. and H.F. bands to the V.H.F. band; whilst this has resulted in many more frequency channels becoming available to relieve the congestion in the lower frequencies, the principal reasons for the change lie in the improved performance obtainable.

On the lower frequency bands, the available channel width does not exceed 10 kc/s and, therefore, amplitude modulation is universally used; since circuits for suppressing impulsive interference have a poor performance in receivers with a narrow band-width, ignition interference is extremely serious at these frequencies. In the V.H.F. band, on the other hand, channel widths are at least 20 kc/s and this allows the use of either frequency modulation or wide-band amplitude-modulation systems with efficient noise-suppression circuits: thus ignition interference can be largely overcome.

Another disadvantage of the M.F. and H.F. bands lies in their ability to traverse very long distances, especially at night, so that long-range interference between many widely separated services is extremely serious. Mobile R.T. services are mostly of a relatively local character and these are adequately served by ground-wave propagation. Due to the need for penetration in built-up areas and hilly country, the U.H.F. band, with its nearly optical propagation properties, is generally unsuitable, but part of the V.H.F. band approaches very near to the ideal. Frequencies in the region of 30 to 40 Mc/s have been, and still are, extensively used, but are liable to cause excessive long-range interference. 70 to 100 Mc/s is a very suitable range of frequencies: it suffers to some extent from long-range interference, but much less than the lower frequencies, and it gives reasonable coverage in built-up areas. With the increasing demand for frequency channels, a further band, around 160 Mc/s, is now coming into use; this does not penetrate obstructions as easily as the lower bands, but it is fairly adequate for local services. Long-range interference is likely to become more serious, due to tropospheric propagation, as the frequency increases above 100 Mc/s. The features of the various frequency bands are summarized in Table 1.

The service area of the V.H.F. bands can be considerably extended by the use of the multi-carrier A.M. system.[2,3]

2. Planning of Systems

It is not within the scope of this Paper to discuss in detail the planning of complete systems of various types, but a brief survey of some of the relevant points is desirable before the design of the equipment is considered.

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Table 1. Comparative performance with different frequency bands.

Frequency Band	Channel Width	Impulsive Interference	Long-range Interference	Penetration
M.F. & H.F.	$\not>$ 10 kc/s	Very bad	Very bad	Good
30-45 Mc/s	20-60 kc/s	Good	Bad	Good
70-100 Mc/s	40-100 kc/s	Good	Fairly good	Usually adequate
150-180 Mc/s	60-120 kc/s	Good	Fair	Fair
U.H.F.	Wide	Good	Bad at times	Bad

2.1. *Location and Power of Fixed Stations*

The various factors affecting these are (*a*) the allocated frequency, (*b*) the range of service required, (*c*) the nature of the terrain to be covered, e.g. heavily or lightly built-up urban areas, flat or hilly rural areas, etc. In general, the higher the site of the fixed station the greater the coverage obtainable, but the aim should be to restrict the coverage as nearly as possible to the required area, in order to reduce interference outside this area. When a choice is available, it may be possible to use hills, not only to provide a good site, but also to curtail the coverage of unwanted territory.

To serve a small town, it is usually sufficient to employ a low-power transmitter at some central point, with the aerial raised above the level of the surrounding buildings. For a larger area, the fixed station should be placed at a suitable high point in open country, where conditions for reception are good, and a high-power transmitter should be used. When a single station is inadequate, two or more stations should be used in a multi-carrier, or similar, scheme.

2.2. *Methods of Modulation*

The two methods in general use are (*a*) wide-band A.M., with impulsive-noise limiters, (*b*) F.M., with a deviation of 12 to 20 kc/s. It cannot be said that either is always superior for mobile R.T. services and the relative merits of the two systems should always be considered in relation to each particular application.

For coverage of large areas, the multi-carrier A.M. system is at present predominant, although efforts are being made to develop equivalent F.M. systems. For single-station schemes, F.M. can give better results than A.M., due to its greater immunity from random noise interference: this presupposes, however, that the F.M. equipment is correctly adjusted, as its performance in respect of noise rejection deteriorates rapidly with detuning. For coverage of small areas in which the ambient noise level is not unduly high, there is little difference between A.M. and F.M. and the former has the advantage of requiring less skilled maintenance.

Up to the present, A.M. equipment has generally been less expensive than F.M. equipment, but their costs are now tending to become more nearly equal; this is due partly to the fact that A.M. receivers are becoming more elaborate to meet increasingly stringent performance requirements, and partly to the development of simpler techniques in F.M. circuit design.

2.3. *Simplex and Duplex Operation*

In the simplex system, communication is possible in only one direction at a time and the receiver and transmitter at each station operate alternately, manual switching being employed; a limitation of this system is that one operator is unable to interrupt the other. Either the same or different frequencies may be used for inward and outward transmission; J. R. Brinkley has shown[3] that, where the same frequencies must be used for different services in adjacent areas, there is a definite economy in employing two-frequency operation.

Duplex working makes simultaneous transmission and reception at each station possible, so that conversations can take place as on an ordinary telephone circuit ; two frequencies are, of course, essential. As will be explained in Section 3.2, the difficulties of aerial installation are increased, particularly at the mobile station, but there are two distinct advantages in having a duplex main station, even if the mobile stations are simplex : one is that connection to a telephone exchange is greatly simplified ; the other is that direct conversations can take place between mobile stations, using the main station as a relay station. There are, however, special cases where direct communication between mobile stations is required at times when they are out of range of the fixed station : this condition can arise with sea-going tugs, and can be satisfied economically only by single-frequency simplex working.

3.0. Design of Mobile Equipment

3.1. *General Mechanical Design*

In most mobile set installations, the space available is restricted and it is an advantage if the equipment is small. On the other hand, it is often subjected to high ambient temperatures with poor ventilation, and a limit is set to the minimum size by the maximum safe working temperature of the components used. The ventilation provided in the covers of the set is restricted to louvres or similar methods, as some protection must be given from rain and splashes. Components and finishes should be suitable for use in a damp, salty atmosphere, so that the equipment can be used for marine applications, and, if the export market is to be considered, a fully tropical finish is advisable.

Mobile sets are usually built in two or more units, assembled on some form of cradle or carrier ; this is mounted on shock absorbers to provide protection against vibration and road shocks. It is common practice to build the transmitter and its power supply in one unit and the receiver with its power supply in a second unit. Sometimes the transmitter and receiver are combined on one chassis, but this arrangement is not suitable for duplex operation, owing to interference between transmitter and receiver. In other cases, the power supplies form a separate unit, while in one model, to be described in Section 5.1, the modulator, which is also used as a public address amplifier, is separated from the transmitter, making four separate units. The equipment is remotely controlled through a multi-core cable from a small control unit, which is mounted on the dash-board of a car, or other convenient place. Inter-unit cable connections are made by easily detachable plugs and sockets, which should be provided with some form of retainer as protection against vibration. Ease of maintenance is of great importance and it is desirable that a faulty unit should be capable of speedy replacement by the user, who will often have little knowledge of radio. The use of a number of small units has advantages from this point of view, although it does cause a small increase in cost.

Miniature valves and components are largely used in modern equipment, but it has been found advisable to make the chassis somewhat larger than the minimum necessary to accommodate them, both to improve accessibility and to prevent over-heating. Each unit can consist of either a single chassis or an assembly of sub-units: the single chassis is cheaper for large-scale production of a standardized model, but the second method provides greater flexibility for meeting varied requirements and has some advantages in assisting maintenance and repairs.

3.2. *Aerials*

Where sets are installed in cars, the usual aerial is a quarter-wavelength vertical rod. For frequencies higher than about 60 Mc/s, this should preferably be mounted on the car roof, the objects being to obtain maximum height above ground and to provide a metal earth-plane immediately below the aerial. If the car roof is non-metallic, a horizontal sheet of metal should be fitted below the aerial ; alternatively, an equivalent of an earth-plane, such as four or more horizontal quarter-wavelength rods, projecting radially from the base of the aerial, may be used. It is not always possible to mount the aerial on the roof and sometimes an alternative position such as the top of the luggage boot, is preferred on grounds of appearance : in such cases, the aerial will have an impedance different from that of a quarter-wavelength aerial and the matching to the set must be adjusted ; loss of range will be experienced, due to the loss in effective height, and the polar diagram round the car is likely to be irregular, due to the lack of symmetry in the ground-plane and to screening by part of the metal car body projecting above the base of the aerial. For lower frequencies, such as the 30-40 Mc/s band, a quarter-wave-length rod is usually inconveniently long for roof-mounting, and a lower position must be used. Either the aerial base or the rod itself must be flexible, in order to prevent damage by overhead obstructions.

For tugs and other vessels, a co-axial dipole aerial, fixed to the top of the mast or funnel, can be used. This type of aerial is effectively a centre-fed half-wave dipole, with the lower element taking the form of a tube, concentric with an inner supporting tube, through which the feeder cable is led to the centre of the assembly.

Railway locomotives present a difficult problem, owing to their low clearance for bridges and tunnels. Special types of aerial with a small vertical dimension have been developed for this application.

For simplex operation, a single aerial is used, this being switched alternately to the receiver or the transmitter by a relay ; when two frequencies are used, it is cut to suit either the mean of the two frequencies or that of the mobile transmitter. For duplex operation, it is customary to use two separate aerials for the transmitter and receiver and considerable care is necessary to minimize the coupling between them. In the case of a car, this means that they must be mounted fore and aft, using the body of the car as a partial shield between them ; this results in an appreciable loss in efficiency and irregular polar diagrams. It is possible to use a single aerial with a duplex equipment by fitting filters in the transmitter and receiver aerial leads, but, in practice, there is considerable difficulty in designing filters which are small, efficient and suitable for production. It has been found, however, that the inclusion of a simple L C rejector circuit in the receiver lead enables two aerials to be used with a separation of only a few feet, so that they can both be mounted on the roof of a car.

3.3. *Power Consumption*

In mobile installations, power is almost always obtained from a storage battery, frequently from the one supplying the normal electrical installation of a car. It is, therefore, extremely important to reduce the power consumption to a minimum. The load may be divided into two parts, (*a*) continuous, (*b*) intermittent. The continuous load comprises the receiver L.T. and H.T. and, in many cases, the transmitter L.T. : economy here is vital. The transmitter H.T. load, being intermittent and normally of short duration, is less important.

The receiver H.T. requirements can be reduced by running the valves below their normal rating : performance is quite adequate with a potential of 200 volts or less and many stages can be run at about half their maximum rated current. Vibrator power packs are more efficient than rotary transformers, but the latter have a longer life and are more reliable than vibrators.

The heater current of most types of British indirectly-heated receiving valves is 200 or 300 mA at 6·3 volts, while some American types require only 150 mA. Until valves with a lower consumption become available, the number of valves used in the receiver must be reduced as far as possible. It is not possible to use battery valves with 1·4-volt filaments, partly owing to their inferior performance, and partly owing to the difficulty of keeping their filament voltage within the permissible limits when the source of supply is a 12-volt car system.

When the facility of immediate talk-back is required, no time can be allowed for the transmitter valves to heat up and, so long as they are indirectly heated, they must be run continuously : this represents a serious waste of power when the transmit/receive time ratio is small. This problem has been solved in America by the production of a number of types of transmitting valve with directly-heated filaments, such as the 2E24 and 2E30 : these can be switched off except during the actual periods of transmission.

3.4. *Receiver Design*

While the number of systems in operation was small, the risk of mutual interference was relatively low and selectivity requirements were of secondary importance compared with sensitivity, signal/noise ratio and frequency stability. Now, however, congestion in the frequency bands is already evident in some places and will obviously become an increasingly important factor in the future. On bands below 100 Mc/s, channels are being allotted with a separation of only 50 kc/s. An adjacent-channel selectivity of at least 80 dB is desirable : with less than 60 dB, interference between services on adjacent channels in the same area is very serious. With an F.M. system with 15 kc/s deviation, a band-width of about 36 kc/s is desirable, and to this must be added an allowance for frequency drift and crystal tolerances, so that the designer is set a very difficult problem. With an A.M. system, the position is a little better, partly owing to the lower susceptibility of A.M. systems to adjacent channel interference[4], and partly because the minimum usable pass-band (neglecting frequency tolerance) is only 6 kc/s : such a narrow band would, however, allow very poor suppression of impulsive interference, for which purpose a band at least five times as wide is desirable (this can include the allowance for frequency errors), but in this case a compromise is possible between the two types of interference (adjacent-channel and impulsive-noise), whereas with F.M. any narrowing of the pass-band will cause distortion of the wanted signals. The first necessity is the reduction of frequency errors and temperature effects to the absolute minimum. Secondly, the I.F. should not exceed about 3 Mc/s, as a sufficiently sharp cut-off slope is difficult to obtain at higher frequencies with an economical number of circuits. An attractive solution of the problem would be the use of a crystal band-pass filter, operating on a frequency of 5 Mc/s or more, but it is not yet possible to obtain suitable crystals free from unwanted responses.

Apart from adjacent-channel interference, second-channel and other spurious responses are liable to cause trouble, and selectivity against these should also be of the order of 60 to 80 dB. With an I.F. of 3 Mc/s or less, such a performance cannot be obtained in the R.F. circuits and a double superheterodyne becomes essential, with a first I.F. of the order of 10 to 20 Mc/s. The two frequency-changers may be supplied either from independent oscillators, each controlled by a separate crystal, or from a common oscillator, the first mixer using a harmonic and the second the fundamental component. Either method can give satisfactory results, but in each case care must be taken in the choice of frequencies to avoid spurious interference.

To obtain a good signal/noise ratio, the aerial circuit must provide as much gain as possible and a low-noise R.F. amplifier with a stage gain of at least five times is necessary. Too much gain before the second mixer must be avoided, however, as otherwise cross-modulation will be caused by a strong signal on an adjacent channel.

In an A.M. receiver, an efficient impulsive-noise limiter[5] is essential. Care is necessary to ensure that no path is provided by which the impulses can by-pass the limiter, and that they cannot cause blocking in one of the I.F. amplifier stages.

In an F.M. receiver, a muting, or squelch, circuit must be employed. It also becomes necessary in an A.M. receiver if the sensitivity is better than about 2 μV. If a calling system is used, it includes a muting circuit.

The A.G.C. action should maintain the output reasonably constant over a wide range of input signal strength and should be fast enough to cope with rapid "flutter" fading. No difficulty is experienced in achieving these requirements in an F.M. receiver with a reasonably good limiter ; in an A.M. receiver, equally good results can be obtained by employing A.G.C. on the first A.F. amplifier and by using circuits with short time constants to supply A.G.C. to the controlled stages before the detector.

Great care must be taken in the design of the power pack, whether it employs a vibrator or a rotary transformer, to prevent noise from reaching the receiver. R.F. "hash" interference is, of course, the most troublesome and may cause interference either by direct conduction or by radiation into the input circuits, or even by radiation into the aerial feeder from the other cables or the power pack itself. Methods of prevention include the fitting of R.F. chokes and by-pass condensers at strategic points, screening, and very careful selection of earthing points.

3.5. *Transmitter Design*

The R.F. power output of British mobile transmitters is generally between 7 and 20 watts, while American sets range up to 30 and even 50 watts. With the existing range of British valves, an output of 10 to 15 watts represents a reasonable compromise between performance, on the one hand, and cost and power consumption, on the other. To ensure efficient matching to the aerial in different types of installation, it is necessary to provide means for adjusting the output circuit when the equipment is installed.

Radiation of harmonics, particularly the second, must be kept to a low level to avoid interference with higher frequency bands : for example, the second harmonic of an 80 Mc/s transmitter would interfere with a system on 160 Mc/s. This necessitates a carefully balanced push-pull circuit or the use of a pair of coupled circuits in the final stage.

In an A.M. transmitter, the oscillator is controlled by a crystal, operating on a frequency which is usually between 5 and 15 Mc/s. Crystals employing overtone excitation on considerably higher frequencies have not yet come into very general use, but are likely to do so for both transmitters and receivers when more experience has been gained with them and they have been proved sufficiently reliable in operation. The oscillator is followed by one or more frequency multipliers to obtain the carrier frequency. Anode modulation of the power amplifier is customary.

In F.M. mobile transmitters, a phase-modulation system is universally employed. The crystal-controlled oscillator operates on a relatively low frequency, as a total multiplication of at least 48 times is generally necessary; in consequence, more R.F. stages are required than in an A.M. transmitter. On the other hand, less power is required for the modulator. The alternative system, employing a reactance modulator, does not appear to have been used at all in this type of equipment; indirect control from a crystal oscillator is necessary and this itself requires several valves, which more or less offset the saving in multiplier stages. It is not possible to use a portion of the receiver for control purposes when the equipment may be used for duplex operation; even if the equipment is intended only for simplex operation, the use of different frequencies for transmission and reception would introduce further complications. The low oscillator frequency, necessary for a phase-modulated transmitter, has the disadvantage that the ratio of the frequency separation of the carrier from adjacent oscillator harmonics to the carrier frequency $\left(\text{i.e. } \frac{fc \pm fo}{fc}\right)$ is small: this can cause difficulty in selecting the correct harmonic during alignment; it also increases the difficulty of suppressing the radiation of unwanted signals, spaced on either side of the carrier frequency by the oscillator fundamental frequency.

The H.T. power requirements are beyond the capacity of a single vibrator of any type at present available; although two vibrators are sometimes used in American equipments, the general practice is to use a rotary transformer. Voltages range between 250 and 450, according to the types of valve used and the power output required.

A facility which is often required from a mobile installation is its use as a public address equipment. With an A.M. system, it is only necessary to switch the output of the modulator to a public address loudspeaker instead of to the transmitter power amplifier. With an F.M. system, there is more difficulty: either an A.F. power amplifier must be fitted specially for the purpose, or the power stage of the transmitter must be supplied with the necessary A.F. components and the circuit arranged so that it can be made to act either as a transmitter or as an A.F. amplifier.

3.6. *Control and Calling Circuits*

As mentioned in Section 3.1, mobile transmitters and receivers are always operated remotely from a small control unit, which, in a car installation, is usually fitted to the dashboard. This unit is connected to the radio set by a multi-core cable and it is preferable that this should carry only low-impedance speech circuits and D.C. control circuits operating at low voltages and currents. Switching of the power circuits in the transmitter and receiver and transmit-receive switching are, therefore, performed by relays. A press-to-talk switch is fitted to the hand-microphone to operate the transmit-receive relays, which switch on the transmitter H.T. supply and the filaments of any directly-heated transmitter valves; in simplex equipments, the aerial is switched over and the receiver H.T. supply switched off during transmission.

Some equipments employ a loudspeaker and a hand-microphone, while others use a combined microphone and telephone handset. In the latter case, unless a calling device is fitted, a loudspeaker must also be used and this is switched off when the handset is picked up. The microphone may be either a carbon or a moving-coil type.

Calling devices, e.g. a bell, buzzer or lamp, may be operated by any received carrier, by speech, or by a tone modulation transmitted by the fixed station. The last of these is the only method which can be made proof against operation by interference. If the tone is not continuous, but takes the form of a series of coded impulses, this is termed a selective-calling system and the mobile set can be arranged to respond to one particular set of impulses; this requires an appreciable number of extra components, including some form of stepping relay.

4.0. Design of Fixed Station Equipment

4.1. *General Considerations*

Various types of fixed station installation are needed to suit different requirements; they may be classified as follows:

(*a*) Single-station schemes, employing either a low-power or a high-power transmitter. In some cases, one or more satellite receiving stations, linked to the main station by landline, may be required.

(*b*) Multi-carrier schemes[2, 3]. These require a headquarters station, which may also be a main transmitting and receiving point, and one or more satellite main stations, linked to headquarters by radio or landline.

Three types of control system are used:

(*a*) Direct control, where the operator is beside the equipment.

(*b*) Extension control, where the operator and the equipment are in different parts of the same building.

(*c*) Remote control, over a landline, where the equipment is unattended and may be some miles from the operating point.

Apart from the control units, the various parts of the equipment are all built on standard panels for rack-mounting, thus permitting maximum flexibility in station design.

For omni-directional working, the aerial is usually either a single half-wavelength element, such as a co-axial dipole or an end-fed half-wave with quarter-wave matching section, or a vertically stacked array of two or three such elements. Reflectors and/or directors are sometimes added to introduce a measure of directivity if uniform omni-directional propagation is not wanted.

4.2. *Receivers*

It is common practice to adapt the mobile receiver for fixed station operation by building it on to a panel for rack-mounting and by fitting a mains-operated power pack instead of a battery-operated unit. This has the advantage of a measure of standardization and thus simplifies maintenance and the stocking of spare parts. It has the disadvantage that the performance is restricted by the limitations imposed on the design of the mobile receiver by the need for economy in size and in power input. In many applications, particularly where a high-power fixed transmitter is used, there is a need for a superior performance from the fixed receiver, and, in addition, various refinements can be added, which are not justified in the conditions pertaining to the mobile equipment.

All the performance requirements, which were discussed in Section 3.4 in relation to the mobile receiver, apply equally to the fixed receiver and their attainment is somewhat easier. The sensitivity, noise factor and rejection of spurious responses can be improved by the addition of a second R.F. amplifier; adjacent-channel selectivity can be increased by the use of more or better I.F. circuits.

4.3. *Low-power Transmitters*

In this case, there is usually no disadvantage in adapting the mobile transmitter for rack-mounting and A.C. operation. Minor alterations for the external connections may be necessary and it is a useful refinement to add a meter and switch on the panel for metering and monitoring. Where a calling system is required, an addition for this purpose must also be made to the transmitter.

4.4. *High-power Transmitters*

The R.F. power output is usually between 50 and 500 W, a commonly used value being 100 W. The transmitters are usually built on a number of separate panels, assembled in a rack or cubicle. In the case of the smaller models, the receiver and auxiliary equipment may be housed in the same rack as the transmitter, while the larger transmitters form a separate, self-contained unit. In most cases, high-power stations are required for unattended operation and adequate safety devices must be included, both to prevent damage in the event of a fault occurring and to protect maintenance personnel from dangerous shocks. All components must be designed and operated with an adequate margin of safety, with a view to ensuring reliable operation over long periods without maintenance.

Circuits usually follow a conventional pattern and need not be discussed here.

4.5. *Control Systems*

4.5.1. *Extension Control*

Where a station must be operated from two or more different points in a building, or where it is not convenient to instal the equipment at the operating point, it is possible to extend the microphone, receiver output and control circuit leads to a small control unit through a multi-core cable, without the need for any amplifiers at the control point ; power switching should be performed by relays, but these are commonly fitted in any case. The length of cable which can be used depends on the circuit impedances and the resistance of the leads ; 50 ft. is adequate for many requirements and does not introduce any difficulty ; greater lengths may require heavy cables, or the use of transformers in the speech circuits to raise their impedance to, say, 600 ohms, and an increase in the voltage which operates the control relays.

4.5.2. *Remote Control*

When a station is too remote from the control point for the provision of an extension cable, it is usually operated over either one or two pairs of telephone lines. The control unit at the operating point may include speech amplifiers for the microphone and receiver output circuits ; if a loudspeaker is used, an amplifier is essential, due to the limitation on the maximum A.F. power (5mW) which may be imposed on a telephone line ; if a low-sensitivity microphone is used, it must be followed by a pre-amplifier to raise the input to the line to 5 mW, in order to obtain maximum signal/noise ratio at the far end of the line.

In a simplex system, a single-pair line is required and the speech circuits are switched at both ends of the line, a relay being used at the remote end. A second relay is required at the remote end to switch the power supply to the equipment on and off. The two switching circuits can be carried on one pair of lines, either by using relays with different operating currents, or by using one line and an earth return for each circuit.

With a duplex system, the simplest method is to use two pairs of lines, one for inward and the other for outward speech, thus avoiding the need for hybrid circuits. If only a single pair is available, or if connection to a telephone exchange is required, hybrid circuits must be employed.

Whenever connection is made between radio equipment and a Post Office line, special precautions are necessary to conform with Post Office safety regulations ; items particularly affected are the design of transformers and relays.

4.6. *Radio Link Equipment*

For linking a satellite main station to headquarters in a multi-carrier system, or for the remote control of a station when a landline is not available, use is made of a V.H.F. or U.H.F. radio link. This requires two transmitters and two receivers in duplex operation. In addition, a trigger unit, which may be incorporated in the link receiver, is required to switch on the main transmitter when a message is sent over the outward link. A second trigger unit could be used to switch on the remote link transmitter when a message was received on the main receiver, but it is more usual to leave this link transmitter permanently switched on. Both link receivers are also permanently in operation, but the carrier of the link transmitter at the control station is switched on only when a message is to be sent, as this carrier controls the operation of the main transmitter.

The link transmitters can be similar to the low-power transmitter, discussed in Section 4.3, except for such modifications as are necessary for the difference in operating frequency. The link receiver may be somewhat simpler in design than the main receiver discussed in Section 4.2 ; the chief simplifications result from the less stringent requirements for sensitivity, A.G.C. and muting. On the other hand, the signal/noise ratio must be as high as possible and the fidelity of the complete link system should be good, compared with that of the main system.

Directional aerials should be employed in order to obtain the best signal/noise ratio and to reduce both the radiation in unwanted directions and the reception of unwanted signals or other interference.

5.0. Description of a Particular Range of Equipment

To illustrate some of the points which have been discussed in the previous sections, a brief description will be given of a commercial range of equipment, for the design of which the author has been largely responsible ; the more interesting features will be described in greater detail. Most of the items were originally developed in 1946-7 and a description of an early model of the mobile set has previously been published[6]. Many changes have been made since it was first designed, as operational experience has been gained and as fresh requirements have arisen. Although a new design would differ considerably in many respects, this range is still in full production for the police and many other services, and has the advantage that it has now been thoroughly proved in a wide variety of applications.

Amplitude modulation is employed, and models are available for any frequency between 27 and 125 Mc/s, each set operating on a single crystal-controlled frequency.

5.1. *Mobile Set*

5.1.1. *General Description*

The various parts which comprise a mobile station are illustrated in Fig. 1. The set itself consists of four units, receiver, transmitter, speech amplifier and power unit, mounted together on a shock-absorbing cradle ; this assembly occupies a space of 1 cu. ft. and weighs about 40 lb. The aerial is a quarter-wavelength flexible rod, attached to a flexible rubber base for mounting on the roof of a car. The other items are a moving-coil hand microphone and a control unit, which contains the loudspeaker and the control panel in a single box, for mounting under the dashboard ; as an alternative to the latter, there is a small control unit with a separate loudspeaker. For marine installations, the radio set is assembled in a flat metal cupboard, which can be attached to a bulkhead and projects only about 6 in. ; a special waterproof control unit and loudspeaker are employed for installation on the bridge.

5.1.2. *Receiver*

This contains 11 valves in a single superheterodyne circuit ; the arrangement is shown as a block diagram in Fig. 2. The R.F. amplifier employs a high-slope pentode (EF91), coupled to the mixer (EF91) by a parallel-fed π network. The multiplier anode circuit is coupled to the control grid of the mixer by a small capacitor. The R.F. circuits are illustrated in Fig. 3.

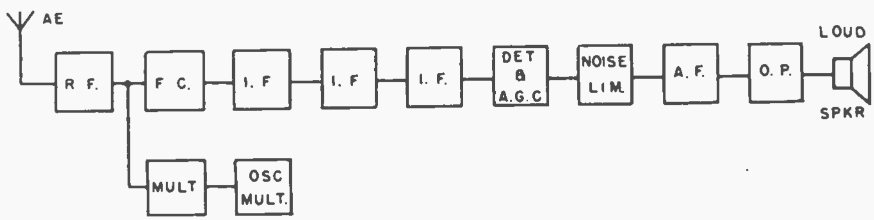


Fig. 2.—Receiver Block diagram.

The intermediate frequency is 4·5 Mc/s and four pairs of coupled circuits are used. The amplifier was originally designed with a pass-band of 50 to 60 kc/s, to be suitable for police multi-carrier systems, but this model was found to be insufficiently selective for use in the " business " frequency bands, with a channel spacing of 50 kc/s, so a new design was produced with greatly increased selectivity. The selectivity curves of the two models are reproduced in Fig. 4.

The detector and A.G.C. stage uses a double-diode valve. Full A.G.C. is applied to the first two I.F. amplifiers, which use variable-mu valves, and a fraction of the A.G.C. is applied to the R.F. and mixer stages. The time constant of these circuits is less than 20 milliseconds so that the receiver can follow rapidly fading signals.

The standard noise-limiter circuit employs a combination of series and shunt diodes, which limits impulsive interference to approximately 80 per cent. modulation at any carrier level. An improved circuit[7] has recently been developed, which provides carrier-controlled differential limiting by the introduction of a differentiating circuit and clamping diode before the series-shunt diode limiter ; an integrating circuit is included in the A.F. amplifier following the limiter. Post-detector A.G.C. is applied to this A.F. amplifier from a point in the limiter circuit at which a carrier-derived negative voltage, free from A.F. signals, is present. The two limiter circuits are shown in Fig. 5.

The output stage uses a 6V6GT valve and

delivers 2 to 3 W A.F. to the loudspeaker. The overall performance figures for the receiver are 1·5 to 2 μV sensitivity for 50 mW output, signal/noise ratio 10 dB at about 2 μV input, and noise factor approximately 10 dB. A preset control is fitted on the front panel and can be used to reduce the gain of the first I.F. stage when full sensitivity is not required.

The H.T. voltage supplied to the receiver is about 230 V and this is further reduced by the decoupling circuits for most of the valves to about 200 V. The total current drawn from the battery for the receiver is slightly over 5A at 12 V.

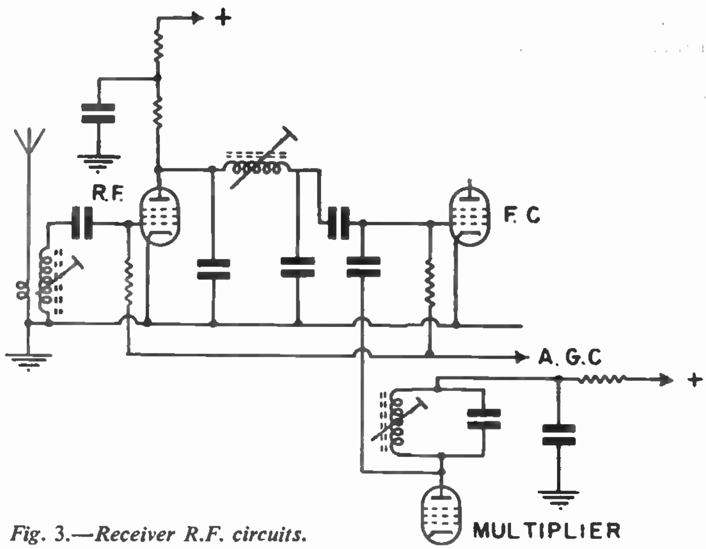


Fig. 3.—Receiver R.F. circuits.

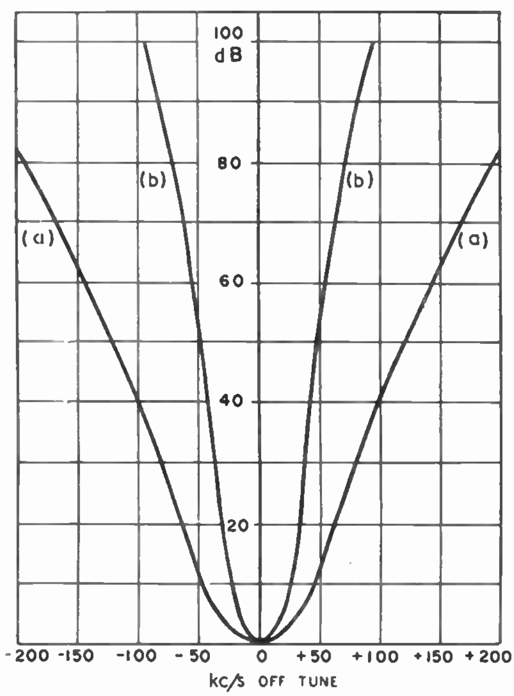


Fig. 4.—Receiver selectivity curves.

5.1.3. *Transmitter and Modulator*

The transmitter unit contains only the R.F. circuits and the modulation transformer. It uses six valves in four stages, the first of which is a crystal-controlled oscillator-multiplier and the second another frequency multiplier. This is followed by an amplifier, employing two EF91 valves in parallel, which drives the power amplifier, comprising two QVO4/7 valves in parallel. The R.F. output is approximately 12 W. The original output arrangement consisted of a single tuned circuit with the aerial connection taken from a tapping on the coil, but this gave excessive second harmonic output. A push-pull output stage was tried, but this produced only a small reduction in the second harmonic, gave no increase in output power, and introduced the need for carefully matching the drive to the two valves. A second tuned circuit, coupled to the tank circuit, was then tested and this gave a very satisfactory improvement and has recently been put into production. Fig. 6 is a block diagram of the transmitter circuit and Fig. 7 gives details of the two output circuits.

The speech amplifier consists of a push-pull pre-amplifier, resistance-capacitance coupled to a push-pull output stage. The latter employs two 6V6GT valves, operating in class AB_1, and delivers 10 W A.F. to the output transformer. The output is fed at low impedance into the power unit, where it is switched by a relay, either to the public address loudspeaker or to the transmitter modulation transformer.

5.1.4. *Control System*

All normal operation is effected by two switches on the control unit and by the press-to-talk switch on the microphone handle. A three-position switch adjusts the loudspeaker volume and a selector switch operates relays in the power unit to switch on the units which are required. By means of the latter it is possible to use the receiver alone for a listening watch, or to switch on the transmitter and modulator heaters in addition to the receiver, so that the transmitter is ready for immediate use. It is also possible to use the speech amplifier for public address purposes, with or without the receiver in operation.

The H.T. supply for the simplex equipment is obtained from two rotary transformers. One is switched on with the receiver heaters and its output is transferred to the transmitter when the microphone switch is pressed for transmission; the other supplies the speech amplifier and is switched on when the microphone switch is presssed, either for transmission or for public address. In the duplex model, there are three rotary transformers, one each for the receiver, the transmitter and the speech amplifier.

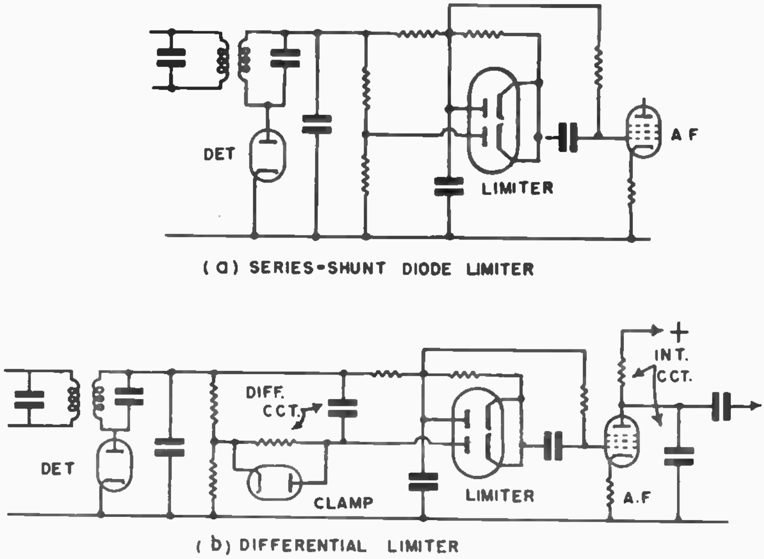


Fig. 5.—Limiter circuits.

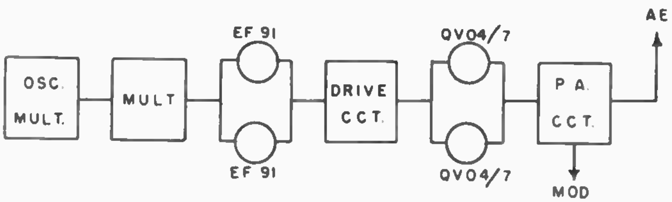


Fig. 6.—Transmitter block diagram.

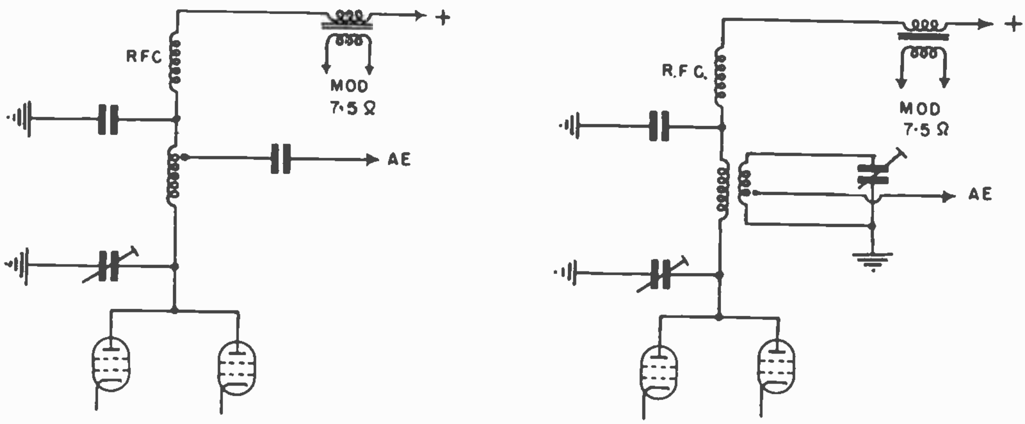
A miniature aerial relay is fitted in the simplex transmitter and is operated when the microphone switch is pressed for transmission.

5.1.5. *Calling System*

The original design made no provision for any type of calling system, but a simple method has recently been developed for inclusion in the equipment when required. Owing to the extreme compactness of the mobile receiver, there was no room to add any extra parts to it and, in fact, the only unit to which additions could conveniently be made was the combined control and loudspeaker unit. At the fixed station, the alterations were confined to a small modification to the transmitter. The system operates in the following manner. The fixed transmitter is fully modulated by a low-frequency tone when a push-button on the panel is pressed. At the mobile station, the A.F. input to the control unit is switched to the calling circuit instead of to the loudspeaker ; the signal on this line, which has an impedance of 3Ω, is stepped up to a high impedance by a transformer, the secondary of which is tuned to the frequency of the calling tone. The secondary voltage is rectified and the resultant D.C. operates a relay, which completes the circuit of a bell or buzzer, energized from the battery. When a call is received, the loudspeaker is switched on and the calling circuit is disconnected, either by a toggle switch on the control unit, or by a switch built into the microphone rest and operated when the microphone is picked up. While this method does not provide an automatic selective-calling system, it does permit a limited variety of manually-coded calls to be made. Its main advantage is that nothing at all is heard on the loudspeaker between calls.

5.2. *Fixed Station Equipment*

In low-power stations, the receiver, transmitter and control panel (if fitted) are assembled in a small rack-type cabinet, an example of such an equipment being illustrated in Fig. 8. In high-power stations, the transmitter occupies a separate cubicle.



(a) SINGLE CIRCUIT (b) COUPLED CIRCUITS.

Fig. 7.—Transmitter output circuits.

The original designs of the receiver and low-power transmitter were completed early in 1947, but various improvements, mainly in the receiver, were made later in that year. The high-power transmitter is the latest addition to the range and was not completed until mid-1948. The various types of control equipment have been developed to meet different requirements as they have arisen and their range is continually being extended.

5.2.1. *Receiver*

This consists of a 16-valve receiver chassis and an A.C. power-pack chassis, both mounted on a panel 19 in. × $8\frac{3}{4}$ in., which also carries a $6\frac{1}{2}$ in. diameter loudspeaker. Most of the circuit is similar to that of the mobile receiver, but it includes two R.F. amplifiers and the A.G.C. circuit is somewhat more complex. In addition, there is a new muting circuit[8], controlled differentially by carrier strength and noise, which can be adjusted to operate effectively at a given signal/noise ratio. In its most sensitive condition and in the absence of external noise, it opens on a signal input of about 0·7 μV. The circuit arrangement is shown in Fig. 9. In the absence of an incoming signal, the muting control is adjusted so that the muting diodes are just conducting ; these diodes, each in series with a large condenser, are connected between the A.F. amplifier and chassis, so that the amplifier is effectively short-circuited when the diodes are conducting. The control grid of the D.C. amplifier is biased negatively with an increase in carrier level and positively with an increase in noise level. If the resultant change in bias is negative the potential at the anode rises, so that the muting diodes cease to conduct and the muting circuit is opened. On the other hand, if the resultant bias becomes more positive, the muting diodes continue to operate.

The approximate overall performance figures for the receiver are 0·5 μV sensitivity for 50 mW output, signal/noise ratio 10 dB at 1 μV and noise factor 7 dB.

5.2.2. *Low-power Transmitter*

The transmitter and modulator are built on a common chassis, attached to a panel 19 in. × 7 in., which also carries an A.C. power-pack chassis. The circuit is similar to that of the mobile transmitter and modulator, except that the driver stage uses a QVO4/7 valve instead of two EF91 valves and the modulation transformer couples the modulator valves directly to the anode circuit of the P.A. valves. A meter is fitted on the panel and can be switched into each stage of the transmitter for test purposes and can also be used as a modulation indicator. A high-quality moving-coil microphone is normally supplied and provision is made for speech input from a 600Ω line ; a telephone handset can be used as an alternative when required.

5.2.3. *High-power Transmitter*

This is built in six units, assembled in a steel cubicle, measuring approximately 5 ft. high × 2 ft. wide × 1 ft. 8 in. deep. Each panel can be readily withdrawn for maintenance and the cubicle is fitted with lock-up doors. Forced cooling is provided by a blower unit at the bottom of the cubicle, with air ducts to those points at which most heat is generated. Interlock switches are provided to cut the mains supply when the rear doors are opened or when a panel is withdrawn. As the transmitter is designed for unattended operation, a comprehensive system of fuses and safety relays protects the equipment from damage in the event of a failure occurring.

Models are available for any single frequency between 27 and 185 Mc/s, the final stages employing resonant lines at the higher frequencies and LC circuits at the lower frequencies.

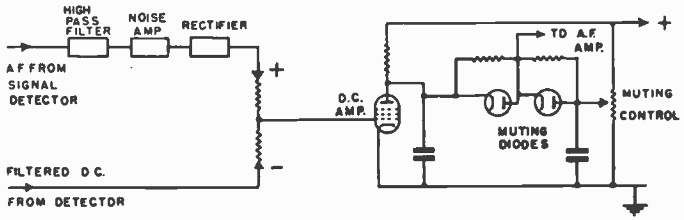


Fig. 9.—Muting circuit.

When extreme frequency stability is required, for example in multi-carrier schemes, a temperature-controlled crystal is fitted. The final stage employs two triodes in push-pull, which deliver 100 W R.F. to the aerial.

Anode modulation is obtained from a class AB_2 modulator and the speech amplifier is arranged to permit considerable flexibility for meeting different requirements. In the standard model, the frequency response is from 100 to 10,000 c/s and a speech clipper is included to prevent overmodulation.

5.2.4. *Control Equipment*

Any of the stations described can be operated through a cable from an extension control unit. This contains a loudspeaker, a volume control and a transmit-receive switch ; either a high-quality microphone or a micro-telephone handset can be used. The transmitter and receiver each contain an H.T. relay, operated from the 6·3 V A.C. L.T. supply and controlled by the transmit-receive switch. To prevent hum induction into the speech circuits, the leads for these in the cable are screened and separate earth conductors are used.

Fig. 10 shows the basic arrangement used for controlling a simplex station over a single pair of telephone lines. The two wires and the line transformers form a balanced circuit for speech currents ; one wire and earth are used for controlling the mains relay, which switches the power supply to the radio station ; the other wire and earth are used for controlling the transmit-receive relay.

For operating a duplex station over two pairs of lines, one pair is used for the microphone circuit and one of the control circuits ; the other pair is used for the receiver output and the other control circuit. If remote control of a calling device is also required, then one pair of lines carries two control circuits with an earth return, as in the two-wire simplex system.

When a radio station is to be linked to a telephone switchboard, the control equipment is made to suit the particular requirements of each case.

receiver and no de-emphasis is included ; the frequency response extends from 200 to 10,000 c/s.

A two-valve trigger unit is built into the receiver, when required, and this unit is also available on a separate panel. The A.G.C. voltage of the link receiver is amplified by a two-stage D.C. amplifier to operate a relay, which controls the main transmitter, or other equipment if required. Power supplies for this amplifier are obtained from the receiver power-unit.

5.3.2. *Transmitter*

This is identical with the main station low-power transmitter except for the following points :

(*a*) The frequency of operation is in the 144-156 Mc/s band.

(*b*) An additional frequency multiplier is included.

(*c*) The R.F. power output is reduced to 7 watts.

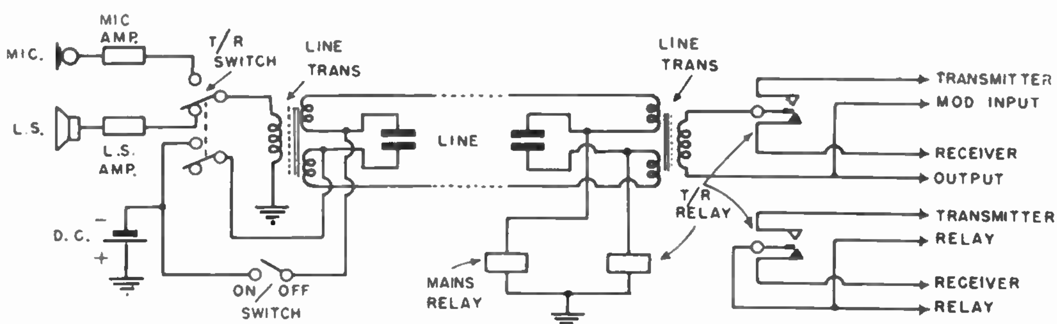


Fig. 10.—Simplex control circuit.

5.3. *Radio Link Equipment*

5.3.1. *Receiver and Trigger Unit*

The receiver is a simplified version of the main station receiver, adapted to work in the 144-156 Mc/s band. One of the I.F. stages and the whole of the muting circuit are omitted, since there is no need for very high sensitivity ; the A.G.C. circuit is also simplified. The signal/noise ratio is 30 dB at about 10 μV signal input. The overall fidelity is higher than that of the main

(*d*) The A.F. fidelity is better, particularly at high modulation levels, and the frequency response extends from 200 to 10,000 c/s.

6.0. Future Developments

In the immediate future, there will be a large demand for equipment to operate in the 160 Mc/s band, in addition to a continuing need for sets for the lower frequencies. It will be necessary to provide for both F.M. and A.M. systems, since neither appears likely to obtain a clear superiority over the other for all purposes for some time to come.

There is an obvious need for the development of simplified calling systems, not merely to isolate calls to a particular mobile station, but also to reduce interception of interfering signals and to provide a simple means of muting the receiver between calls.

Looking further ahead, it is logical to expect link circuits to be moved up to the U.H.F. band. Whether mobile systems can be used effectively on frequencies higher than about 200 Mc/s appears to be doubtful, but this may become necessary to a limited extent for non-essential short-range systems as the lower frequency bands become increasingly congested.

One of the factors which is tending to limit the use of radio-telephone systems is the cost of equipment. Unless some radically new developments occur, it does not seem likely that this can be reduced very much. As experience is gained in both the design and the production of this type of equipment, costs will tend to fall, but, to offset this trend, the performance requirements are becoming stricter as more services come into use; furthermore, the raising of the operating frequencies also increases the cost. Any attempt to lower prices by making inferior equipment would fail in the long run, since such sets would not only give unsatisfactory service to those using them, but would also be likely to cause increased interference with other systems.

7. Acknowledgments

The author is indebted to Messrs. Pye, Ltd., for permission to publish this paper. The equipment described in Section (5) is manufactured by Messrs. Pye Telecommunications, Ltd., and acknowledgment is also due to various of the author's colleagues who contributed to its development.

The author also wishes to thank Mr. J. R. Brinkley, of the Home Office Directorate of Communications, for much valuable advice on the design of equipment for multi-carrier systems.

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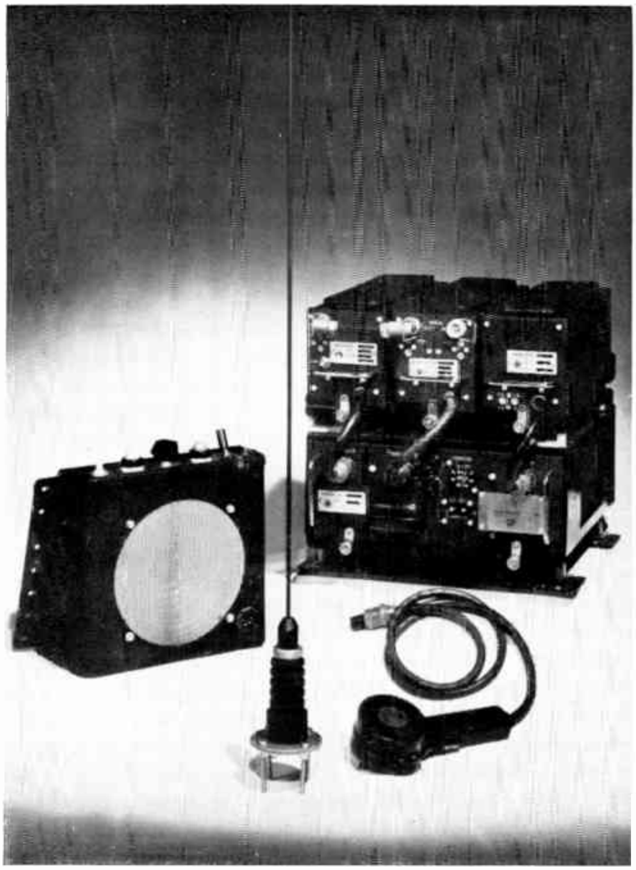

Fig. 1. Mobile Station.

Fig. 2. Low Power Fixed Station.