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

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THE FUTURE OF NATIONAL CERTIFICATES

Now that so much attention is being given to the establishment of a new Diploma in Technology, the significance of the Ordinary and Higher National Certificates in Engineering needs to be evaluated.

The National Certificate scheme was an invaluable and fundamental contribution to properly organized courses, nationally recognized and understood. Whilst the Electrical Engineering Certificate scheme has been in existence for 33 years, there is no comparable scheme in radio and electronic engineering; to meet an increasing demand, it is true that a number of colleges are now introducing radio and electronics subjects in their Higher National Certificate curricula, particularly in the A1 and A2 years.

One of the fundamental needs for the radio and electronics engineer is a good basic knowledge of physics; unfortunately physics at a suitable level for radio training is only taken as an optional endorsement subject. Much more important from the point of view of the radio profession is the fact that the inclusion of radio and electronics in the earlier stages of the course would allow a more advanced standard to be reached in radio and electronic engineering at the Higher level.

One interesting proposal made to the Institution is that physics should be introduced as a compulsory subject at Ordinary level. This would mean some modification of the existing compulsory subject of "Engineering Science."

Whilst some technical colleges have introduced an elementary study of thermionics in the final year of the Ordinary course, most students whose enthusiasm is directed towards radio and electronics feel discouraged by not having even started on more detailed radio engineering principles by the second year of their National Certificate course. Bearing in mind the comparatively small percentage of students who go on to the Higher National Certificate course, some alteration in the existing curriculum for the Ordinary National Certificate would provide an incentive qualification for the radio and electronics technician.

Even more important is the proposal that by including a greater radio content in the Ordinary National syllabus it would be possible to introduce a higher standard of radio engineering in the Higher National Certificate.

No doubt the newly appointed Council for Diplomas in Technology has benefited from the extensive experience gained in the administration of the Ordinary and Higher National Certificates. A particularly noteworthy advance is the decision to include external examiners in the panel who will be approved by the Council to conduct the final examinations leading to the Diploma in Technology. This will provide an opportunity to secure national uniformity of standard. A valuable feature of the present National Certificate schemes is the work done by assessing bodies in conjunction with the Ministry of Education; it is to be hoped that in any re-organization of technical education greater use will be made of organizations able to make useful contributions, particularly as assessors.

It is also to be hoped that, when details of the structure of the new Technological Diploma scheme are being formulated, due regard will be taken to the real status of radio and electronic engineering today. Both the extent of the industrial activity, and the wealth of scientific and engineering knowledge in this particular field, indicate that education and training will be retarded by taking the view that radio and electronics can still be classified as a sub-section of electrical engineering.

NOTICES

Annual General Meeting

The Council announces that the 31st Annual General Meeting of the Institution will be held at the London School of Hygiene and Tropical Medicine on *Wednesday, October 31st, 1956, at 6 p.m.*

The Agenda of the meeting will be published in the August issue of the *Journal*, together with the Council's nominations for vacancies about to occur in various offices, namely President, Honorary Treasurer, and ordinary members of Council; there will be six vacancies for ordinary members of Council, of whom four must be Full Members and two Associate Members.

"Not less than 21 days after the date of circulation of the Council's list, any 10 corporate members may nominate any other duly qualified person to fill any such vacancy by delivering such nomination in writing to the Secretary, together with the written consent of such person to accept office, if elected." (Extract from Art. 32 of the Articles of Association.)

The Annual Report of the Council will be published in the September issue.

Specimen Examination Papers

The November 1956 of the Institution's Graduateship examination will be the first held under the revised regulations. Specimen papers have been prepared for those parts of the examination where the syllabus has been extensively amended, and may be obtained on request from the Institution in London, price 2s. 6d.

If overseas candidates require the papers to be sent by second-class air mail, an additional 2s. 0d. should be included with the remittance.

A Museum of Sound Recordings

For many years there have been suggestions that a national, publicly accessible, collection of gramophone records should be set up, which would bear the same relationship to sound recordings as the British Museum Library does to the printed word. This library would have the object of ensuring the permanent existence of recordings of historic, artistic or scientific value.

During recent months, a British Institute of Recorded Sound has been set up for this purpose, under the Presidency of Viscount Esher, and legally incorporated as a non-profit-making company. Premises have been obtained at 38 Russell Square, London, W.C.1, which provide potential storage space for several hundred thousand records. The building includes a room for gramophone recitals and lectures, and sound-proof cubicles with equipment specially adapted for the reproduction not only of modern disks and tapes but also of the older "hill and dale" disks, and even cylinders.

The Institute is at the moment a privately sponsored body, and seeks support both in subscriptions and donations, and in gifts of obsolete recordings. Further information may be obtained from the Institute's Secretary, Mr. Patrick Saul, at the address given above.

Insurance Concessions for Research Students

A recent decision of the Ministry of Pensions and National Insurance means that research students will be treated as being in full-time education. Thus they will be excused from paying national insurance contributions.

Glasgow Radio Show

The Radio Industry Council has announced that it will hold a Scottish Radio and Television Exhibition at the Kelvin Hall, Glasgow, in mid-May, 1957.

It is expected that all the leading manufacturers of radio and television receivers will exhibit and there will also be exhibits of the latest high-fidelity sound equipment and of valves and components. The B.B.C. will also take part. The only previous exhibitions held in Scotland by the radio manufacturers were at Kelvin Hall in 1933, 1934 and 1935, although the retailers' organization has sponsored two shows in Glasgow since the war.

Expansion of I.T.A. Facilities

Low power test signals are now being radiated for test purposes from the site of the I.T.A.'s new television transmitter at Emley Moor, Yorkshire, preparatory to the opening of the station in late Autumn. The effective radiated power of the vision transmitter is 1 kW.

Additional sound and vision transmitting equipment is now installed at the Authority's Lichfield station as a standby in the event of breakdown. When certain ancillary equipment has been installed early in the Autumn, the Authority intends to raise the regular power of this transmitter from 50 to 200 kW effective radiated power.

SOME PROBLEMS OF SECONDARY SURVEILLANCE RADAR SYSTEMS *

by

K. E. Harris, B.Sc. †

Read before the Institution in London on March 28th, 1956
In the Chair: Mr. R. N. Lord, M.A. (Associate Member)

SUMMARY

This paper considers the system engineering of a secondary surveillance radar, comprising a ground interrogator and airborne transponders, for use in an air traffic control system. It deals in general terms with the choice of frequencies, considers how closely it should be associated with the primary radars now used for the control of air traffic and deals with other system problems as follows: the elimination of interrogations by the sidelobe of the ground radar radiation, the design of ground and airborne aerials, transponder saturation and methods of countdown, the difficulty of airborne units being "captured" by one ground station to the exclusion of others, second-time-round signals, unlocked responses due to remote ground stations, and the form of coding which may be used, with particular emphasis on the air-to-ground channel. The discussion is kept intentionally at the systems level so that the general considerations can be appreciated unhampered, so far as possible, by technical detail. It has nevertheless been thought necessary at some places to point the argument with sufficient detail to give some further insight into the problems involved.

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* Manuscript first received 24th February 1956 and in final form on 24th May 1956. (Paper No. 359.)
† Director of Research and Development, A. C. Cossor Ltd., Highbury, London, N.5.
U.D.C. No. 621.396.962.38.

1. Radar and Air Traffic Control

This paper deals with a radar assembly which may be used as an electronic aid in an Air Traffic Control System. The control system may be a military or a civil one: the way in which the system works, the job which it is required to do, may differ broadly in the military and civil applications; or if the civil control system alone is considered, the requirements may differ so greatly from place to place and time to time, that techniques which are apparently totally different may be developed to solve or ameliorate the local problems. The electronic or mechanical aids used in the control system, and the wits and abilities of the controllers, may never be adequate to provide what is really required of the system: hence much time and energy are spent in claiming universal merit for some gadget which is useful in some part of a particular control system. The author does not consider secondary surveillance radar or any other sort of radar to be essential to an Air Traffic Control System: it is a useful and perhaps extremely useful tool at the present stage of technical development. Equally, he does not consider human controllers to be essential to an air traffic control system: they too are useful tools at the present stage of development, but inevitably little by little they will be ousted by computer methods.

The functions of any air traffic control system are likely to include the following:

(*a*) To find out where all the "vehicles" are, whether they are in the air or not,

(*b*) To predict where they will be,

(*c*) To decide if therefore a potentially dangerous situation exists, and

(*d*) If so, to give instructions to eliminate the danger.

Radar and secondary surveillance radars deal only with a part of (*a*) and it is with this knowledge in mind that the following argument is presented.

2. Definitions

Primary radar locates objects by the reception of some part of the transmitted energy after it has been reflected by the object.

Secondary radar locates objects by the reception of signals which are automatically transmitted by a transponder at the object when it is interrogated by the locating radar.

A "Secondary Surveillance Radar System" is a system comprising ground and airborne elements which determines the positions of aircraft which are carrying airborne transponders and provides this information to the ground controller.

A "Dependent Secondary Surveillance Radar System" is one that uses a primary radar transmitter to interrogate the airborne transponder.

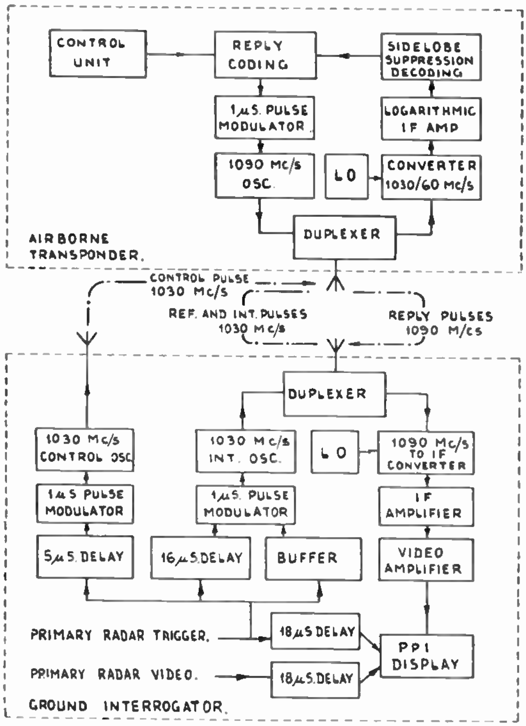


Fig. 1.—System schematic diagram.

An "Independent Secondary Surveillance Radar System" is one that uses a separate transmitter from that of the primary radar to interrogate the transponder. Fig. 1 is a schematic diagram of the ground and airborne equipments for a typical system.

Figure 2 shows the complete airborne equipment required for one such system. It comprises, from left to right, Control Unit, Transponder with its associated power supply unit and the airborne aerial. Fig. 3 shows the interior of the transponder.

The ground-to-air path is the "interrogate" channel and the air-to-ground path the "response" channel. An auxiliary ground-to-air or air-to-ground path used for sidelobe suppression is the "control" channel.

3. System Status—1956

In the past few years many authorities have stated that there is an operational need for a secondary surveillance radar system as an aid to air traffic control but to date no such system has been put into widespread service. One reason for the delay is that aircraft equipped with transponders will need to use them in conjunction with ground equipments which are located in several countries. International agreement has not yet been reached on a universal system. Early systems developed in the United Kingdom and the United States used primary surveillance radars as interrogators. Problems such as "sidelobe interrogation," "aerial polarization" and "second-time-round signals" were encountered with such systems. Some of the solutions proposed for these problems gave rise to further difficulties such as "capture," transponder saturation and increased complexity of the airborne and ground equipments. More satisfactory solutions to these problems result when a separate ground transmitter is used for secondary radar interrogation. The operating conditions of the secondary radar system can then be chosen for optimum secondary radar performance. A civil secondary radar system with an independent interrogator has been under development and trial in the United Kingdom since 1951 and a similar system for civil use has been under evaluation for two years or so in the U.S.A. Military I.F.F. techniques will be familiar to many readers of wartime history and current text books.

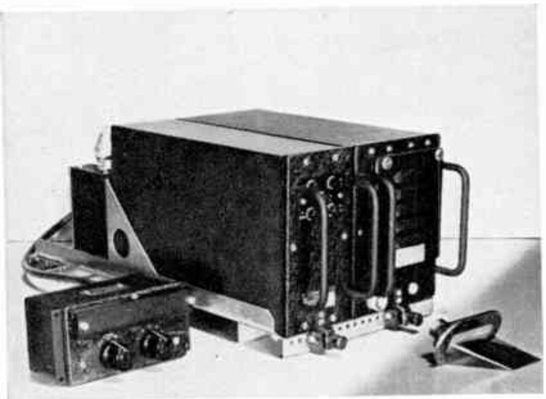


Fig. 2.—An airborne control unit, transponder with associated power supply unit, and airborne aerial for use in secondary surveillance radar system.

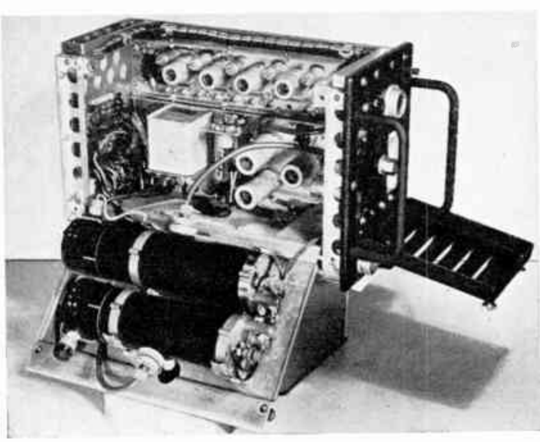


Fig. 3.—The T-R unit interior.

4. Advantages and Disadvantages of Secondary Surveillance Radar

In the last ten years primary surveillance radar has been extensively applied as an aid to air traffic control at large civil airports. For this purpose primary radar has its limitations, some of which can be overcome if a combination of both primary and secondary radar is employed.

The major operational requirements for Terminal Area and Long Range radars at civil airports are:

(*a*) To provide for continuous tracking by the controller of all types and sizes of aircraft under all conditions of weather and visibility.

(*b*) To provide identification of aircraft without the need to request special manoeuvres.

(*c*) To provide accurate determination of the geographical position of the aircraft in azimuch and distance and of relative aircraft positions.

(*d*) To provide information on relative heights of aircraft.

(*e*) To reveal areas of adverse weather to permit aircraft to be guided away from or advised of such areas.

(*f*) To provide tracking of aircraft at short ranges in regions of strong ground clutter.

4.1. *Coverage performance of primary and secondary radar systems*

The order of coverage at present required is 60 and 150 nautical miles for terminal area and long range radars respectively at altitudes up to 50,000 feet on all types of aircraft. The performances of available primary surveillance radars fall considerably short of this requirement and it will probably be some time before a simple and reliable primary surveillance radar of the required performance is in widespread use. Longer range performance, particularly on small aircraft, can more readily be given with a secondary radar system than by primary radar. Transponders used to improve radar coverage in this way have been called "radar assist beacons" in the U.S.A.

4.2. *Aircraft identification*

Primary radar alone does not provide sufficient information to identify the replies from a particular aircraft on the primary radar display without the need to request special manoeuvres. Auxiliary aids such as v.h.f. direction-finders and radio beacons have been used to assist identification but they are unreliable when two aircraft occupy approximately the same plan position. Secondary radar can considerably simplify the problem of identification by means of coding the reply from the transponder. In the simplest case, in response to a request on the radio-telephone, the secondary radar replies from the aircraft can be modified in a way that is distinctly visible on the display of an associated system. In a more complex case the identity of every aircraft may be ascertained without the need for radio telephone request.

4.3. *Positional accuracy*

The slant range accuracy of a radar system depends upon the pulse width and the accuracy of the display calibration. The pulse widths of an associated primary and secondary radar system may be of the same order so that, provided the variation of delay in passing through transponders is negligible, the accuracy of range measurement will be the same for both primary and secondary radar.

The bearing accuracy of a radar system depends amongst other things upon the horizontal aperture of the ground aerial and the operating frequency. The majority of existing and projected primary surveillance radar equipments operate in the S- or L-bands* and have a horizontal beam width of between 1 and 2 degrees. Discrimination of this order has been found to be necessary by controlling staff: it is not easy to provide it in the secondary system because of conflicting claims for low pulse recurrence frequency and high rates of rotation.

4.4. *Height determination*

Primary radar alone has been used to determine the heights of aircraft but equipments at present available are not sufficiently accurate and flexible for air traffic control purposes. Secondary radar can be used to report the readings of the aircraft altimeter to the ground station by means of suitable coding of the transponder replies. Several methods of applying height coding have been proposed but the equipment required is relatively complex.

4.5. *Weather information*

Primary radar is being used to observe precipitation as an aid to directing aircraft away from regions of severe turbulence. The information obtained in this application has been found to be very useful but not, unfortunately, totally reliable. It neither follows that turbulence is present where precipitation is observed, nor that turbulence is absent where no precipitation is observed. A secondary system contributes no help in this function.

Precipitation can have a serious masking effect on radar signals and can cause a considerable reduction in the range performance of a primary radar system. The effect is progressively more pronounced in the S- and X-bands than in the L-band. Since only one-way attenuation is involved, the effect of precipitation on the performance of an L-band secondary radar system is negligible in comparison with the effect on the S-band primary radar system.

4.6. *Freedom from clutter*

Moving target indicator methods have been applied with some success to eliminate stationary ground clutter from primary radar displays. These methods have their own shortcomings, limited "sub-clutter visibility," "velocity gaps" and "phase gaps" for example, and they are not always effective in removing precipitation. Secondary radar provides a clutter-free display without recourse to M.T.I.

* L band: 390-1,550 Mc/s; S band: 1,550-5,200 Mc/s; X band: 5,200-10,900 Mc/s.

methods, but on the other hand it gives no indication of the presence of aircraft if they are not carrying transponders or if the transponders are unserviceable.

Such considerations strongly suggest that both primary and secondary radar are required together to give maximum radar information in a control system.

5. System Design Problems

It is now proposed to deal with a number of system design problems. These are:

(*a*) Choice of frequencies.
(*b*) Dependent or independent working.
(*c*) Associated or dissociated working.
(*d*) Sidelobe interrogation.
(*e*) Ground aerial design.
(*f*) Capture, the exclusion of service to some ground stations by others.
(*g*) Second time round signals.
(*h*) Unlocked interference.
(*i*) Transponder saturation.
(*j*) Airborne aerials.
(*k*) Coding.

5.1. *Choice of Frequencies*

The choice of operating frequencies for an independent secondary surveillance radar system depends upon the international allocation of frequencies, the availability of suitable components, and the required horizontal definition of the radar system.

The I.T.U. (International Telecommunication Union) Atlantic City 1947 Conference allocated only three bands specifically for secondary surveillance radar use: 3,246 to 3,266 Mc/s, 5,440 to 5,460 Mc/s, and 9,300 to 9,320 Mc/s. The bands 960 to 1,215 Mc/s and 1,600 to 1,660 Mc/s are allocated world-wide for aeronautical radio-navigation. The bands 960 to 990 Mc/s and 1,186 to 1,215 Mc/s have been exploited for Distance Measuring Equipment. In the United Kingdom 1,600 to 1,660 Mc/s has been exploited for radio altimeters which sweep through the whole band. The band 1,300 to 1,600 Mc/s is allocated for aeronautical radio navigation in regions 2 and 3 (N. and S. America, and E. Asia and Australasia, respectively) and for fixed and mobile services in region 1 (Europe, W. Asia and Africa). However, a note in the table of frequency allocations envisages the possible future world-wide application of the band 1,300 to 1,600 Mc/s for an integrated system of electronic aids to air navigation and traffic control.

The bandwidths of the ground and airborne i.f. amplifiers must be great enough to compensate for the drifts of simple local oscillators and transmitting oscillators. In general terms, as the operating frequencies are increased from 1,000 Mc/s towards 9,000 Mc/s, the following factors affect the design of the ground and airborne equipments:

(*a*) Oscillator efficiencies decrease.
(*b*) Oscillator frequency drifts increase.
(*c*) Transmission line losses increase.
(*d*) Polar patterns of airborne aerials become less uniform.
(*e*) Vertical coverage diagrams of ground aerials of small vertical aperture become less uniform.
(*f*) Receiver sensitivities decrease due to the wider bandwidths of the i.f. amplifiers which are required to overcome oscillator drifts.

It can be seen then, that to keep the airborne and ground equipments as light as possible one should choose the lowest possible frequencies for both interrogate and reply channels. With available components the given factors are not very significant over the bands 960 to 1,600 Mc/s, but beyond 1,600 Mc/s the oscillator problems become more difficult to solve.

Operationally it is desirable that the horizontal beamwidth of the secondary system shall not be worse than that of the primary radar with which it is to be associated. Two factors militate against this. Firstly, the additional range to which the secondary system will operate demands a lower pulse recurrence frequency to avoid second-time-round problems. Unless the primary radar designer has left a considerable factor of safety in the number of pulses per beamwidth, or unless a slower rotation rate is tolerable, this is likely to demand a doubling of the secondary over the primary beamwidth. Secondly, if the secondary aerial is to be mechanically associated with the primary scanner, it will not often be permissible to make the horizontal aperture of the secondary aerial much greater than that of the primary. If the primary radar is S-band and the secondary radar is L-band there will therefore be another factor of three times greater beamwidth of secondary over primary radar. Taking both these points together, the secondary system may be forced to have a beamwidth five or six times greater than the primary radar: this in turn may demand separate primary and secondary display units. Altogether this consideration demands a frequency as high as possible for the secondary radar.

On balance the International Civil Aviation Organization (I.C.A.O.) have chosen 1,030 Mc/s ± 8·5 Mc/s as the interrogating frequency, and 1,090 Mc/s ± 8·5 Mc/s for the response. This choice permits a tolerable solution of the technical points listed above, but fails to satisfy the operator's demand for high resolution. This decision rests on the assumption that secondary surveillance radar will not be used as a traffic control tool in the absence of primary radar, except possibly to solve special low traffic density problems. The prospect of transponder failure and the presence in the control area of unequipped aircraft strengthens this assumption.

5.2. *Dependent or Independent Working*

At first sight it would appear an advantage to use the narrow beam of the primary radar and the existing transmitter as an interrogator. When the widely divergent primary radars which are used for air traffic control purposes are considered it becomes obvious that it is impracticable to design a transponder to work with them all. They differ in frequency, in power output, in receiver sensitivity, in polarization, in pulse length, in beamwidth and in recurrence frequency to an extent which makes a universal transponder impossible. To use the primary radar as an interrogator of an internationally standardized transponder would demand international standardization of the primary radar.

Even if this were attempted it would result in an unpleasant compromise between primary and secondary requirements. For example the primary radar recurrence frequency may need to be chosen for optimum M.T.I. working, the secondary p.r.f. needs to minimize second-time-round problems; the power in the primary transmitter must be adequate to deal with the fourth-power law and the target scattering factor in the radar equation and it may be some thousand times greater than that required to work the double-path square-law secondary system; the secondary interrogator may use multiple pulses to give security or freedom from unwanted interrogations on the ground-to-air path, the primary radar must rely on a single pulse at each pulse interval; the primary radar may wish to discriminate against rain echoes with circular polarization, the secondary system prefers vertical polarization.

Thus there is an overwhelming case for "independent" working which was defined (Sect. 2) as the use of a separate transmitter from that of the primary radar. The arguments above also show that an "independent" aerial is also necessary.

5.3. *Associated or Dissociated Working*

In some respects there is an advantage in mounting the secondary radar aerial on top of the primary scanner and in using a common plan position indicator, but there are also advantages in separate scanners and in separate displays.

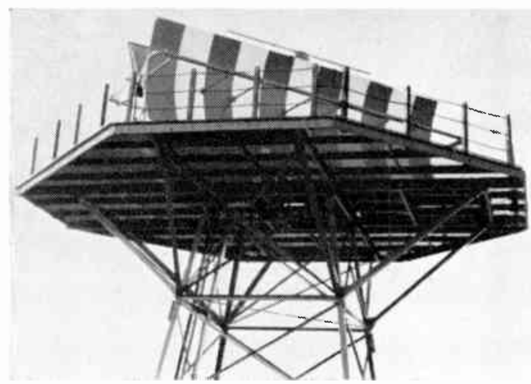


Fig. 4.—Secondary surveillance radar aerial on top of the MEW scanner at London Airport. A detailed view is given in Fig. 17.

5.3.1. Factors to be considered with "associated" working

(*a*) If the secondary radar aerial is associated with a primary radar scanner, no additional turning gear is required. There may be problems in finding enough slip rings, and the additional load and windage, though small, may still be the last straw.

(*b*) With two aerials mechanically mounted together there results one difficult siting problem in place of two independent ones, each already difficult enough. The two together may result in a compromise completely satisfactory to neither: on the other hand, it should be remembered that the one problem, although more difficult, may be soluble where the two independent ones are not—particularly in the environs of a modern airport where the finding of two sites at reasonable distances from the display centre is often almost impossible irrespective of their suitability for radar scanners.

(*c*) Complete "association" reduces the number of displays required and keeps all the radar information in one place for each controller, thus saving space and reducing the possibility of "liaison" errors. In fact the amount of additional equipment required to provide the complete secondary system can be provided in a two-foot cube box (to give an approximate notion of its size), besides which the only other requirement is the aerial.

5.3.2. Advantages of separate displays

(*a*) With separate displays the shortcomings of each, e.g. rain clutter and permanent echoes on the primary radar and unlocked responses on the secondary radar display, do not detract from the other. The resolution of the primary system is not impaired by the greater beamwidth of the secondary: the sensitivity of the secondary system is not reduced by the addition of noise from the primary.

(*b*) If the displays are separate either can be serviced while the other remains working.

5.3.3. Advantages of separate scanners

(*a*) With separate scanners, although two turning gears are required, rotating mechanisms can be simplified. The associated system either demands a triple rotating joint or else that the secondary transmitter/receiver be rotated with the scanner. For this reason it is sometimes suggested that the scanners should be separate but that they should be electrically locked so that common displays can be used.

(*b*) Siting is independent with separated scanners. The siting of each system is difficult: the primary radar problem is well known; with secondary radar the problem of large plane surfaces, such as the sides of hangars, becoming radar mirrors is important.

(*c*) If separate scanners and displays are used the rotation rates can be optimized for each p.r.f. and beamwidth.

Although the technical advantages of separation are clearly great, and provisioning, installation, and maintenance considerations add their weight to the argument, the history of I.F.F., the reduction of the total amount of equipment and especially the number of displays, and the feeling that secondary radar may need to be "independent" but must not be allowed to be as independent as all that, have combined to make the systems now under evaluation "associated" ones. The possibility that in the future they may not be so is little comfort to the designer who is constrained, by the requirement to tie the two systems together, to accept a number of difficulties.

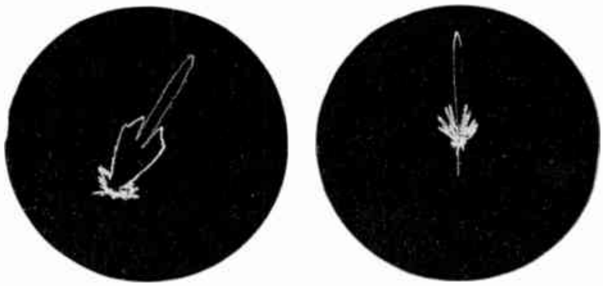


Fig. 5.—Interrogator aerial patterns.

The first of these penalties is the restriction in vertical aperture thereby imposed on the secondary radar aerial. If this aerial is to be a parasite on the primary radar turning gear it must evidently be light and add little "windage" as it rotates; its vertical aperture must therefore be small. Some of the resultant problems are considered in a later section, but it must be remembered that at 30-cm wavelength there is little hope from the outset of getting a vertical aperture sufficiently great to make such difficulties negligible. Fig. 4 shows a Secondary Surveillance Radar Aerial mounted on a MEW scanner at London Airport.

5.4. *Sidelobe Interrogation and Methods of Sidelobe Suppression*

The interrogator aerial of a secondary surveillance radar system is designed to produce a narrow beam in the horizontal plane. Imperfections in the design and siting of practical aerials permit the radiation of some power in other directions (see Fig. 5). At short ranges this sidelobe radiation is sufficient to interrogate an airborne transponder. Similarly, the reply from the transponder may be received upon a sidelobe of the receiver aerial pattern so that a reply may be displayed on the plan position indicator at a wrong bearing. At very

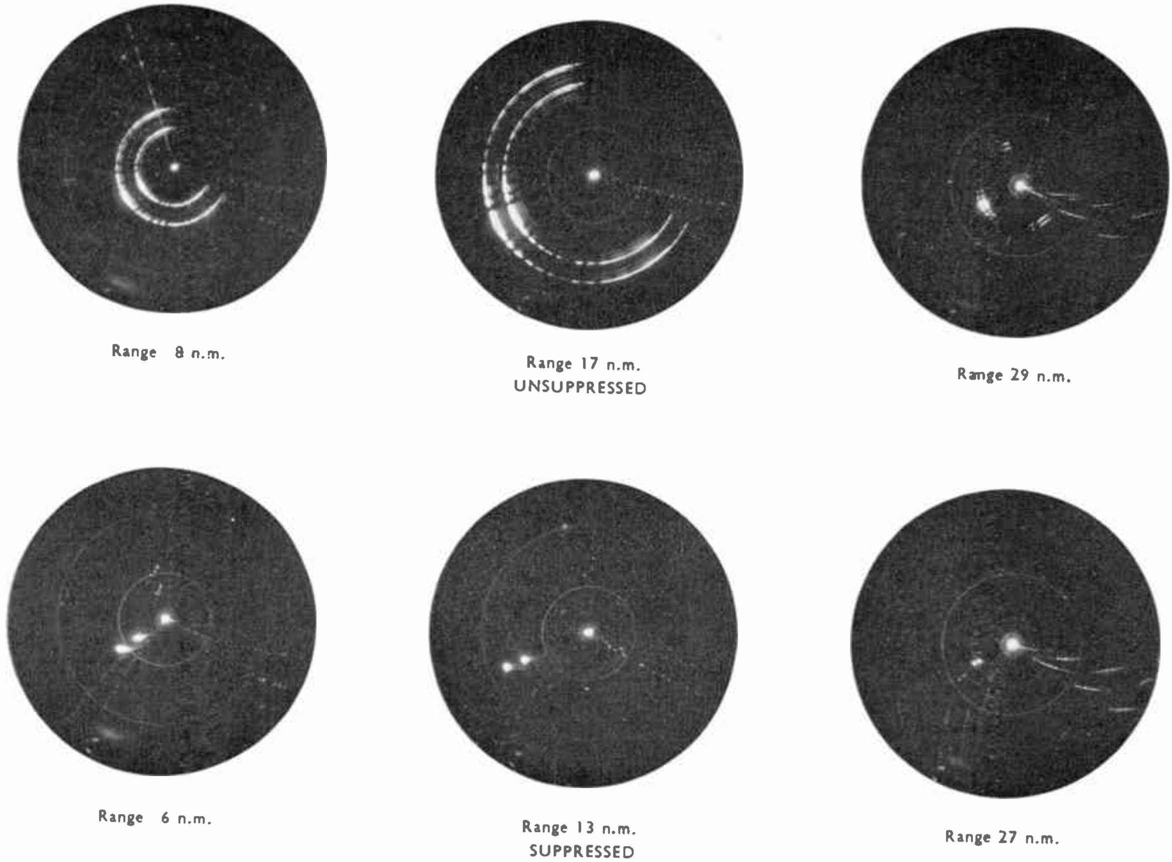


Fig. 6.—Typical secondary radar PPI responses showing the variation of ring-around with changing range. Double pulse reply code is shown.

short ranges interrogations by sidelobes occur in all horizontal directions causing a ring of responses to appear on the indicator so that the bearing of the aircraft is indeterminate. Fig. 6 shows typical p.p.i. responses with and without sidelobe suppression.

Suppose that it is required to design a secondary surveillance radar system which will give a reliable range performance of, say, 50 miles on aircraft which fly at elevation angles up to 30 deg. Sufficient power must then be transmitted to produce discernible responses from aircraft at a range of 50 miles when the system is operating under the most unfavourable conditions. It then follows that, at a given range under the most favourable conditions, the signal amplitude at the ground receiver exceeds the signal obtained under the most unfavourable conditions by the sum of:

(*a*) The variation of the airborne aerial polar pattern throughout the operational coverage,

(*b*) The variation of the vertical coverage diagram of the ground aerial at elevation angles up to 30 deg. at the given range,

(*c*) The variation of the ground receiver sensitivity under service conditions,

(*d*) The variation of the airborne transmitter power under service conditions.

Taking practical values at 16 db, 20 db, 3 db and 6 db respectively for each of these variables gives a possible 45 db increase of signal strength under favourable conditions. Thus sidelobe responses may be displayed from aircraft at a range of 50 miles in directions where the horizontal polar pattern of the ground aerial is not more than 45 db down relative to the peak of the main lobe. In addition, sidelobe responses may be displayed at a range of

5·0 miles in directions where the pattern is not more than 65 db down relative to the main lobe. It may not, of course, be possible in practice to transmit sufficient power to guarantee such performance, but the relative argument remains the same irrespective of the transmitted power. Although the practical range at which sidelobe responses are troublesome will be reduced as power is reduced, so also to the same degree will be the guaranteed "system performance."

It may be noted in passing that the 20-db allowance for the vertical coverage diagram of the ground station is one of the penalties for a small vertical aerial aperture.

There are two main ways in which the appearance of sidelobe responses may be removed from the radar display:

(*a*) Responsor suppression: the sidelobe transmissions are allowed to interrogate the transponder but the sidelobe replies are suppressed on the air-to-ground channel.

(*b*) Interrogator suppression: the sidelobe transmissions are prevented from interrogating the transponder by suppression on the ground-to-air channel.

Responsor suppression demands a higher mean rate of triggering of the airborne transponder than interrogator suppression so that saturation and interference problems are more serious with the former method.

5.5. *Responsor suppression*

Responsor suppression may be realized in either of two ways.

5.5.1. Sensitivity/time control

The sensitivity of the ground receiver is varied in accordance with the slant range of the aircraft from the ground station so that at short ranges the receiver passes only strong signals. For totally successful performance it is theoretically necessary that the sum of the following variables shall be less than the difference between the main lobe of the aerial and the strongest sidelobe which is to be suppressed:

(*a*) Variation of the airborne transmitter power from aircraft to aircraft and with operating conditions during flight.

(*b*) Variation of the airborne aerial polar pattern throughout the operational coverage.

(*c*) Variation of the ground aerial operational polar pattern throughout the operational elevation coverage of the system.

(*d*) Variation of the sensitivity/time control law and the ground receiver sensitivity from optimum with operating conditions between maintenance.

Taking practical values of 6 db, 16 db, 20 db, 6 db for each of these variables gives a possible total variation of 48 db whereas the operational horizontal pattern of a typical aerial may have many sidelobes less than 30 db below the main lobe. It is, of course, unlikely that all these conditions will be at one extreme at the same time, but it is quite probable that the occasional suppression of a main lobe signal or occasional sidelobe break-through will be found; thus the sensitivity/time control method will not be absolutely positive. The principal objections to it, however, are that it effectively modifies the vertical coverage of the secondary radar system and that, as stated above, it is more susceptible to saturation and interference problems.

From the point of view of simplicity this method is most attractive. It has also been demonstrated to be effective in practice in cleaning up the radar display when the setting up of the range gain law has been made by engineers fully understanding the conditions involved. It might prove to be less effective even in this respect if it had to be adjusted and maintained by normal station personnel having a wide range of equipments to service.

5.5.2. Cancelled reception method

The response from the aircraft is received upon two ground aerials, one directional and the other omnidirectional in the horizontal plane. The two aerials have similar vertical coverage diagrams. The aerials are connected to separate receivers which have non-saturating characteristics and video signals from the omnidirectional receiver are subtracted from those from the directional receiver. The relative gains of the receivers are adjusted so that a positive output is obtained only in directions contained within the main lobe of the directional aerial.

This method of sidelobe suppression suffers from the saturation and interference problems of the sensitivity/time control method, and is also subject to the following disadvantages:

(*a*) R.f. interference between the signals from aircraft at the same range affects the operation of the suppression circuits.

(*b*) Back edge elongation of the signal pulses, due to second path reflections, affects the operation of the suppression circuits.

(*c*) Higher airborne transmitter power is required for the same range performance as for an unsuppressed system, since the noise from the omnidirectional receiver adds to the noise from the directional receiver.

5.6. *Interrogator suppression*

Interrogator suppression may be achieved by the following methods.

5.6.1. The very long time constant method

The trigger level of the transponder is determined by pulse stretching the strongest signal which is received from the ground station. The time constant of the trigger level storage circuit is chosen to give a decay of about 3 db during the time of one rotation of the ground aerial. The time of one scan is 5 seconds for a radar operating at 12 rev/min so that the recovery rate of the trigger level is of the order of 3 db per 5 seconds.

The very long time constant method of suppression is satisfactory provided that only one ground transmission is received by the transponder. The problem of "capture" arises when the transponder is within the coverage of two or more ground stations. The main beam of the strongest of the interrogating signals then sets the trigger level of the transponder and service is denied to weaker interrogating signals. The transponder is then said to be captured by the stronger interrogating signals. It follows that one can consider volumes of air space around each ground interrogator in which an airborne transponder will reply exclusively to that station. In brief, the coverage of every ground station will contain dead zones due to other ground stations.

5.6.2. The long time constant method

The trigger level is determined by pulse stretching control signals which are received from a ground control transmission. The control transmitting aerial pattern is omnidirectional in the horizontal plane and is similar to the interrogating aerial in the vertical plane. The control and interrogating transmissions are coded differently and are not necessarily locked in time, as shown in Fig. 7. The time constant of the trigger level storage circuit is chosen to give a decay of about 3 db during the interval between successive control pulse transmissions. It is convenient to make the repetition time for the control transmissions of the same order as the interrogator, so that the recovery rate of the trigger level is about 3 db per 3 or 4 milliseconds. The amplitudes of the control and interrogating signals are adjusted so that triggering occurs only on the main lobe of the interrogating aerial.

The long time constant method suffers from capture but the control level can be set further down on the main lobe than in the very long time constant method so that the magnitude of the capture areas is not so great.

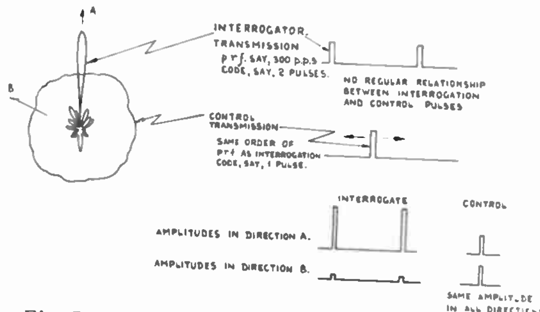


Fig. 7.—Control and interrogating transmissions.

5.6.3. The short time constant method

A control transmission is emitted that is closely related in time to the interrogator transmission as shown in Fig. 8. To take as an example a timing which is in use, the control pulse may be emitted 8 μsec prior to the emission of an interrogator pulse and both pulses may be 1 μsec wide. The trigger level of the transponder is set by pulse stretching the received control signals. It is arranged that the transponder is only triggered if the amplitude of the interrogator pulse exceeds the residual amplitude of the control pulse at the time of reception of the interrogator pulse. The recovery rate of the trigger level may be set to be 5 db per 7 microseconds so that the transponder is only triggered if the amplitude of the interrogator pulse exceeds a level 5 db below the amplitude of the control pulse. The recovery to full sensitivity, after reception of a pulse 60 db above the minimum trigger level, then takes place in 84 μsec provided that the characteristic of 5 db/7 μsec is maintained over a dynamic working range of 60 db. Thus the transponder is only captured for an interval of 84 μsec even when quite close to a ground station and for less time at greater ranges. Hence some twenty or more ground stations of average random pulse repetition frequencies about 250 per second can obtain service that is effectively free from capture.

The short time constant method is undoubtedly the best known means of sidelobe suppression that is suitable for a universal system. It is proposed therefore to deal in rather more detail with problems encountered in the application of this method.

5.6.4. The tolerance of the suppression level

The relative amplitude of the interrogator pulse to the control pulse may be defined as the "decoder suppression level" at the boundary condition where triggering of the airborne transponder is just suppressed. The decoder suppression level in a practical case is liable to vary between -1 db and -9 db when the control pulse precedes the interrogator pulse by 8 μ sec and the amplitude of the interrogator pulse lies between 0 and 60 db above the minimum trigger level.

The boundary amplitude on the control aerial horizontal polar pattern may conveniently be defined as the "pattern suppression level." The level at which this needs to be set is dependent upon variations of the decoder suppression level, relative variations of the operational vertical polar patterns of the control and interrogator aerials, and relative variations of the control and interrogator transmitter powers.

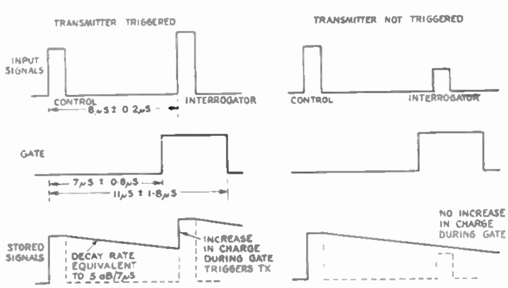


Fig. 8.—Two-pulse method of transmission.

It will be observed that for satisfactory working the statistical sum of these variations must be less than the difference between the peak of the main lobe and the highest amplitude sidelobes of the operational horizontal polar pattern of the interrogator aerial.

5.6.5. The control transmission

The gain of a directional interrogator aerial of beamwidth 5 deg. is about 17 db greater than that of an omnidirectional control aerial with the same vertical pattern. Thus, if the interrogator power is 1 kW and the control power is 17 db above 1 kW, i.e. 50 kW, it follows that the interrogator and control pulses received at a transponder in the direction of the main lobe, are of equal amplitude. The maximum range of the system is then equal to that obtained with the same transponder set to operate when it receives two interrogation pulses spaced by 8 μ sec transmitted by the 1-kW peak pulse power interrogator.

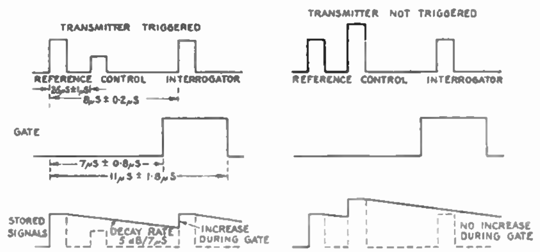


Fig. 9.—Three-pulse method of transmission.

The side lobes of a typical interrogator aerial may be everywhere greater than 20 db down relative to the peak of the main lobe so that it would be permissible to reduce the control transmitter power by, say, 10 db, i.e. to 5 kW, whilst leaving the interrogator power at 1 kW. The maximum range of the system in such a case is reduced, working as it is with a transponder requiring an 8 μsec spaced pair to operate it, and 10 db of surplus interrogator pulse power is then received by a transponder located at the maximum range of the system. The control pulse is required to be present at the maximum range of the system in order to operate the pulse pair decoding circuits of the transponder. It is, however, possible to emit a reference pulse from the interrogator transmitter prior to the interrogator transmission so that the reference and interrogator pulses operate the pulse pair decoding circuits and the control pulse is used solely to set the trigger level for the interrogator pulse. The control pulse is then preferably emitted in between the reference and interrogator pulses so that simple timing circuits may be used for the pulse pair decoding operation. This arrangement has been called the three pulse system because the transmission consists of two interrogator plus one control pulse, and the timing arrangements are illustrated in Fig. 9.

It is possible to design the transponder to operate with either the two pulse or three pulse method of ground transmission. For the three pulse method, the spacing between the reference and interrogator pulses can be 8·0 μsec and the control pulse may, for example, be emitted 2·5 μsec after the reference pulse. The decoder suppression level is between 0 and -6 db for the three pulse method of transmission. Thus, if the interrogator power is 1 kW and a pattern suppression level between -14 db and -20 db is acceptable, the control transmitter power is required to be only 3 db greater than 1 kW when emitted on an aerial of gain 17 db less than the interrogator aerial.

The power required for the control transmitter may be reduced if the control aerial pattern is somewhat directional in the horizontal plane and is rotated in synchronism with the interrogator aerial. The pattern suppression level then varies with azimuth and is required to exceed the sidelobe level in all horizontal directions. For example, suppose that the sidelobe level is more than 20 db down within directions ± 30 deg. relative to the main lobe, and more than 26 db down in all other directions. The pattern suppression level for the two pulse methods may be set at, say, 10 db below the main beam in order to avoid wastage of interrogator power, so that a reasonably efficient control aerial would have a pattern of maximum value extending over the main lobe of the interrogator pattern and falling to not less than -10 db at ± 30 deg. nor less than -16 db in all other directions. A practical aerial meeting this specification may have a gain of 10 db greater than an omnidirectional aerial.

The way in which the airborne transponder may be made to operate equally satisfactorily for either a two pulse or three pulse system can, for example, be as follows: The two input pulses or three input pulses are applied as positive signals on the grid of V3A (see Fig. 10). A positive signal on this grid causes its anode voltage to fall and X2 to conduct. V3A and X2 then form a low resistance path through which the capacitor C3 charges, the charging current being drawn from the h.t. supply and C21. V3A can be regarded as a cathode follower, or as a charging resistance of variable value depending on its grid voltage. C3 will charge to nearly the maximum value of the pulse and the fall in grid voltage at the end of the pulse will cut off the anode current in V3A because the cathode is biased by the capacitor voltage.

When anode current is cut off, C3 is isolated and starts to discharge through R11, and the voltage on C3 decreases linearly with time at a rate $V/C3.R11$ where V is the voltage across R11. This voltage rundown is equivalent to a rundown in the level of input signal at which the transponder will trigger and the rate used at present in practice is about 5 db/7 μsec.

In Fig. 11 idealized input signals A are drawn to the same time scale as the corresponding voltage waveforms B across C3. Two cases are of importance: in the first, illustrated by solid lines, the C pulse is greater than the I and R pulses (i.e. the I and R pulses are due to sidelobe signals). The R pulse raises the voltage across C3 to a level corresponding to peak R pulse amplitude and at the end of the pulse linear rundown takes place. This rundown continues until the C pulse with its greater amplitude again raises the level of the voltage and the rundown recommences from

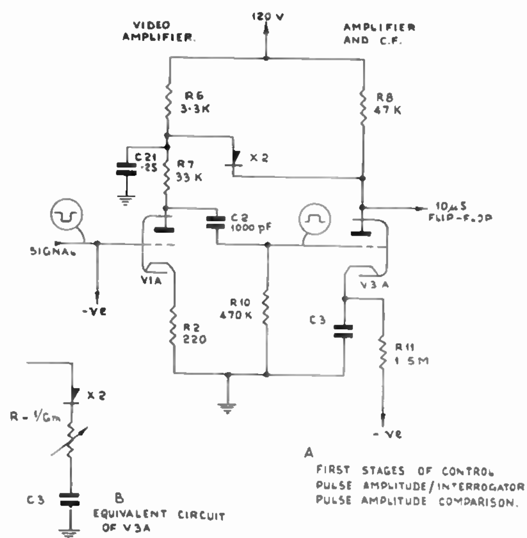


Fig. 10.—Decoder for two-pulse or three-pulse methods of ground transmission.

this higher level. When the I pulse occurs the voltage across C3 (the cathode voltage of V3A) is greater than the peak amplitude of the I pulse on the grid of V3A so that the valve remains cut off, the rundown of voltage across C3 continuing uninterrupted.

The second case, drawn with dotted lines, shows the state of affairs when the interrogator pulses are much larger than the control pulse. The R pulse raises the voltage across C3 to a level corresponding to the R pulse amplitude

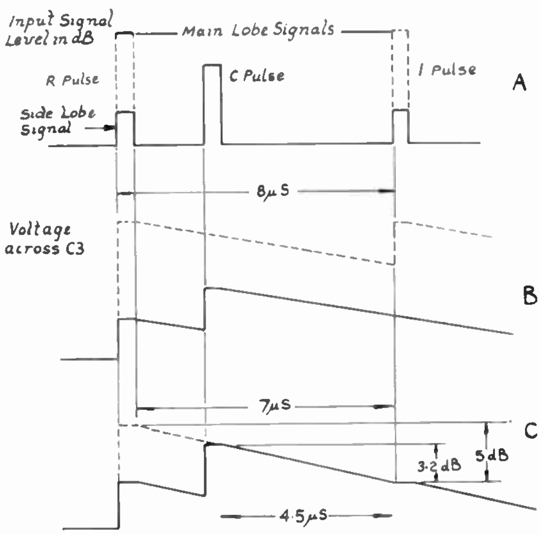


Fig. 11.—Waveforms of circuit in Fig. 10.

and a linear rundown of voltage takes place as before. This time, however, the C pulse amplitude is not sufficient to cause current to flow through V3A and interrupt the rundown which continues until the I pulse is received. This is the condition which results in the transponder being triggered.

Waveform C (Fig. 11) shows the limiting case in which the I pulse just fails to interrupt the capacitor voltage rundown, and it is seen that the relative levels are such that the C pulse must be 3·2 db greater than the I pulse. Similarly in a two-pulse system the C pulse must exceed the I pulse by 5·0 db. In practice these figures are subject to considerable tolerance: for a three-pulse system the C pulse amplitude can exceed I pulse amplitude by 3 db $\pm$3 db, and for a two-pulse system the R pulse amplitude can exceed I pulse amplitude by 8 db for triggering to be always suppressed.

Thus the circuit produces a positive pulse in the gate whenever the interrogator pulses are adequately greater in amplitude than the control pulse, and there is no pulse in the gate if this condition is not met. This applies equally for two or three pulse interrogation and evidently the resultant pulse in the gate can be caused to initiate the airborne transmitter sequence.

5.7. *Ground Aerial Design*

To obtain the required horizontal and vertical interrogation patterns within the limitations already discussed of small vertical aperture and horizontal aperture no greater than the associated primary radar is a matter of established technique. In practice sidelobes in the horizontal plane can be kept everywhere more than 20 db and in general 30 db below the main beam by design reproducible in production. To minimize the overall size of the aerial and sidelobe and backlobe radiation a broadside array of dipoles is preferable to a parabolic reflector particularly if the vertical aperture is reduced to an absolute minimum, i.e. to the height of a resonant vertical dipole.

5.7.1. Interrogator aerial patterns

The free-space horizontal pattern of the aerial depends upon the distribution of power to each element in the broadside array. Uniform distribution of power gives the narrowest beamwidth but the sidelobe level is high, the major sidelobes being only about 13·5 db below the main lobe. The sidelobes are reduced in amplitude when the power is progressively reduced from the centre towards the outer edges of the aperture and are completely eliminated when the elements are each fed in a binomial distribution of power. The pattern then falls progressively to zero in directions $\pm$90 deg. of the main lobe, and the beamwidth between half-power points is about twice that for a uniform distribution of power to the elements in the aerial. In practice, a compromise must be made between the desired sidelobe level, the beamwidth and the degree of complexity involved in procuring a given distribution of power over the aperture.

A typical production aerial in the system previously discussed, having an aperture of 15 feet and composed of 24 elements fed in a six-step distribution, exhibits a beamwidth of about 5 deg. as compared with 3½ deg. for a uniform distribution of power. The back radiation is more than 20 db below the forward radiation for elevation angles up to 30 deg. (see Fig. 12).

The operational horizontal polar pattern of

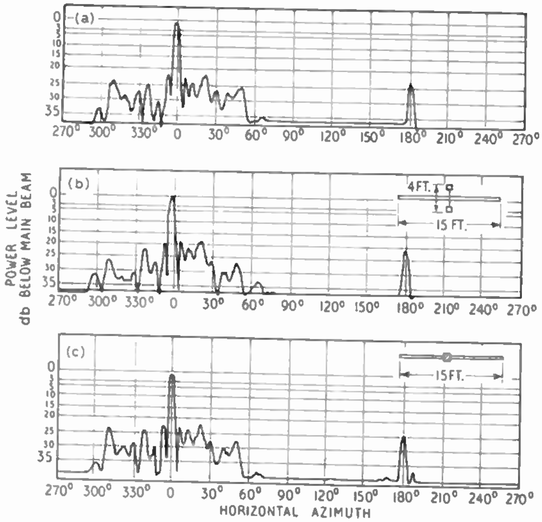


Fig. 12.—(a) Azimuth/amplitude diagram of an interrogator array of 15 ft. aperture with a gap at its centre for control aerials but no such aerials in position.
(b) As (a), but with forward and rear control radiators spaced 4 ft. apart.
(c) As (a), but with control radiators close together.

a secondary radar aerial exhibits more sidelobe radiation than the free-space pattern of the aerial because irregularities in the terrain surrounding the aerial cause scattering of some of the energy radiated below the radar horizon. Objects such as nearby buildings, hangars, hills and other aerials tend to make the practical sidelobe level worse than that measured from the aerial alone. The operational horizontal polar patterns at any site should be measured relative to an omni-directional test aerial located at various points around, and sufficiently distant from the site. Measurements made with a directional test aerial located in a direction free from irregularities may be unfairly optimistic of the situation since the directivity of the test aerial may discriminate against building and ground reflections.

The free space vertical pattern of a single dipole and reflector is broad in beamwidth and about half the power is emitted in directions below the horizon. The signal received at an aircraft is then the resultant of the direct path signal and ground reflected signals. Curves for radio propagation over a plane earth show that the reflection coefficient decreases from unity for small angles of incidence to about 0·3 at 10 deg. for vertical polarization and typical terrain with a dielectric constant of 10. Calculations show that destructive interference causes dips, of about 19 db, 13 db and 10 db below the direct signal amplitude, to occur at elevation angles of 1 deg., 2 deg. and 3 deg. respectively for an L-band aerial mounted 25 feet above the ground (see Fig. 13). The peak at 0·5 deg. is about 6 db above the direct signal amplitude so that the total signal variation is 25 db from 0·5 deg. to 1 deg. Sufficient transmitter power must be emitted to permit operation of the system on aircraft within the dips in the operational elevation pattern.

5.7.2. The control aerial vertical pattern

The operational vertical pattern of the control aerial is required to be similar to that of the interrogator pattern throughout the vertical coverage of the system. Ideally, this requirement can only be met by arranging that the free-space patterns on both aerials are identical and emanate from the same point in

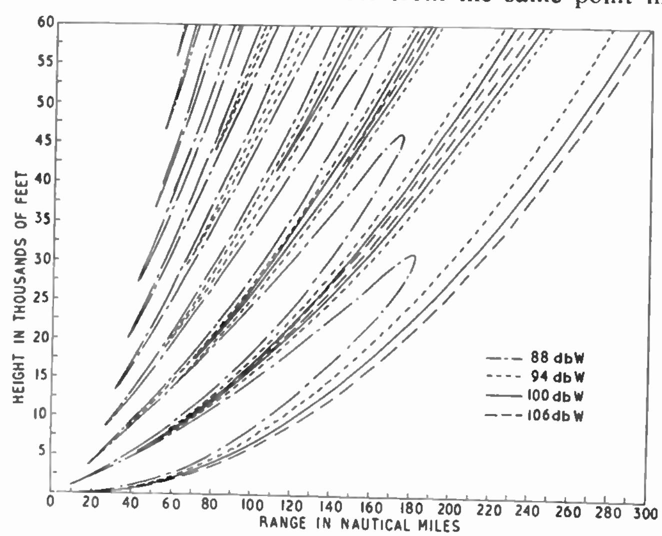


Fig. 13.—Pattern given by a radiator which is isotropic in the vertical plane of the diagram and is 25 ft. above a spherical earth. Powers on the curves are developed across the airborne aerial load for the following conditions:
(a) Ground power transmitted = 1 kW.
(b) Ground aerial gain over isotropic radiator = 20 db.
(c) Airborne aerial gain over isotropic radiator = 1 db.

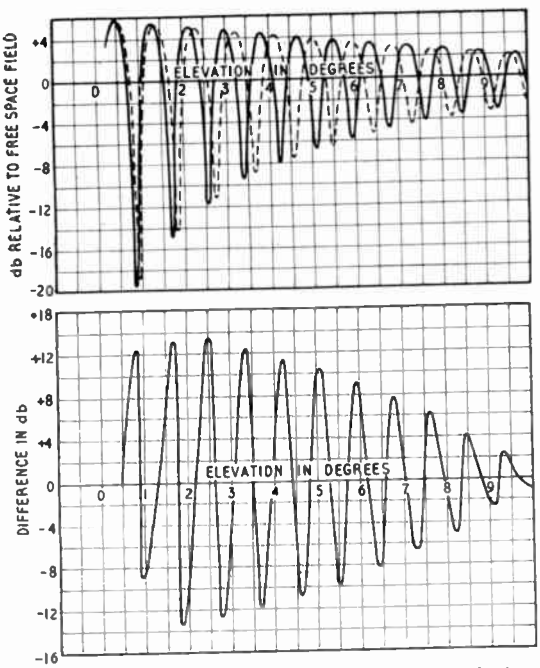


Fig. 14.—(*a*) *L-band aerials, vertical polarization. Dielectric constant of earth* = 10; *Earth's radius* = 4,585 *n.m.; Earth's conductivity* = 10^{-14} *e.m.u. Dashed curve is for dipole at* 25 *ft. Full curve is for dipole at* 27 *ft.*

(*b*) *Difference between aerials at* 25 *ft. and* 27 *ft. against elevation angle.*

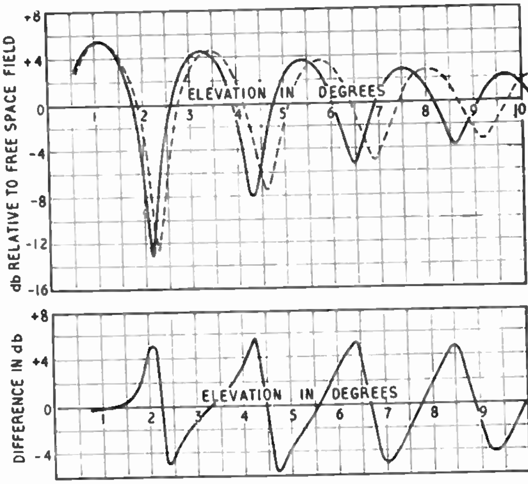


Fig. 15.—(*a*) *Dashed curve for dipole at* 10 *ft. Full curve for dipole at* 10 *ft.* 8 *ins. Conditions otherwise as in Fig.* 14.

(*b*) *Difference between aerials at* 10 *ft. and* 10 *ft.* 8 *ins. against elevation angle.*

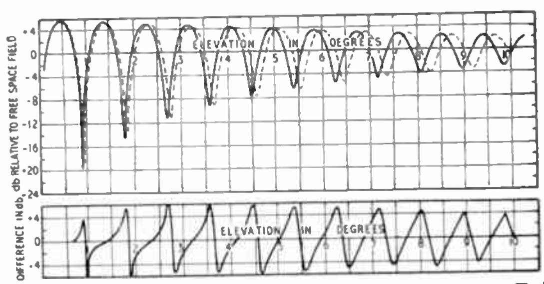


Fig. 16.—(*a*) *Dashed curve for dipole at* 25 *ft. Full curve for dipole at* 25 *ft.* 8 *ins. Conditions otherwise as in Fig.* 14.

(*b*) *Difference between aerials at* 25 *ft. and* 25 *ft.* 8 *ins. against elevation angle.*

space. Calculations for propagation over a plane earth show that when two aerials of identical broad free-space patterns are located at a height difference of 8 inches, the relative variation of signal strength with elevation is about ±6 db for aerial heights between 10 feet and 100 feet (Figs. 14, 15 and 16). If, however, the spacing is less than 10 yards in the horizontal plane and of the order of 1 inch in the vertical plane, the difference in the vertical patterns is negligible, provided that the screening effect of one aerial upon the other may be ignored.

5.7.3. Combined interrogator and control aerials

The interrogator aerial of the secondary radar equipment now being evaluated by the Ministry of Transport and Civil Aviation (Fig. 17) comprises a broadside array of 24 dipoles in front of a vertical reflecting screen. The control pattern is produced by an extra dipole placed at the centre of the broadside array. The vertical reflecting screen is omitted from behind this dipole only, so that some back radiation exists from the control aerial. The control aerial vertical pattern is broader than that of the interrogator aerial, but the difference is negligible for elevation angles up to 30 deg. The interrogator array screens the radiation from the control aerial in directions between ±80 deg. and 100 deg. relative to the main lobe, so that the horizontal control aerial pattern is an asymmetrical figure-of-eight (Fig. 18). In practice, the free-space control aerial pattern falls by 30 db in the directions of minimum radiation; but, since the free-space pattern of the interrogator array falls by more than 40 db in these directions, it was at the outset of design considered unlikely that the interrogator signals

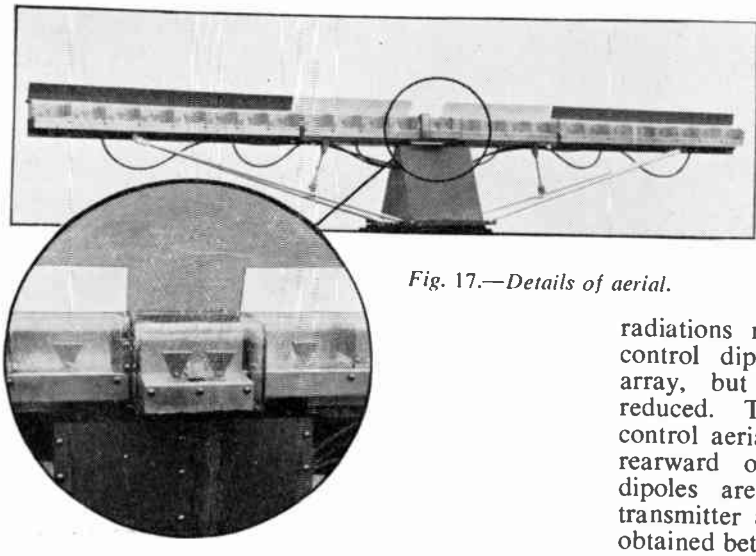


Fig. 17.—Details of aerial.

would exceed the control signals in these directions. Field trial experience of the equipment indicates that this assumption is true when the ground aerial is sited reasonably clear of obstructions. However, under some siting conditions occasional spurious responses have been observed with aircraft flying closer to the ground station than 10 miles. These responses have almost certainly been due to reflection of the main lobe by buildings, causing high amplitude artificial sidelobes to appear in the horizontal polar pattern of the aerial. In one such case the ground aerial was sited at a height of 10 feet about 40 feet away from an assembly of brick built buildings of heights about 15 feet. The sidelobe responses only appeared when the aircraft were located in certain directions and the paints were only about 2 deg. wide compared with the main lobe paints of width about 5 deg. (see Fig. 19). These sidelobe paints only appeared erratically and could possibly be tolerated for some operational purposes. It is possible that this effect may not appear at all at a more carefully chosen site, but if it proves to be an important source of trouble, it will be desirable to make the control aerial pattern more omnidirectional. This can probably only be done by taking advantage of the fact that a horizontal spacing of 10 yards between the two aerials is permissible; but a spacing as large as 10 yards cannot easily be used in practice due to the mechanical difficulty of keeping the vertical spacing within a few inches.

The side radiation from a control dipole placed at the centre of an interrogator array may be increased by moving the dipole forward of the array; the back radiation is then reduced due to the screening effect of the interrogator array on the control aerial pattern. Conversely, the side and back radiations may be increased by moving the control dipole rearward of the interrogator array, but the forward radiation is then reduced. Two dipoles may be used for the control aerial with one forward and the other rearward of the interrogator. When such dipoles are fed in phase from the same transmitter an r.f. interference pattern will be obtained between the signals radiated from each dipole (Fig. 20). Whether this pattern will give a satisfactory operational control pattern depends on the spacing, the relative arrangement of, and the power fed to the forward and rear dipole elements. These factors control the number of interference lobes and the depths of the nulls. There is some hope that without further complexity a good enough control pattern can be got in this way. Alternatively it

Fig. 18.—Combined horizontal interrogator and control patterns, illustrating the screening of the control pattern by the interrogator array when the control radiators are close together as shown in Fig. 17.

may be possible to feed forward and rear control radiator elements with suitably random non-coherent r.f. at the same time. Failing adequate operation from any of these methods it is possible to feed the forward and rearward control radiators from two transmitters, the control transmission on the rearward dipole being made 1 μsec after that on the forward dipole. Destructive interference cannot then occur and the trigger level of the airborne set is determined by that control pulse which is of greatest amplitude in any direction. Fig. 21 shows forward and rearward control patterns superimposed on the interrogator pattern for a particular case under test. In this case the boxes containing the control radiators can be put together or moved up to 4 feet apart. The fact that spacing between a control and

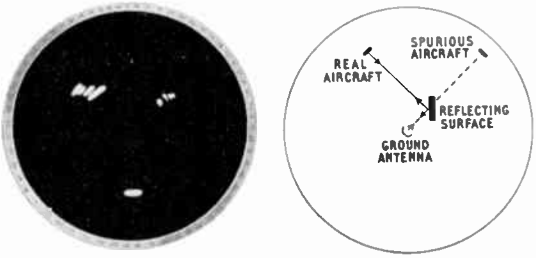


Fig. 19.—Reply signals of aircraft in northwest sector reflected off airport hangar building to produce spurious targets visible in northeast sector.

interrogator pulse is only 7 μsec in the rearward direction, as against 8 μsec in the forward direction, for the two pulse method of transmission, is unimportant since interrogation in this direction is to be suppressed. Destructive interference between the rearward control pulse and an echo of the forward control pulse that arrives at the transponder at the same time as the rearward pulse cannot seriously effect the

Fig 20.—Pattern for forward and rear control radiators spaced apart and radiating coherently.

operation of the system since the trigger level will by then have been set by the amplitude of the forward control pulse. A system incorporating this final elaborate arrangement is under test; as indicated above it may prove possible to simplify it when sufficient field experience is available. With forward and rear control radiators the question arises how much they will modify the interrogator pattern. Figs 12 (b) and (c) answer this question in one practical case.

5.8. *Capture*

Capture results when the control pulse method of suppressing interrogations by sidelobes is employed. The control pulse renders the transponder insensitive to pulses of amplitude similar to that of sidelobes; thus where the main beam of a distant station produces signal strengths of the same order as local sidelobes, service may be denied to the distant station.

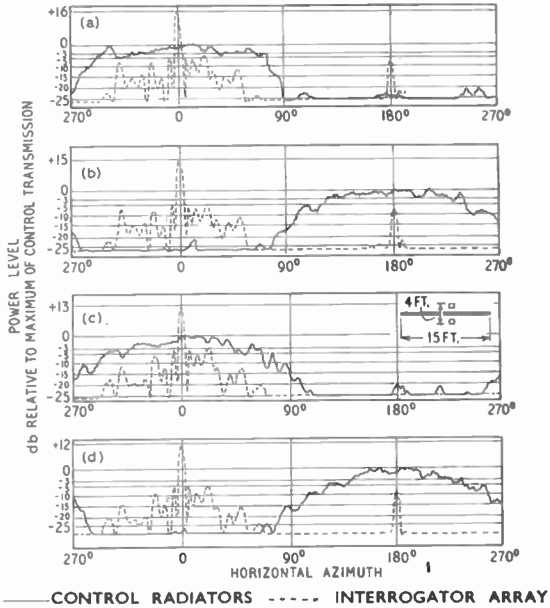


Fig. 21.—Radiation patterns of interrogation array and control aerials.

(a) Control aerials close together and only forward control radiator being used.
(b) As (a) but with only rear-looking control radiator being used.
(c) As (a), but with control radiators spaced 4 ft. apart, only forward radiator in use.
(d) As for (c) but forward control aerial unused, and rear-looking control radiator in operation.

The transponder is captured in this way whether the very long time constant, long time constant, or short time constant techniques of suppression are employed; but whereas with the first two methods it is effectively captured continuously, with the short time constant technique the transponder is captured for a time of the order of 100 microseconds only every 3 or 4 milliseconds. Thus capture-free service is given to many ground stations interrogating at random. This is the paramount

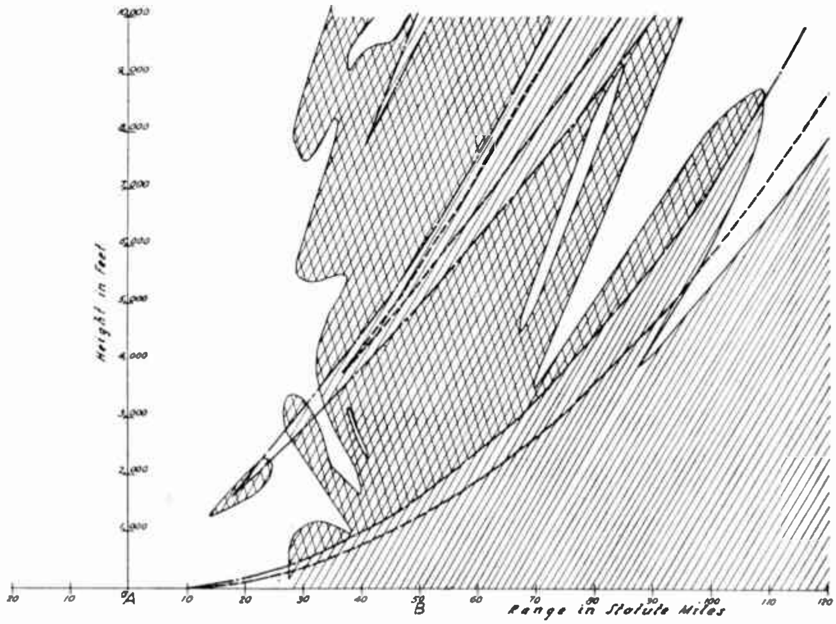
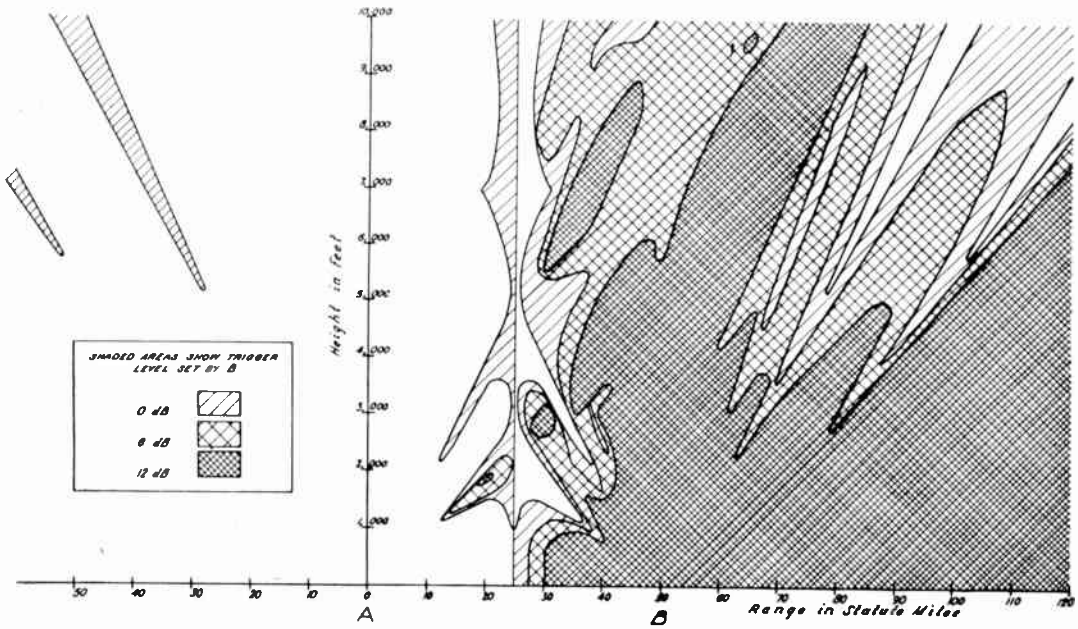


Fig. 22.—Capture effect. Cross section in line of stations A and B, 50 miles apart. Trigger level set by B, 6 db below main beam.
20 db contour shown thus: —·—·—·—
26 db contour shown thus: – – – – – – – –

Fig. 23 (below).—As Fig. 22 but illustrating the result of trigger levels at various settings.



advantage of the short time constant method of sidelobe suppression.

It might be supposed that with sidelobes 20 db to 30 db down on the main beam and the transponder sensitivity set just to preclude sidelobe interrogation the volume of airspace captured will not be very large and the basically simple "very long time constant" method might prove operationally adequate. In practice, however, it is not possible to set the sensitivity so far down on the main beam or sidelobes will break through. Allowance must be made for:

(*a*) The fact that after the level has been set, the sensitivity must be allowed to return to a maximum in a finite time.

(*b*) Changes of aircraft aspect between successive settings of the trigger level.

(*c*) System amplitude variations.

(*d*) The ability of field maintenance technicians to set conditions at the right starting point.

(*e*) Variations in amplitude between the interrogating and control transmitters on the ground.

More detailed analysis shows that a setting between 6 db and 12 db down on the main beam is optimum for sidelobes between 20 db and 30 db down on the main beam. This analysis has been extended to calculate the capture areas for two stations 50 miles apart and making the following assumptions:

(*a*) The elevation radiation pattern is that of the dotted curve of Fig. 13.

(*b*) The beam is tilted 4 degrees in elevation.

(*c*) Vertical polarization is used.

(*d*) The interrogation frequency is 1,215 Mc/s.

(*e*) The mean scanner height is 25 feet above the ground.

(*f*) The effective earth's radius is 5,280 miles.

(*g*) The dielectric constant of the earth is 10.

Figures 22 and 23 show the kind of capture areas involved in such cases.

5.9. *Second-Time-Round Signals*

It was suggested in the discussion of sidelobe suppression that to ensure any guaranteed system range performance it is necessary to design for 45 db greater power under the most adverse system conditions than under the optimum conditions. That is, with an aircraft in a maximum of the ground station pattern, presenting an optimum "view" of its own aerial, having a transponder of sensitivity at the top limit and maximum transmitter power, 45 db less system gain will be required than for an aircraft in a minimum of the ground station pattern, and with the other factors all adverse. Thus under the most favourable circumstances, even in a medium range system, aircraft will sometimes be seen at very great ranges. If it may be assumed that transponder-equipped aircraft will not operate at heights above 50,000 ft., the radio line of sight is limited to about 280 nautical miles. Suppose that the interrogating signals are emitted at a repetition frequency of 500 per sec, then the interval between pulses is 2 millisec, which is equivalent to 167 nautical miles. The reply from an aircraft at a range of 280 nautical miles is then received at the ground station after a second ground transmission has taken place and the reply cannot be distinguished from that of an aircraft at a range of 113 nautical miles. Such replies are called "second time round" signals as they are when encountered in primary radar systems. Fortunately with primary radar they are rarely troublesome except under anomalous propagation conditions. Sensitivity/time control on the receiver will help to reduce second-time-round signals but as has been seen the vertical coverage diagram of a ground aerial of small vertical aperture may vary by as much as 25 db with elevation at any given range, thus sensitivity/time control cannot be totally effective even at short ranges. Further it cannot be applied without mutilating the vertical coverage diagram and must therefore not be applied too stringently or to any considerable range. A typical case of such mutilation is illustrated in Fig. 24. The evident recourse is to a reduction of the pulse recurrence frequency, and if the operational height limit is 50,000 feet the recurrence frequency must not exceed 300 pulses per second. As aircraft fly higher there will be a need further to reduce this p.r.f. limit. When an independent interrogator is to be associated with a primary radar of a high pulse repetition frequency it is necessary to trigger the interrogator at a suitable sub-multiple which is less than 300 p.p.s. if the same display unit is to be employed for both. It will be recalled that this small recurrence frequency spoils the horizontal resolution of the secondary radar, unless it can be dissociated from the primary radar and allowed to rotate much more slowly and use a greater horizontal aerial. It will also be seen in the ensuing argument that the low p.r.f. limits the possible reply coding techniques.

5.10. *Unlocked Responses*

The secondary radar ground station receiver will accept signals from every transponder which transmits on the universal frequency whether the response is made to an interrogation by the local transmitter or by any one of the other ground stations within range. In high traffic density areas where there are many ground

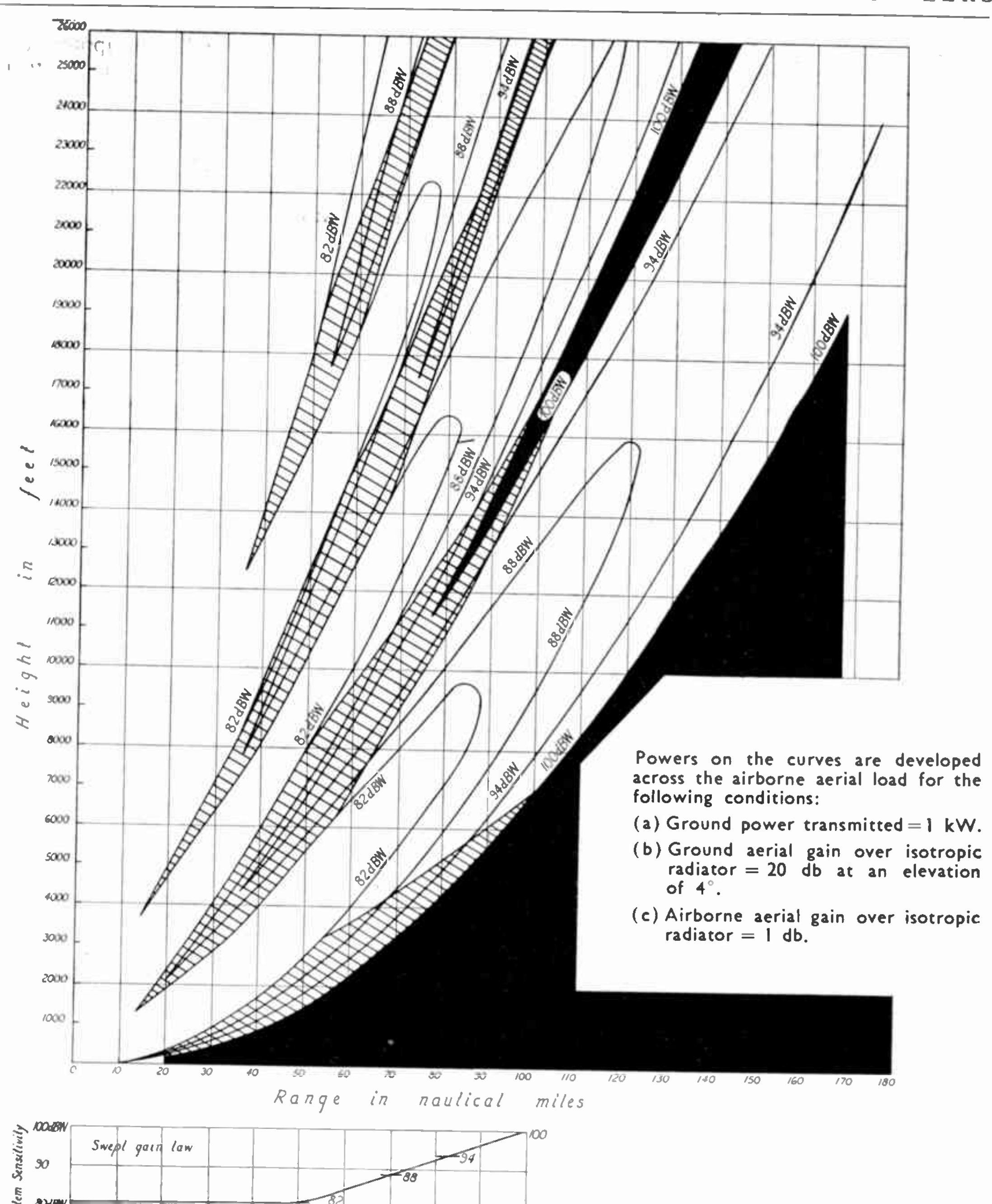


Fig. 24.—Vertical coverage. A system of sensitivity 100 *dbW gives solid cover to the left and above the solid black area with the exception of the single black wedge.*

The use of the swept gain law illustrated above introduces further gaps in the coverage as shown by the hatched areas. It is not implied that this particular swept gain law has any operational significance; the principle remains the same, however, for any swept gain law.

stations there will, therefore, be many unlocked signals appearing on the ground radar display. Such signals have been called "fruit," and devices to minimize their interfering effects are, in the U.S.A., called "defruiters." Evidently the problem of "fruit" requires to be minimized by all possible means. First, unnecessary interrogation should be avoided, (*a*) by coding the interrogation so that only the interested ground stations trigger the transponder and only the aircraft appropriate reply, and (*b*) by suppressing sidelobe interrogation on the ground-to-air channel. Secondly, steps may be taken at the ground station to prevent unlocked signals appearing on the display. This may be done: (*a*) by using a decoder on the ground and allowing only certain reply codes to go forward to the display; and (*b*) by the use of a suitable coincidence circuit "defruiter."

For example, an ultrasonic delay line of the M.T.I. radar type may be used so that signals immediately received may be compared with those obtained in response to the preceding local interrogating pulse group. The line length is made to equal the local station pulse period, so that signals are only passed forward to the display if there are two successive responses at the same range.

5.11. *Transponder Saturation*

Transponder saturation is said to occur when the transponder is no longer capable of replying to each correctly coded interrogating signal which is received. Transponder inability to reply may arise due to the following causes:

(*a*) The interrogation signals from one station may arrive during the period when the transponder is captured by another station.

(*b*) The mean rate of interrogations may exceed the maximum tolerable duty cycle of the airborne transmitter and "count-down" may become necessary.

The first of these is not true saturation and its solution is to make the capture period short compared with the mean spacing of interrogating signals and to rely for an adequate service upon the random unlocked pulse repetition frequency relationship of the interrogators.

The solution to the second problem is evidently to make the maximum duty cycle of the airborne transmitter adequate to meet the most stringent operational conditions. However, to limit the complexity of the airborne equipment, one should not demand too large a duty cycle. The ways in which the required duty cycle may be limited are as follows:

(*a*) Make the pulse repetition frequencies of interrogators as low as possible.

(*b*) Specify that interrogating aerials shall not be "laid on" aircraft, i.e. they should always be rotating.

(*c*) Suppress sidelobe interrogation on the ground-to-air channel.

(*d*) Code the ground-to-air channel to reduce triggering by interference pulses.

(*e*) Limit the coded reply to as few pulses as possible.

In the limit, however, it will be necessary to count down so that the service to ground stations is reduced when too many interrogations are received. The way in which this is done may be of operational importance and the best method is dependent on some of the other system characteristics.

5.11.1. The magnitude of the problem

In one available transponder which is quite typical in this respect, the transmitter is able to produce pulses at a peak rate of 15,000 per sec for a period of the order of 40 milliseconds and count-down circuits are fitted so that the mean rate of reply is limited to not less than 4,000 p.p.s. averaged over a period of longer than 1 second.

Thus suppose a locality to be dotted with ground interrogators each having a beamwidth of 5 deg. and rotating at 20 rev/min (approximately 40 milliseconds transit time), and having a p.r.f. of say 400 p.p.s. Then each station puts 16 interrogating pulse groups into the transponder each 3 seconds. One form of reply code now discussed in the U.S.A. for civil traffic may require 9 pulses to be sent in the reply group for each interrogation; thus if this maximum reply code is in use each ground station demands 144 reply pulses every 3 seconds and assuming that the beams sweep at random over the aircraft it can be seen that more than 50 ground stations will get adequate service; incidentally they would still all get service if their interrogating beams all coincided in sweeping over the aircraft because in that case less than 15,000 pulses would be demanded in the same 40 milliseconds.

The foregoing argument stands if all sidelobe interrogation has been eliminated on the ground to air channel and provided that no interrogator is "laid on" to the transponder. If, however,

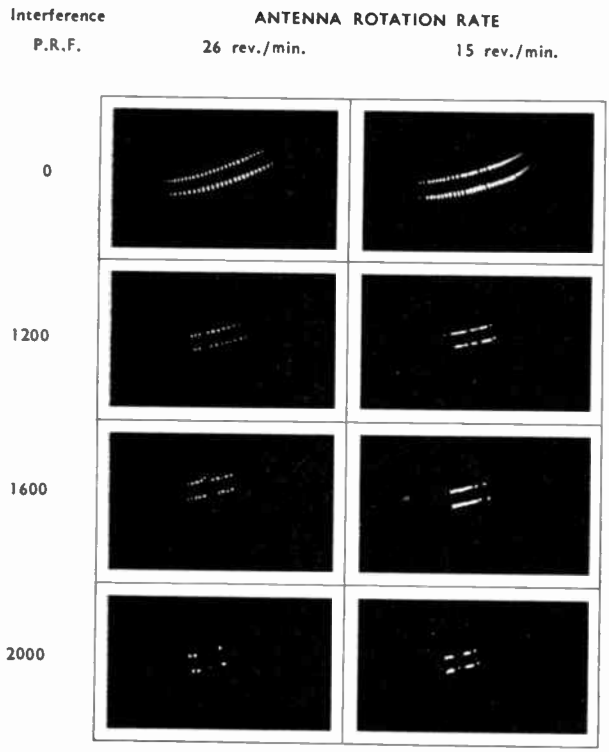


Fig. 25.—Expanded view of decoded target showing effects of increased interference on target quality. (Participating I-R unit p.r.f. 300.)

the transponder from which all 9 pulses are being demanded is within the range at which there is sidelobe interrogation from one station throughout 360 deg. and there is no ground to air channel sidelobe suppression then, at a p.r.f. of 400 p.p.s., this one ground station alone demands 3,600 of the 4,000 pulses each second which will saturate the transponder: two local ground stations will saturate all the aircraft within the range at which there is sidelobe interrogation through 360 deg. from which five or more pulses are demanded in the reply (Fig. 25).

This is one of the most powerful arguments in favour of sidelobe suppression on the ground to air channel, but as this has not been provided in some existing systems the method of count-down remains of considerable importance. It must also remain of importance so long as there is a possibility that laid-on aerials may be used for some purposes, e.g. for more complex intelligence transmission systems.

5.11.2. Methods of count-down

Count-down has been described in recent discussion as "random" or "sensitivity." With "random count-down" the transponder sensitivity is reduced for a period following each successful interrogation when the mean interrogation rate rises towards the limit. This means that all the ground stations within range tend to get a reduced percentage service; at first occasional and then more frequent pulses are lost. The effect on the radar indicator is the same as reducing the number of pulses per beamwidth in primary radar, the radar blips becoming thinner and ultimately range performance is reduced. With "sensitivity count-down" it is arranged that when the mean interrogation rate rises, the transponder sensitivity is reduced through a relatively long time constant. Thus the weaker interrogations fail to get service but stronger ones get full service. It will be observed that local sidelobe interrogation 20 db to 30 db down on the local main beam may be weaker than a remote station main beam of equal power to the local station; the remote station may well get continuous service to the exclusion of the local sidelobe interrogations when it is ten times more distant from the aircraft than the local station. To this extent sensitivity count-down acts in the right direction, but stations more distant still may lose service entirely when they would still be getting adequate results with random count-down.

The great advantage of sensitivity count-down is, however, that when service is provided it should be nearly 100 per cent. This will be of importance where a "defruiter" is employed; the loss of every other response pulse train to a system which only recognizes an aircraft to be present when it gets two successive pulse trains at the same range and azimuth means a total failure to paint. Evidently with random count-down, pulses are not likely to be lost in this uniform way, but the loss of occasional pulses is more serious when a "defruiter" is used than when it is not. The full comparison of the advantages of random and sensitivity count-down is a statistical and operational matter, but because of the foregoing argument an American specification has recently changed its requirements from "random" to "sensitivity" count-down.

5.11.3. Count-down techniques

With either count-down system it is first necessary to measure the rate at which the transponder is triggering. This is best done in the modulator where a sufficient power level is available without introducing amplification stages. The pulsing requirement of the triode oscillators employed demands a pulse transformer and suitable polarity signals can be derived from it. After integration this measure of trigger rate can be applied to a biased diode or some other suitable threshold measuring device. For example, a modified form of random count-down can be achieved by feeding the derived signal to the suppressor of a short suppressor base pentode in the video amplifying chain after the i.f. amplifier. In this case the suppressor base of the valve acts as the threshold determining arrangement. Thus after the reply rate has reached a predetermined level no replies are transmitted for a period, say 50 milliseconds, which is long compared with the minimum possible space of reply groups of the system. If this method is employed occasional complete paints will be lost as the transponder begins to count down, but the method satisfies the "defruiters." Alternatively, random coun-down can be applied so that occasional pulses are lost, by lengthening the recovery time of the transponder from its normal 50 microseconds or so progressively as the trigger rate increases by using the amplified integrated trigger rate information to lengthen the recovery time of the gate associated with the coding circuit. An advantage of this method is that it can readily be made continuously variable. Finally, the sensitivity count-down method is simply achievable by using the trigger derived signal as a variable bias to the i.f. amplifier grids.

When countdown is required the designer is therefore faced with the choice of reducing sensitivity, thus possibly counting out the distant ground stations, or putting on a long period dead time which results in complete paints being lost from the indicator, or using a short period dead time which thins out the number of pulses in a paint and may be specially troublesome if a "defruiter" is employed. All three systems may be met in the field; as stated, the American civil authorities are showing a preference for the first.

5.12 *Airborne Aerials*

Ideally a line of sight should exist between the airborne aerial and the ground station throughout the whole of the required coverage of the system. In practice the aspect of an aircraft varies throughout flight and it is not possible to find a single point on the aircraft surface which is always in optical view from the ground station (see Fig. 26). Airborne aerials are generally best located at the lowest or highest points in flight and the diffraction from such aerials is relied upon to maintain radio contact when an optical line of sight does not exist. In general terms diffracted signals are stronger at lower signal frequencies so that it may be expected that the coverage of an L-band airborne aerial will be more uniform than that of a similar S-band aerial. Vertical polarization is to be preferred although horizontal and even circular polarized airborne aerials have been made.

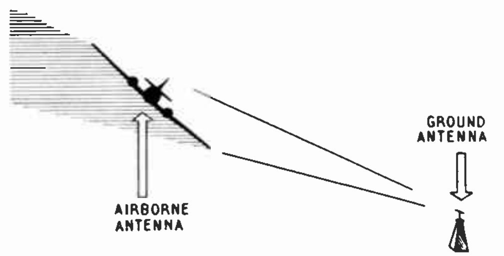
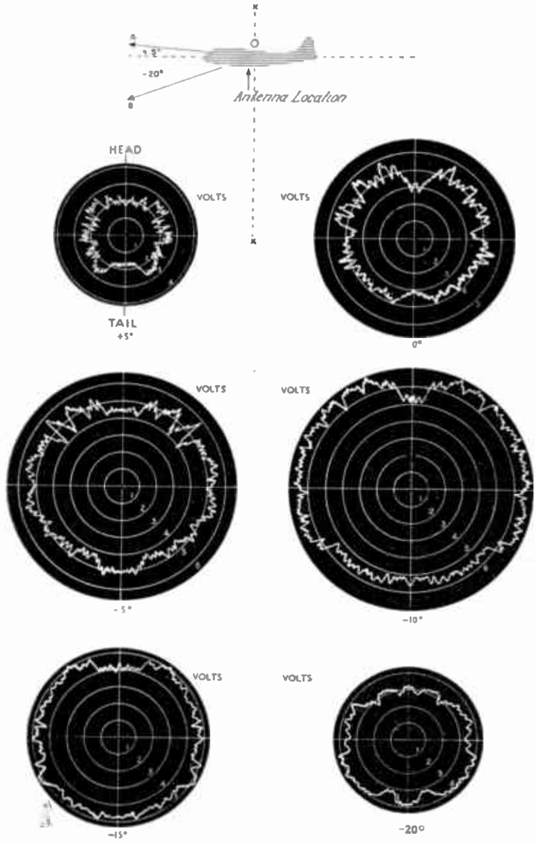


Fig. 26.—Shadowing effects.

A single vertical polarized L-band airborne aerial mounted at the lowest point in flight on a typical aircraft will have a polar diagram which varies some 6 db in the horizontal plane. Under conditions of bank of 15 deg. or pitch of 5 deg. the masking effect of the aircraft will cause a further reduction of some 10 db in the masked direction. A typical set of conditions is shown in the six curves of Fig. 27a. At +5 deg. above the aircraft the radiation is fairly well screened, for higher angles of elevation the screening is even more complete. Below the aircraft is the familiar semi-doughnut (Fig 27b) and in the cross sections of Fig. 27c are seen the developments of successive angular sections of this doughnut which are modified as parts of the aircraft screen the antenna or as reflections from the aircraft produce minor interference patterns.

Thus if an allowance of 14 db is made in the system design and one belly-mounted



(a)

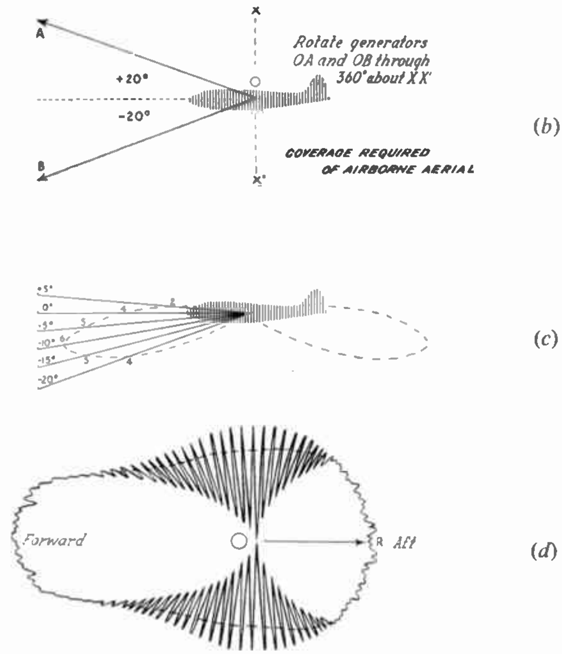


Fig. 27.—(a) Actual coverage of belly mounted aerial.
(b) Coverage below aircraft.
(c) Cross-sections of coverage pattern.
(d) Coverage using two discone aerials with reflectors.

antenna is used, then at maximum range the aircraft will be "seen" if it presents any horizontal aspect between say − 20 and + 2 deg. (20 deg. down and 2 deg. up from the horizontal). As the range of the aircraft decreases it will be seen over a wider and wider range of vertical angles; at one-quarter maximum range it should be seen over the whole lower hemisphere and maybe up to 10 degrees bank away from the ground station in most aspects. A single aerial used in this way may be satisfactory for general purposes.

It has, however, been required in some systems that a coverage of $\pm$ 20 degrees to 6 db down on the maximum shall be given and this cannot be obtained with any single airborne aerial location. A nose antenna is screened in all backward directions, for example, and an antenna on the tip of the stabilizing fin is considerably but erratically screened in all downward directions and taken alone is inferior to belly mounting. Dual installations, or duplication to the video stages solve the problem but are obviously expensive. High speed switching between two aerials has been tried, but is also complex; finally, two widely spaced aerials mixing at r.f. have been employed experimentally. For example, one installation used a discone aerial with a 150 deg. corner reflector in the nose and a discone with a 300 deg. reflector on the stabilizing fin. In the particular aircraft considered this gave about 100 λ spacing at 1,000 Mc/s and a diagram of the general form of Fig. 27d in the reference plane of the aircraft. Allowing that the coverage in the interference pattern area may be represented by the dotted lines, such an arrangement gave a performance within 6 to 8 db of the vector OR for all angles $\pm$25 deg. of the reference plane. The assumption involved in accepting the dotted lines as the coverage is probably justifiable if the aircraft "flies through substantially more than one lobe-cycle" in a rotation of the ground aerial period.

5.13. *Coding*

Both interrogation and response can be coded. The purpose of coding on the ground-to-air channel is to contrive that the airborne units shall not be needlessly interrogated, thus reducing the over-interrogation and unlocked response problems, and to arrange that replies are obtained only from aircraft of immediate interest to the air traffic control system. The use of a simple double pulse with 8 µsec spacing between the pulses (see Fig. 28) is the presently proposed civil form and it prevents interrogation by primary radars for example.

Considerable intelligence can be transmitted from air to ground if required. In accordance with normal communication theory, one or both of two things will have to be sacrificed progressively as the amount of information to be conveyed is increased. Either the rate at which the information is renewed must be decreased or the effective resolution of the radar system must deteriorate. For example, a recognition aerial system may be used in addition to the interrogating and control aerials. This recognition aerial can be "laid on" to the aircraft from which intelligence is required and left illuminating it for long enough to read the message. Alternatively, for each received pulse the aircraft responder can transmit a number of pulses in succession to convey the required intelligence. In the first case time is required; in the second, radar resolution is sacrificed. It is important in designing the coding system to get the best compromise.

5.13.1. Purposes of coding on the air-to-ground channel

The possible purposes of response coding are:

(*a*) To establish which radar "echo" corresponds to an aircraft known to be in the ground station coverage.

(*b*) To identify an aircraft which is calling on its radio telephone.

(*c*) To permit the ground controller to identify any "echo" on his radar display without radio telephone contact with the aircraft.

(*d*) To distinguish one group of aircraft from another, e.g. the aircraft coming to land at London from those over-flying the area, or in the extreme to distinguish friend from foe.

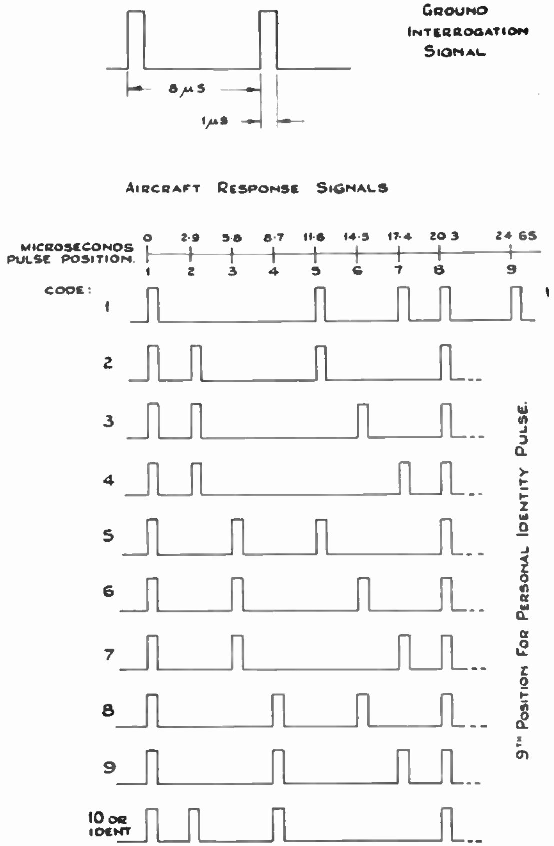


Fig. 28.—Pulse codes used in ATC radar beacon system.

(*e*) To add security to the distinction between friend and foe by making it difficult to copy what the friend is doing.

(*f*) To include information about aircraft height taken from the altimeter.

(*g*) To add other information, for example, to read the aircraft position from the airborne navigational aid.

The first three of these may be classed as "Recognition," the next two as "I.F.F.," and the last two as "Additional Intelligence." In any discussion of coding it is necessary to establish to which of these seven purposes the system is directed. The required complexity increases as one progresses from (*a*) to (*c*) or from (*d*) to (*e*) and can become extreme with the inclusion of (*f*) and (*g*).

A single code switched on by the pilot when asked "Where are you" by the ground controller meets requirement (*a*). A single code switched on with the telephony transmitter meets requirement (*b*), though it is evidently preferable to have one code per r.t. channel if the aircraft may communicate on different channels with different ground stations; and if (*a*) and (*b*) are to give simultaneous service the code for (*a*) must differ from that for (*b*). The minimum number of codes for (*c*) is the maximum number of aircraft which may be found in the coverage of a given ground control authority in a chosen "organizational period"—plus a margin of safety. The organizational period is the time which must elapse before a given code can be used again without causing confusion. Estimates of the requirement vary from 10 to 200 codes. The profligate way to satisfy (*c*) is to allot each aircraft an immutable code and the order of the requirement then becomes ten million codes. (*d*) demands a single code shared by all friendly aircraft as a minimum; this may be increased to two or three codes to permit "Distress" or other special signals to be given. (*e*) may be of any amount of complexity. (*f*) might for example use something of the order of 100 codes to transmit height in steps of 250 feet to 10,000 feet, 500 feet from 10,000 to 30,000 feet and then 1,000-foot steps to 50,000 feet. (*g*) again might demand any amount of complexity depending on the area to be covered and the positional accuracy required.

The present requirement then is indisputably for one code, strongly weighted in favour of perhaps half a dozen, and showing signs of growing, although the upper limit required is difficult to forecast.

5.13.2. Methods of coding

To code a pulse transmission one can in theory vary one or more of four things:

(*a*) The number of pulses sent in reply to an interrogating pulse group,

(*b*) The pulse amplitude,

(*c*) The pulse width, and

(*d*) The time interval between two or more pulses.

The pulse amplitude varies with range in the nature of things; it is not simple to vary it accurately, and it is undesirable deliberately to demand more power for coding than is required to achieve the desired range. Amplitude variation can therefore be discarded.

The choice of a suitable system of coding must consider pulse elongation which occurs because signals arrive from the aircraft to the ground by indirect paths as well as the direct one. Practical experience shows that in a typical system a reply pulse may be elongated or have indirect path "echoes" associated with it of significant amplitude to five or ten times its own duration. This means that, if no pulse amplitude measurement is to be made, no two pulses in the reply may be put closer than some ten times the response pulse length unless operational "guard" techniques are employed in the system, e.g. pulses are only sought at certain specific points in time and a total count is made to eliminate erroneous information.

A straight measurement of number of pulses can also be discarded as a possibility if more than two or three codes are wanted because the "back edge elongation" may include discrete reflections separated from the main body of the response. More important still, if many codes are required the duration of a long train would with such a system reduce the effective range discrimination, possibly to an intolerable extent. For example, ten pulses spaced even as close as 3 μsec from one another occupy almost 3 miles of airspace and although steps can evidently be taken to remove overlap on the radar indicator, false information is likely to follow when the two aircraft get within three miles range of one another at the same bearing.

A straight measurement of pulse width could also only be made in increments of, say, 3 μsec or whatever the minimum practical safe spacing may be. This makes heavy demands on the airborne transmitter power and requires the same length of time base as the counting of pulses for a given number of codes.

Interval between two pulses can be measured in suitable increments and the system is capable of expansion by adding further pulses again at variable and measurable intervals. Such a system demands too much stringency in both the airborne and ground timing circuits and must therefore be abandoned.

Another system is to combine number and interval. The system looks at each T μsec after the first pulse to see if a pulse is present there or not. Here T is the minimum practical interval considering pulse elongation. For T μsec this gives 1 code, for $2T$ μsec 3 codes, for $3T$ μsec 7 codes, and so on, $(2^n - 1)$ codes where n is the possible number of pulses after the first.

Yet another system is to modulate the back edges of the transmitted pulse with a mixture of audio frequencies. To modulate simultaneously with n frequencies gives $(2^n - 1)$ codes. The modulation varies the pulse width from say 1 to 3 μsec in the absence of back edge elongation. The greater advantages of this system are, (i) that the radar resolution is only reduced by this amount, and (ii) that it can be arranged that the absolute lengths are unimportant as it is the rate of variation which is measured, thus the back edge distortion is immaterial.

The last two methods are the only reasonable ones, when "all the coding information is required all the time," that is when the complete answer is required in a very short time interval after each burst of interrogating pulses. The first has the advantage of immediate display on the radar plan position indicator if the number of codes is very small; difficulties of distinguishing codes arise even with three or four pulses, for example, pulse, space, pulse, space may look very similar to pulse, space, space pulse, so that beyond two or three codes a ground decoder is required. The loss of radar resolution may also become serious at more than three or four pulses. The audio frequency modulation of pulse length essentially requires a separate decoder; this requirement is to be weighed against the better resolution. In a practical system the number of codes obtainable with this technique is not large. The available bandwidth for the modulating frequencies is confined to one half of the radar system pulse recurrence frequency. It has been shown that to reduce second time round problems a p.r.f. of the order of 300 per second is desirable and thus the band for modulating frequencies is reduced to 150 c/s. This means that even with modern filter techniques not more than three modulating frequencies are available and back edge modulation is only of academic interest until a "sequential" method of coding is employed. Thus the multiple pulse technique is the only practical available possibility unless the alternative of using a "laid on" coding aerial is employed. With multiple pulse coding, resolution in range is sacrificed and a practical maximum of the order of 100 codes is the limit.

By a sequential coding system is meant one in which the form of code is varied in a measurable sequence. If full advantage is to be taken of a sequential system a separate interrogating aerial must be used with the coding equipment. A large number of codes is then practicable given a detection time of a few seconds. A modified form of the pulse width modulation with audio frequencies is then thought to be the best system. It allows separate recognition with less interference on the plan position picture. The preferred form is to use three modulating frequencies (a mark, a space, and a control frequency), one, and only one, of which is present at any given time. The probability of spurious audio frequencies within the band of modulating frequencies being generated by the changes in back edge distortion with time has been investigated theoretically and practically and is very small. The fact that pulses and not continuous waves are used as carrier for the audio frequencies does not introduce false signals in the wanted frequency range of any considerable magnitude. The circuits can also be made sufficiently linear without great difficulty, to ensure no trouble from intermodulation products.

5.13.3. The present United States proposal

The transponder being developed for the civil airlines in the U.S.A. (where it is known as the Airborne Safety Beacon) will at present give ten codes which are obtained as follows: Four pulses are sent in response to an interrogating group; of these, two (called the framing pulses) are always present and are spaced with leading edges 20·3 μsec apart and the others occupy two of the six available points spaced at 2·9 μsec intervals between the framing pulses. It has been pointed out that by filling 0, or any 1, 2, 3, 4, 5 or all 6 of these spaces 64 codes are obtainable. A requirement also exists for associating a "Personal Identity" with each of the codes by the addition of a ninth pulse 4·35 μsec after the final framing pulse. Thus effectively 128 codes will be provided using a minimum of 2 and a maximum of 9 pulses.

The received information will be presented to a ground decoder and the arrangement permits a single blip, or a double blip, or a widened blip to be passed on to the ground station display. The decoder and the display unit arrangements are such that the controller can choose whether he will display all aircraft, aircraft showing a specific code only, or a specific code plus all "Personal Identities" distinguished on the display by a broad or a double blip.

This decoder arrangement is subject to one specially objectionable defect under some circumstances. When two transponder-equipped aircraft within one beamwidth in azimuth approach within 3·3 nautical miles of each other ($2 \times 20{\cdot}3$ μsec), false signals may be produced as they pass each other. Care has been taken to minimize this defect in the decoder, but it may lead to false information on the indicator in some circumstances. The way in which this can occur is shown in Fig. 29; it will be seen that the problem may be really serious if all the available 64 codes are used.

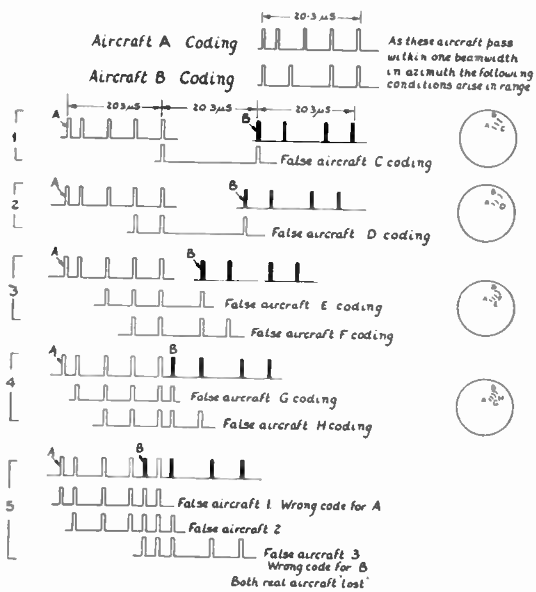


Fig. 29.—Derivation of false signals. The analysis can be extended to show, that, as these two particular aircraft pass, a total of 41 false aircraft are generated. Of these, 17 occur when one of the real aircraft apparently has a code other than its true one.

Evidently steps can be taken in the ground decoder to select for presentation on the display certain code combinations only, so that many of these false aircraft may never be seen. Potentially however they represent a serious operational hazard and much care will have to be exercised with respect to them. There is less of a problem if the 10 originally suggested codes only are used because then there must always be four pulses and four pulses only or the answer is false; in this case fewer false aircraft are generated but the real aircraft may still be eliminated a number of times as they pass each other.

6. Conclusion

The treatment of this subject has been made intentionally broad in scope. Much difficulty has arisen in the past, and continues now, because the airborne transponder has been considered as a separate entity and the problems of integrating it into a system have been postponed until the airborne unit has become available. It is hoped that this paper has shown that such an approach is not likely to provide the best secondary surveillance system.

It remains to stress again that secondary radar complements but cannot, except in limited applications, replace primary radar as an aid to Air Traffic Control, and to repeat that radar *in toto* helps to solve but one part of a complex problem when it is applied to the control of air traffic. The major contribution is still demanded in organization and executive skill by personnel in an intricate Air Traffic Control System.

7. Acknowledgments

The author has pleasure in acknowledging the permission given by Messrs. A. C. Cossor Ltd., to publish this paper. He is also indebted to colleagues for the work which has led to the accumulation of information summarized herein: in particular credit is due to Mr. D. A. Levell whose original contributions in this field have been considerable.

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TRAINING IN THE B.B.C.

This year the Engineering Training Department of the B.B.C.* celebrates the tenth anniversary of its establishment at Wood Norton Hall, Evesham. The mansion, which was used by the B.B.C. as an emergency broadcasting centre during the war, was taken over by the Training Department in 1946. All the courses operated there are residential, and following the completion of a new dormitory block accommodation is now available for just over two hundred students. The number of students trained annually is expected to rise from six hundred to eight hundred this year.

Under its head, Dr. K. R. Sturley, Ph.D., B.Sc., the Department is concerned largely with the specialist training of engineers and technicians, and the development of their skills and capabilities in the application of their knowledge to B.B.C. methods and equipment. Thus the Department does not attempt to duplicate the training provided by technical colleges and universities, but rather to supplement it. None of the courses are of more than twelve weeks' duration, so that all training must be intensive and essentially practical and non-mathematical in its approach.

There are two twelve-week courses; one of these is the initial entry course, designed to train new recruits (technical assistants), who have little or no previous radio training, to be useful members of the operation and maintenance departments. The other is a promotion course for selected technical assistants aiming to reach the engineering grades. This course is divided into two parts, the first covering broadcasting fundamentals, and the second part dealing with some aspects of B.B.C. practice and equipment.

Other courses of from one to three weeks' duration are operated to bring staff up to date on particular developments; one such course is that on colour television, now being run for senior staff. There are also conversion courses, designed to prepare staff for transfer from one branch of the service to another, and workshop courses.

Facilities exist for practical training on transmitters, recording, acoustics, aerials, television lighting, camera channels, studio control rooms, and continuity. Some of these facilities are for non-engineering staff, particularly Studio Managers, who are responsible for the technical control and balance of programmes.

* "B.B.C. Engineering Training Department," *J.Brit.I.R.E.*, **11**, p. 203, May 1951.

Apart from the specialist training provided by the Department, the Corporation is now considering an increase in the formal training facilities. At present, six students are selected from those who have attained the highest marks in the examination at the end of the promotion course, and these are sent on a Higher National Diploma sandwich course. It is possible that their number may be increased considerably. Owing to the nature of the work and the relative isolation of many stations there are difficulties in introducing apprenticeship training of the part-time day release type, and the Corporation does not at present recruit men of under twenty years of age. Present conditions may however change, and the Corporation does not rule out the possibility of recruitment under the age of twenty.

Twelve University graduates are recruited each year, and are given a post-graduate apprenticeship of two years' duration. Although their subsequent employment depends on their ability and inclination, they are normally recruited for the specialized departments such as Research, Design and Planning.

Specialist training techniques have been developed, and much thought and care has been given to the planning of laboratory demonstrations and practical work which clarify and supplement the normal theoretical approach. The Training Department produces a large proportion of the laboratory equipment used, also its own training manuals and text books, several of which have been made available to the public through publishing houses. The Technical Writing Section also regularly issues training supplements which are available to staff in the Corporation.

Engineers in the B.B.C. are given every encouragement to obtain additional qualifications and many do, in fact, qualify for membership of the professional institutions entirely on their own initiative, and in their own time. Those engineers wishing to attend at Technical Colleges are assisted wherever possible by re-arrangement of shift work and by employment at centres near to suitable colleges.

Apart from the training provided for B.B.C. staff, the Department also arranges courses of training for the staff of Commonwealth and overseas broadcasting organizations. Of particular importance is the work of training engineers who will eventually set up similar training facilities in the Colonies.

APPLICANTS FOR MEMBERSHIP

New proposals were considered by the Membership Committee at a meeting held on July 5th, 1956, as follows: 24 proposals for direct election to Graduateship or higher grade of membership, and 25 proposals for transfer to Graduateship or higher grade of membership. In addition, 37 applications for Studentship registration were considered. This list also contains the names of five applicants who have subsequently agreed to accept lower grades than those for which they originally applied.

The following are the names of those who have been properly proposed and appear qualified. In accordance with a resolution of Council and in the absence of any objections being lodged, these elections will be confirmed 14 days from the date of the circulation of this list. Any objections received will be submitted to the next meeting of the Council with whom the final decision rests.

Direct Election to Honorary Member

HOWE, Professor George William O., D.Sc., LL.D. *Glasgow.*

Transfer from Associate Member to Member

TURNER, Capt. Geoffrey Cater, R.N. *Portsmouth.*

Direct Election to Associate Member

AGARWAL, Major Arjun Deo, B.Sc., Indian E.M.E. *New Delhi.*
DE LAISTRE BANTING, Hugh. *Ixelles, Belgium.*
GARDNER, Flt. Lt. Norman John, R.A.F. *Bletchingley.*
GOODER, Alfred William. *Swindon.*
KASON, John. *London N.W.2.*
MORTON, Alexander Hugh, B.Sc.(Hons.). *Paisley.*
REES, William John, B.Sc. *Edgware.*
WEBB, Flt. Lt. Ernest Albert Norman, R.A.F. *Upavon.*
WILLMOTT, John William. *Lincoln.*

Transfer from Graduate to Associate Member

ELDRIDGE, Dennis Arthur George. *Waltham Abbey.*
HAMILTON, Jerome John, B.Sc.(Eng.). *Carshalton.*
LAX, Sqdn. Ldr. Bernard, R.A.F. *Northwood*
PROCTER, Anthony Charles. *Marsh Gibbon, Bucks.*
SCHILD, Rolf. *London N.W.2.*

Transfer from Student to Associate Member

CHAPMAN, David Stanley John, R.A.F. *Enniskillen.*

Direct Election to Associate

BOWER, John Bingham de Courcy. *Chelmsford.**
KYNASTON, John Alfred Charles. *Cardiff.**
TAYLOR, Wilfred Ernest. *London S.E.6.*
WRIGHT, Frank William. *Kirkuk, Iraq.*

Transfer from Student to Associate

PLEETH, Michael Jacob. *West Wickham.*

Direct Election to Graduate

CAIRD, John Aitchison, B.Sc. *Wellington, New Zealand.*
CZAJKOWSKI, Zbigniew, B.Sc.(Eng.). *London W.9.*
LARKWORTHY, Flt. Lt. Lionel John, R.A.F. *Hong Kong.*
MADELEY, Bryan. *Bournemouth.*
NEWLING, Robert Jefferson Grant. *Entebbe, Uganda.*
SECKER, George William. *Harrow.*
THOMAS, Flt. Lt. Caradog, R.A.F. *Watton, Norfolk.*
WINCHCOMBE, Thomas. *Gwbert-on-Sea, Cardigan.*

Transfer from Student to Graduate

BRONSTEIN, David. *Tel-Aviv.*
GEORGE, Panthradil Samuel. *Singapore.*
MUNRO, Kenneth Neil. *Giffnock, Renfrewshire.*
SAXENA, Mahinder Sahai. *Ferozepore.*
TURNER, Brian Robert. *Bermuda.*

STUDENTSHIP REGISTRATIONS

AKINYEMI, Isaac Olaonipekun. *London N.19.*
ANAND, Nand Kumar, B.Sc. *Bangalore.*
ANANTHASWAMY, Nagalapur Krishnamurthy, B.Sc., M.Sc. *Roorkee.*
ARABINDA BURMAN, B.Sc. *Serampore, West Bengal.*
ARJUN SINGH, M.Sc., B.Sc.(Hons.). *Udaipur, Rajasthan.*
ASTHANA, Omesh Chandra, B.Sc. *Bangalore.**
BHARDWAJ, Hari Sarup. *Nagpur.*
BHATLA, Shadi Lal. *Bombay.*
BOLSIUS, Ewald Petrus. *Queenscliff, N.S.W.*
BYNG, Raymond Edward. *Coventry.*
CHUNG, Allison Kaihee Lucius. *Toronto.**
DAS, Pulak, B.Sc. *Delhi.*
DAVIS, Geoffrey Harry. *Wellington, New Zealand.*
DIMSDALE, Ernest Edward. *London S.W.19.*
FRASER, William Morrison. *York.*
GREENHALGH, Daniel Joseph. *Coventry.*
HENG POH KIA, Thomas. *Singapore.*
KESHAV DUT. *Kishangarh, India.*
LEE CHING CHIN. *Hong Kong.*
MALPASS, Alan. *Sedgley, Staffs.*
MARSDEN, Jim. *Huddersfield.*
MENON, Aravindaksha P. *Bangalore.*
NATARAJAN, Ramakrishnan Iyer. *London N.W.6.*
NATESAN, Tanjore Sundara, B.E. *Madras.*
OJHA, Vinod Krishan, B.Sc. *Agra.*
PAI, A. Venkatramana, B.E. *Coventry.*
PAIS, Aloysius Francis, B.Sc. *East Croydon.*
POINTER, Geoffrey. *Norwich.*
PRITAM SINGH. *Bangalore.*
RANSLEY, John David. *Leigh, Lancs.*
REID, Alexander, M.A. *Ardrossan, Ayrshire.*
ROSARIO, Frederic N. J. *Bombay.*
RUBEN, Moshe. *Haifa.*
SURI, Gajindar Lal. *Simla.*
TEWARI, Man Haran Kumar. *Jaipur.*
WAGNER, George David. *Parramatta West, N.S.W.*
ZIJDEMANS, Leendert Johannes. *London W.3.*

* Reinstatement

TECHNIQUE OF MICROWAVE MEASUREMENTS*

A Discussion Meeting held in London on February 29th, 1956.

In the Chair: Professor Emrys Williams, Ph.D., B.Eng. (Member).

SUMMARY

The following are described: Power measurements using the torque vane wattmeter, Hall effect, thermistor and bolometer, and calorimetry (vaporization of liquid gases); attenuation measurements using rotary attenuators, cut-off attenuators, resistive film attenuators; general considerations of accuracy of measurement of attenuation; impedance measurements using directional couplers, swept-frequency reflectometers, slotted waveguide standing wave meters, "three probe line" method, substitution methods; frequency measurements using resonant cavities; echo boxes for assessment of radar performance; noise source measurements; particle accelerator measurements using perturbation techniques; dielectric loss measurements; mechanical construction aspects including materials used, and choke coupling. A select bibliography of thirteen relevant papers published in the *Brit.I.R.E Journal* is appended and numerous other references are given.

INTRODUCTION

E. M. Wareham (Associate Member)†

Power measurement.—Energy in any form is more readily converted to heat than anything else, and at radio frequencies it is converted to heat only too readily. Still, it is an ill wind that blows nobody any good, and when we want to measure r.f. power this is convenient because there are accurate devices for the measurement of heat. Fundamental power measurement at r.f. is, in fact, calorimetry, and the big problem is how to get all the r.f. energy into the calorimeter. Despite the best efforts of all those engaged in measurements of this nature, accuracies of better than 1 per cent. are not very often claimed. Incidentally, water does not seem to be a very good choice of fluid for this purpose, for, with a dielectric constant of about 80, there are serious problems both of absorption and leakage of energy. It occurs to me to wonder why there are no closed system calorimeters using a more convenient fluid.

The bulky size of calorimeters, and their inconvenience, leads to the application of temperature sensitive elements used in conjunction with low frequency a.c. or d.c. measurements of power. The use of the thermistor semi-conductors for this purpose seems to find the most general favour on our side of the Atlantic, and fine wire bolometers on the other side. The thermistor has the advantage of sensitivity, but, being a low temperature element, it is also very sensitive to ambient temperature. The bolometer, which is of course a high temperature element, has to have a very fine wire indeed if it is to be at all sensitive.

A device of considerable interest is the torque vane wattmeter due to A. L. Cullen of University College.‡ It was developed as a result of a study of radiation pressure in which a mechanical force is observed on a surface which is reflecting electro-magnetic energy. It is of intermediate sensitivity, and has an accuracy of about 1 per cent. at an r.f. power level of one watt.

In an interesting development of this device, suggested by C. M. Burrell,§ the waveguide in which the vane is situated is made to resonate. In fact it is made into a cavity resonator with a Q factor of perhaps 20,000 or more. The magnification of field strength thus obtained leads to a very considerable increase in the sensitivity of the device. The Q factor must be known, but cavities are quite stable if they are sealed, and it is possible to measure Q with an accuracy of better than 1 per cent.

* Discussion Meeting No. 11.

† W. H. Sanders (Electronics) Ltd.

U.D.C. No. 621.317.029.63/4:621.37.

‡ A. L. Cullen and I. M. Stephenson, "A torque-operated wattmeter for 3-cm microwaves." *Proc. Instn Elect. Engrs*, **99**, Part 4, pp. 294-301, December 1952.

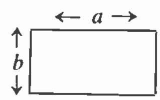
§ University of Cambridge, formerly Radar Research Establishment. Unpublished communication.

Attenuation measurements.—Fundamentally the standard of attenuation can be defined as the ratio of two power measurements, but this is rather slow and cumbersome and power meters do not normally have a wide range of measurement, nor are they very accurate. An alternative and rather better way is to define a standard of attenuation in terms of decay of field strength in a waveguide beyond cut-off. This latter method is a good one because the properties of the cut-off waveguide can be derived mathematically from its dimensions, so that ultimately a cut-off attenuator can be referred to a single fundamental unit, the standard of length. As a result of the convenience of this kind of attenuator, there is a tendency for power meters to be made with very restricted ranges and to use attenuators to extend them.

Attenuation and gain are, of course, dimensionless ratios, and in the decibel scale or the scale of nepers this ratio is the ratio of two powers. For instance it is quite incorrect to specify the voltage gain of an amplifier in decibels unless the input and output impedances are equal. A cathode follower may well have a gain of 40 db and a voltage amplification of 0·95. On the other hand, a filter working between its iterative impedances is free from this difficulty. This emphasizes the fact that when we specify an attenuation as so many decibels or nepers, it is necessary to specify the source impedance, the input and output impedance and the terminating impedance; unless this is done the statement is valueless because the attenuation will be different under different external conditions. In transmission lines we usually design attenuators as filters to operate between their iterative impedances. It is too easy to forget, nonetheless, that they must be operated in this way if they are to be accurate.

Impedance measurement.—Any discontinuous change in an otherwise uniform transmission line will, of course, result in a reflection of energy back towards its source. The ratio of the field components of the reflected wave to those of the incident wave defines a characteristic property of the discontinuity, i.e., its reflection coefficient. The magnitude of this is a unique property of the discontinuity. Its phase suffers a progressive transformation along the line.

When it is necessary to analyse such a discontinuity, we can separate these two components in a hybrid transformer, or we can measure their interaction one upon the other by examining the standing wave set up. The standing wave can be analysed either into the reflection coefficient or into a complex number which we call the normalized impedance. This parameter is of great use in the analysis of transmission line circuits, and it is in fact the ratio of two impedances Z/Z_0, the denominator being the characteristic impedance of the transmission line. Now apart from being dependent upon the electrical and magnetic properties of the medium within the wave guide, it is also a function of frequency and of the physical dimensions of the tube:



$$Z_0 = \sqrt{\frac{\mu}{\varepsilon}} \cdot \frac{\lambda_g}{\lambda} \cdot \frac{b}{a}$$

for a rectangular waveguide in the dominant mode.

From this we can infer straight away that a waveguide of given dimensions filled with a specified medium is a standard of Z_0 against which other impedances can be normalized.

In a standing-wave meter the slotted section is the standard guide. Similarly, a hybrid transformer—such as a directional coupler or a magic T or a rat-race—has got to be a symmetrical structure with a characteristic impedance equal to that of the transmission line being used. It is also possible to use either of these devices as comparators. In this case a measurement is first taken on a standard termination and then on the unknown. The ratio of the two measurements is the impedance of the unknown, normalized to that of the standard. The errors, provided they are consistent, of the measuring instruments themselves are eliminated.

Frequency measurement.—Standardization of frequency has been carried to a very high degree of accuracy by the work of L. Essen at N.P.L. Dimensionally, frequency is $[T^{-1}]$ and therefore frequency measurement is dependent upon time measurement. Time is, of course, standardized by astronomical observation and intervals of time are usually interpolated by means of clocks controlled from quartz crystal oscillators. Suitable valve circuits are used to "count down" the frequency of the crystals, each stage being recorded by suitable electrical counters until a sufficiently long interval is obtained for mechanical counting. The electrical oscillations generated by these crystals can be multiplied, of course, as well as divided in frequency, by the use of suitable circuits, and it is possible, for example, to generate detectable power at frequencies as high as 10^{11} cycles per second directly related to a 100 kc/s crystal. A possible arrangement would be multiplication in decades from 100 kc/s, by means of amplifiers, harmonic generators and filters to cut out the unwanted harmonics. The amplifiers above about 1 kMc/s would be travelling wave tubes, and harmonic generators would be crystal diodes.

It is also possible to add and subtract frequencies by making use of the technique of amplitude modulation and frequency selection. We remember that since an amplitude modulated wave has two sidebands, which are respectively the sum and difference frequencies, a suitable filter will select the sum and eliminate the carrier and the other sideband. By combining the operations of adding and multiplying it is possible to generate frequencies controlled by, and/or referred to, precise crystal oscillators. These oscillations can be compared by heterodyne methods with unknown frequencies in the microwave spectrum, and the accuracy of the crystal is virtually the accuracy of the measurement.

Filtering is conveniently carried out by tuneable cavity resonators. If they are fitted with micrometers and calibrated they are suitable for use directly for measurements of frequency. These devices can give accuracies between 3 parts in 10^3 and 3 parts in 10^5.

Measurements on Systems.—If we now turn to complete systems it is rather pleasing to think of the improvement in measurement ease which has been brought about by the use of the gas-discharge noise source for measurement of receiver noise factor. The equivalent parameter for the transmitter—pulse power output—can be quite easily monitored with the neon glow height indicator. Another device which can be very useful in the analysis of the overall performance of pulse radars, is the echo box high-Q cavity resonator, which is capable of being fully analysed and measured with sufficient accuracy to be used, in conjunction with circuits in the radar itself, to give an absolute measurement of the radar overall performance factor. Defining this as the ratio of the transmitter peak power output to the noise power at the receiver input, it is clear that we thus obtain a good indication of the effective range of the radar.

POWER MEASUREMENTS

Dr. J. Brown*: The torque vane wattmeter, using a single vane, is basically an electrostatic voltmeter and the main significance of this is that it indicates the square of electric field strength which is a function of both the forward and backward power flows. One possible way in which the most valuable property of wattmeters of the radiation pressure type, namely the ability to measure power without absorbing it, may be used is shown in Fig. 1. Power from the oscillator is fed into arm 1 of a three-arm circulator, a non-reciprocal element which has the property that signals can travel only from arm 1 to 2, 2 to 3, or 3 to 1. The power from the oscillator thus arrives at arm 2, where it can be reflected by a radiation pressure wattmeter, and then it travels to the load in arm 3. Any reflected wave from the load is coupled back directly to arm 1 and leaves the wattmeter reading unaffected. This arrangement thus enables the forward power from oscillator to load to be continuously monitored without the need to extract any of the power by a directional coupler as is more usually done. An extension of this idea is the use of a four-arm circulator, in which the power reflected from the load could be directly measured in the fourth arm.

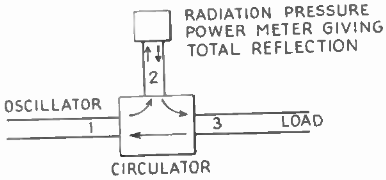


Fig. 1.—Use of radiation pressure power meter to measure without absorbing power.

* University College, University of London.

A further type of power measuring instrument is at present being developed by Professor Barlow and makes use of the Hall effect.* If a small piece of germanium or other semi-conductor is placed in a rectangular waveguide as in Fig. 2, a current flows through the germanium in the direction of OX parallel to the r.f. electric field. The r.f. magnetic field in the direction OY causes a bending of the lines

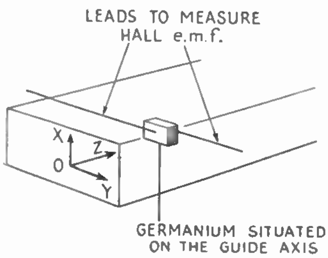


Fig. 2.—Hall effect power meter using germanium.

of current flow resulting in a Hall e.m.f. being developed in the direction OZ. The magnitude of this e.m.f. is proportional to the power flow according to the theory so far developed, and the sign of the e.m.f. changes if the direction of the power flow is reversed. When a standing wave is present, the Hall e.m.f. is dependent on the net power flow towards the load. One possible application of an instrument using this principle, at present being investigated, is to monitor the variation of power in short pulses.

P. F. Mariner†: There are two torque vane wattmeters available, each of which gives the possibility of correcting for errors caused by reflections from the terminal load. One has two vanes spaced by a quarter wave length and carried on the same suspension so that the reflected and incident energy are in opposite phase at the two vanes. The other instrument has one vane and correction is made by making measurements with the load at two positions spaced a quarter wavelength apart.

I. A. Harris (Associate Member)‡: It does not appear to be sufficiently widely known that all the microwave power absorbed by the thermistor mount, as judged by a v.s.w.r. and power measurement, does not appear in the active part of the bead. Actually, comparison with calorimeter methods has shown§ that about 90 per cent. of the power absorbed by the mount appears in the active element, so that the apparent power measured by the usual d.c. substitution method has to be multiplied by a factor approximating to 1·1 to obtain the true power absorbed.

L. C. Walters (Associate Member)‖: Regarding the accuracy of power measurement using a thermistor, I think there is some fairly strong evidence to suggest that at least some of lost energy appears as heat in the material surrounding the thermistor. Originally thermistors were mounted in glass beads, and, while the problem has not yet been overcome, the use of low loss ceramics rather than glass has achieved slightly better results.

J. A. Lane**: Figure quoted of errors of 10 per cent. for thermistors calibrated in terms of d.c. power are typical of thermistor performance at a wavelength of 3 cm, but at a wavelength of 10 cm the error is normally not more than about 2 per cent. for waveguide mounts. At millimetre wavelengths, on the other hand, thermistors may indicate less than the true power by as much as fifty per cent.

The use of film bolometers in microwave power measurement is a technique which seems to have received only a limited amount of attention in the past. In particular, it is not generally appreciated that it is possible to terminate a waveguide with a resistive film which occupies only a small portion of the transverse plane. For example, a film 0·35 cm wide of surface resistivity 180 ohms per square, when located in the centre of 3 cm band waveguide and followed by a reflecting piston at a distance of $\lambda_g/4$ has an input-voltage standing-wave ratio of the order of 0·9 over a 10 per cent. bandwidth. The power in the waveguide can thus be measured by observing the change of resistance in, say, a platinum film deposited on thin mica. Alternatively, the steady-state temperature rise can be observed by means of one or more thermocouples attached to the film and connected to a

* H. E. M. Barlow and L. M. Stephenson, "The Hall effect and its application to power measurement at microwave frequencies," *Proc. Instn Elect. Engrs*, **103**, Part B, pp. 110-112, January 1956.

† Elliott Brothers (London) Ltd.

‡ Ministry of Supply, A.I.D. Laboratories.

§ J. A. Lane, "The measurement of power at a wavelength of 3 cms by thermistors and bolometers." *Proc. Instn Elect. Engrs*, **102**, Part B, pp. 819-24, 1955.

‖ The Plessey Co. Ltd.

** Radio Research Station.

galvanometer. In both techniques the system is calibrated in terms of d.c. power. As an indication of the order of sensitivity, it is possible using a single copper-eureka thermocouple to obtain, at a distance of 1 metre, a scale deflection of approximately 10 cm per milliwatt at a wavelength of 3 cm. Preliminary experiments using a simple instrument of this kind have been carried out recently at the Radio Research Station.

In connection with Dr. Brown's remarks on the cost of many microwave components (see p. 393). I would point out that it is possible to construct a power meter of this type, using carbon-coated paper as the resistive film, for which the total cost (excluding the d.c. instruments involved) is almost negligible.

P. Andrews (Graduate)*: There are many experiments in basic physics which are directly applicable to the microwave field. For instance, an extension of Bunsen's ice calorimeter experiment uses the microwave energy to melt ice at its melting point: by measuring the change in volume, the microwave power may be calculated. A variation of this idea is to use the microwave energy to supply the latent heat of vaporization of a liquid at its boiling point; the volume of gas evolved is related to the energy used. With liquid helium, 1 milliwatt would evolve 15 cm^3 of gas per minute, while for oxygen, 0·2 cm^3 per minute would be liberated per milliwatt. Measurements have actually been made with such an apparatus using liquid oxygen as the medium, and it has been demonstrated that this method can form the basis for a workable system, although at the moment thermal leakages into the apparatus create a very difficult problem.

E. M. Wareham: For thermistors and bolometers, the sort of figures I have heard mentioned are: at S band, about 2 per cent. to 4 per cent. error for thermistors; at X band about 4 per cent. to 8 per cent. error; and at Q band rather higher, in the region of 10 per cent. to 20 per cent. It may be that with this particular bead which is the S.T. & C. "U" type, encapsulated in glass, there is a high loss in the glass for the Q band region. Figures for bolometers of American type are about 2 per cent. to 4 per cent. at X band.

ATTENUATION MEASUREMENT

P. F. Mariner: As a reference of absolute attenuation, Mr. Wareham mentioned the cut-off attenuator whose attenuation is a linear function in decibels with length within the waveguide. At microwave frequencies this device has disadvantages and I suggest that measurement in terms of angle be used instead by using the rotary attenuator.

This consists of a length of circular waveguide divided into three sections, with rectangular to circular transitions at each end. Each of the circular pieces contains a resistive vane which is arranged to absorb substantially all of the energy polarized with its electric vector parallel to the resistive surface. The vanes in the end circular sections are parallel to the broad face of the rectangular waveguide. This makes the instrument a reciprocal device and ensures true conversion from the dominant mode in rectangular waveguide to the dominant mode in circular waveguide.

If the vane in the centre section, which is rotatable, is at an angle θ to the end vanes, then resolving the incident electric field (E_0) parallel and perpendicular to the centre vane, the parallel component ($E_0 \sin \theta$) is absorbed and the perpendicular component ($E_0 \cos \theta$) is transmitted. The polarization is therefore effectively rotated through the angle θ. At the last vane the electric field ($E_0 \cos \theta$) is at an angle θ to the vane and therefore resolving again the transmitted component is $E_0 \cos^2\theta$. The attenuation in decibels is therefore given by $20 \log \sec^2\theta$.

If mechanical tolerances are reasonably maintained absolute attenuation is given by simple computation from the measured angle.

B. E. Kingdon said that a disadvantage of the piston attenuator was the initial attenuation of 20 db which it introduced. The rotary attenuator appeared to cover the range from 0 db upwards which was a useful feature in certain applications.

I. A. Harris: Attenuation should be regarded as an insertion loss, which, as a parameter of the attenuator, is properly defined as the loss in the load brought about by inserting the attenuator between the source and the load,

* Admiralty Signal and Radar Establishment.

both source and load being matched to the nominal characteristic wave impedance of the system. When the source and load are not matched to the nominal characteristic impedance, the actual loss due to insertion of the attenuator is different from the calibrated attenuation as defined above. As a general guide, provided the matching shows a v.s.w.r. of about 0·85, the error in attenuation is not worse than $\pm 0{\cdot}2$ db on account of the mismatches. An accuracy of 0·05 db or better, however, demands a higher degree of matching if the precise values of attenuation are to have an absolute meaning. This is illustrated with a simple formula for the actual insertion loss:—

$$A = 10 \log_{10} \left| \frac{(1-\rho_S\rho_1)(1-\rho_L\rho_2)-\tau^2\rho_S\rho_L}{(1-\rho_S\rho_L)\tau} \right|^2 \quad \ldots\ldots.(1)$$

which is the loss brought about by the insertion of the attenuator between a source of voltage reflection coefficient ρ_S and a load of reflection coefficient ρ_L. In this formula, ρ_1 and ρ_2 are the input and output reflection coefficients of the attenuator respectively, with the opposite end matched in each case. τ is the voltage transmission coefficient of the attenuator, which is the same for propagation in either direction with a passive element satisfying the reciprocity theorem. When the source and the load are matched, ($\rho_S = \rho_L = 0$) then:—

$$A = 10 \log_{10} \left| \frac{1}{\tau} \right|^2 \text{ decibels} \quad \ldots\ldots\ldots(2)$$

and the insertion loss is clearly dependent only on the attenuator. All the ρ's and the τ are complex, but an estimate of the maximum uncertainty on account of mismatches can be found from eqn. (1) with a knowledge of their magnitudes alone. Eqn. (1) was first derived by Beatty.*

Indirectly, the same effect of mismatches applies to a piston attenuator assembly. While the main variation of attenuation is produced by the movement of the pick-up probe in the cut-off tube, there is also a variation due to impedance changes resulting from the flexing of the coaxial line output connector. This latter variation can invalidate the calibration of a high precision cut-off tube. The effect is still present when the outer end of the cable is padded with a fixed lossy attenuator to improve the output match, otherwise often poor on account of the unavoidably reactive pick-up loop or probe.

B. Rogal†: The main feature of the cut-off attenuator is the fact that it can be maintained to a very high degree of accuracy, the latter being mainly a function of mechanical stability. However, at the higher microwave frequencies the guide dimensions are becoming so small that even with electroforming techniques there are unpredicted errors in attenuation per unit length and this failure becomes important. In general such attenuators will need checking and this can be done by using cut-off attenuators at intermediate frequencies of the order of megacycles.

One of the main advantages in this band is that we can amplify the signal passed through the attenuator; the other advantage is that as far as mechanical stability is concerned, dimensional limits in electroformed tubes of 1 in. or 2 in. diameter can easily be maintained to better than 2 parts in 10^4. Of course, a very precise drive is necessary and normally a micrometer would be used.

However, in rapid measurements, when calibration check of the attenuator is required, a micrometer drive which has to cover a number of inches (such as three or four at the check points) is not a very successful solution. There were various methods attempted to overcome this inconvenience, one of which was to produce definite steps of, say, a quarter of an inch travel at a time and to interpolate with a vernier. These steps were produced by means of specially selected rollers or suitably shaped profiles. Rollers can be selected and maintained with extremely close tolerances.

One successful method of accurate calibration using a low frequency attenuator as reference is as follows: two oscillators are 100 per cent. square-wave modulated, one oscillating at 60 Mc/s and the other at r.f. in such a way that one is on when the other is off. The 60-Mc/s signal is coupled through the standard attenuator into the latter stages of the 60-Mc/s i.f. amplifier of a r.f. receiver. The signal from the r.f. oscillator is coupled through the unknown r.f. attenuator to the receiver input. The output of the i.f. amplifier thus contains a train of pulses from the 60-Mc/s source, the spaces of which are filled with a second train of

* R. W. Beatty, "Mismatch errors in the measurement of ultra high frequency and microwave variable attenuators." *J. Res. Nat. Bur. Stand*, **52**, pp. 7-9, January 1954.

† Wayne Kerr Laboratories Ltd.

pulses from the r.f. source after frequency conversion in the receiver. The sum of these at the second detector output is a d.c. level with a square wave superimposed.

When the two signals are of equal amplitude the square wave disappears and it is thus possible to make very accurate comparisons both by visual display (c.r.t.) and by meter (phase sensitive rectifier).

When generating these signals it is important to maintain a power output stability which is better than the required accuracy of measurement in both oscillators. One method which has been found satisfactory in stabilizing the low frequency oscillator is to rectify the output and to apply the amplified voltage as negative feedback into the h.t. supply. This has been found to achieve a stability better than 0·01 db over periods of several hours.

Signal generators usually cover a power range between 10^{-3} and 10^{-13} watts. At these levels it can be claimed that the use of the technique described has led to measurements of incremental attenuation with an accuracy of the order of 0·015 db. The equipment is sufficiently sensitive to detect the presence of r.f. signals at power levels below 10^{-16} watts and so can provide useful information in the investigation of r.f. leakage effects and other extremely low level phenomena. At such low power levels the i.f. signal is considerably less than the noise and the c.r.t. display is less effective than the phase sensitive rectifier circuit, where the bandwidth can be very much reduced.

J. W. Sutherland*: Referring to resistive attenuators, the H_{11} circular double absorbing type has the great advantage that it is both independent of frequency and self calibrating; but it does suffer to some extent in that the scale is wide open at the lower end and very close at the upper end, so that it is inconvenient for general laboratory use on that account. At longer wavelengths than 3 cm it can become rather bulky. Nevertheless, it is a very good instrument.

In the attenuator using a resistive film lying in the E plane and relying upon the sinusoidal distribution of field strength across the guide, problems have been encountered giving an appearance of resonance. Fig. 3 shows such an effect; this effect has been found to be a function both of the shape of the element and of the film resistivity.

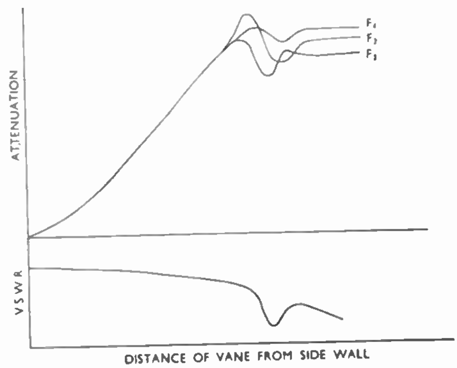


Fig. 3.—"Resonance" in a resistive vane attenuator.

* Marconi's Wireless Telegraph Co., Ltd.

D. A. J. Walliker†: I shall confine my remarks to that type of dissipative attenuator in which a glass plate coated with a resistive film is mounted in rectangular waveguide with the surface of the plate parallel to the narrow wall of the guide. The attenuation is negligible when the plate lies close to the narrow wall, and increases as the plate is moved towards the centre of the guide. Because of the negligible insertion loss, and because rectangular waveguide is employed, this type of attenuator is often more convenient than a cut-off attenuator for many microwave measurements, and when carefully built and calibrated, can yield measurements of extremely high accuracy. The same type of attenuator also finds extensive use in microwave radar and communication equipment, for the adjustment of power levels and for minimizing the effects of mismatching.

It is of first importance that the attenuation shall be unaffected by variations in ambient temperature and humidity, and techniques have now been developed for the manufacture of coated glass attenuating elements which have negligible temperature-coefficients of attenuation, and which are unaffected by exposure to extreme climatic conditions—for example, to 2,000 hours of exposure at 100° C ambient temperature, and to 84 days of tropical cycling between 20° C and 35° C at 95 per cent. relative humidity. Further, these techniques yield a high degree of reproducibility in production between one element and another of the same type.

It is desirable also that the attenuating

† Decca Radar Ltd.

element shall present a good match to the input and output waveguides, shall cover a wide frequency-band with small frequency-sensitivity of attenuation and shall have a substantially linear attenuation characteristic. These desiderata are now achievable, and recent studies have resulted in the evolution of a rapidly convergent design procedure whereby an optimum choice of the various parameters of the element can be made within the limits of manœuvre set by the user's specification. Researches now in hand are designed to place this procedure on a rigid mathematical basis.

The performance of an attenuating element can, to a first approximation, be related to its overall length, and when only a short length of waveguide is available for the accommodation of the element—as is often the case in radar and communication equipment, and in portable test-gear—then the element design must be the result of a compromise.

Experience has shown that it is then best to restrict the bandwidth to a range of frequencies not much greater than that applicable to the remainder of the equipment. In this way, propagation in higher-order modes within the attenuator can be avoided, and good matching and linearity, together with a low frequency-sensitivity of attenuation, can still be obtained in a limited length of waveguide.

Finally, let me illustrate a point which is not always clearly appreciated by the user. Consider two different attenuating elements, A and B, having the characteristics shown in Fig. 4, and let these elements be fitted into attenuator mounts of equal mechanical resetting accuracy.

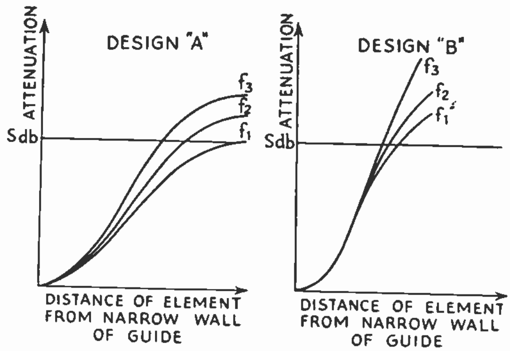


Fig. 4.—Illustrating the need for "matching" resistive vane attenuators to achieve broad-band performance.

Clearly either attenuator could be used to produce the attenuation level S db, but if the frequency is not one of the frequencies of calibration (f_1, f_2 or f_3), then a numerical interpolation is required. Now since design A has a more extensive region of linearity than B, the reading and resetting accuracy of the former design will be practically constant up to the level S db. The interpolation errors of B will also be less than those of A, because of the smaller frequency-sensitivity of attenuation, and further, these errors will be very closely proportional to the attenuation level up to S db. Thus despite the greater slope of its characteristics, design B will yield an overall error (instrumental error plus interpolation error) which is not only significantly less than the overall error of design A but is an almost constant percentage of the attenuation level.

IMPEDANCE MEASUREMENT

Dr. J. Brown: Mr. Wareham has given a definite value for the impedance of a waveguide although it is generally accepted that there is no unique way of defining this quantity. There are many ways in which an impedance can be defined and consistent results can be obtained for any of the values obtained, provided the same definition is adhered to during any set of calculations. I feel that Mr. Wareham should qualify his value by stating under what conditions it is to be used. There is a considerable risk of confusion if a specific value is quoted as being applicable in general: it might be thought, for example, that the reflection coefficient can be calculated from the normal transmission line formula when the waveguide is connected to a transmission line such as a coaxial cable for which the impedance is uniquely defined. This is not true.

It seems appropriate in a general discussion of the present kind to raise the question of the types of instrument which should be developed for microwave measurements. Considerable effort has been devoted to the development of precision instruments capable of yielding results of very high accuracy. It appears to me, however, that there is a corresponding need for equipment at more modest prices, which should be capable of giving reasonably accurate results but be simple to operate and maintain. The most obvious type of measurement for which such equipment is desirable is that of standing wave ratio. The number of occasions on which relatively crude values are needed over a band of wavelengths must lead to an expenditure of time and energy in research laboratories which could well be used in other ways. The reflectometer mentioned by Mr. Wareham provides the most convenient form of instrument for this purpose and many laboratories have developed apparatus using this method to give plots of standing wave ratio against frequency. I am surprised that no complete equipment of this type is available commercially. It has been suggested that the reflectometer may in fact be capable of greater accuracy than the more commonly used standing wave indicator but I am doubtful if this is true of present instruments.

E. M. Wareham (*in reply*): The definition of Z_0 which I gave is one which describes a characteristic property of a uniform rectangular tube propagating only the dominant mode. This definition contains all the parameters which control the disturbance arising at the junction of two dissimilar waveguides of this form. It has a value also in ascertaining the magnitude of the reflected wave which would be excited at such a junction.

P. F. Mariner: I would like to refer to an instrument which is not a slotted waveguide section, but is more akin to a directional coupler. The magnetic field of the dominant mode in rectangular waveguide is formed in loops in planes parallel to the broad face. There is a point in the broad face, therefore, at which the magnetic field has a rotating vector of constant amplitude. A circularly polarized wave will be coupled through a hole at this point and the electric vector in a circular waveguide placed above the hole will rotate in the same sense as the magnetic vector in the main waveguide. If in the main waveguide there is a wave travelling in the opposite direction causing a standing wave, the electric vector of the circularly polarized wave to which it couples will rotate in the opposite sense to that of the first wave. These two electric fields will add to give an elliptically polarized wave the ellipticity of which is equal to the standing wave ratio in the main waveguide. The phase of the standing wave referred to a plane in the main waveguide, is related directly to the orientation of the ellipse referred to the same plane. A rotating plane polarized detector can be used to analyse the field in the circular waveguide. This system could be mechanized and the output shown on an oscilloscope.

I want next to describe an instrument whose accuracy can be, with swept frequencies, very comparable with that of a standing wave indicator. At single frequencies it is possible to achieve even better accuracies than with a standing wave indicator because the equipment contains a facility for backing-off errors arising within the instrument. The waveguide circuit starts with an oscillator of frequency f and whose power is divided. One part goes to the component under test and is then reflected (or transmitted) to form one input to the balanced mixer; the other part passes through a motor driven Fox type rotary phase changer and forms the other input to the balanced mixer. The phase changer produces, effectively, a local oscillator frequency of $f+\delta f$. Since the microwave oscillator is arranged to have a constant output and the mixer is operating on the linear part of the crystal characteristic the output from the balanced mixer, at frequency δf, has a phase and amplitude coherent with that of the reflection (or transmission) coefficient of the component under test. A generator is mechanically coupled to the rotary phase changer to give a reference signal frequency δf with which the output from the balanced mixer is compared. Two signals in phase quadrature are obtained from this comparison and are used in a cathode ray tube to give an Argand diagram of the reflection (or transmission) coefficient or, of course, complex impedance on the circle diagram.

A similar system can be used for measuring the transmission coefficient of aerials anywhere in the radiation field although the phase information is most useful in the near field. In this application, the information is used in servo systems to give a record on charts instead of on the cathode ray tube.

I. A. Harris: Regarding standing wave indicators, the accuracy of measurement depends largely on the way in which they are used. For instance, given a satisfactory uniform slotted waveguide and probe motion, it is well known that probe reflection is a source of error, affecting particularly the peaks of the wave pattern. Probe reflection is minimized by tuning and matching the probe and detector. (Matching enables a given deflection to be obtained with less probe penetration than otherwise.) To do this properly over a wide band and to have the facility of varying the probe insertion, three adjustments are necessary. Usually, only two adjustments are provided—probe tuning and probe penetration. It is also worth while recalling that the effects of residual probe reflection can be minimized by measuring only at the minima of the standing wave pattern, provided the v.s.w.r. differs greatly from unity. Although mentioned in papers and textbooks, these points cannot be emphasized enough. The same applies to the calibration of the detector to an accurate square law, not at a half power point only, but over the whole range of the indicator scale used. With these points given careful attention, the slotted waveguide standing wave indicator is an accurate absolute instrument reliable for measurements of many kinds.

L. C. Walters (Associate Member): I do not think there is any doubt that a solution of Maxwell's equations, if possible, is the best approach. However, there are in practice a great many microwave boundary value problems which are completely insoluble except as approximations, even with the aid of large scale computers. One case in point is the circular standing wave detector: nobody, as yet, has precisely solved Maxwell's equations for a circular bend in a rectangular guide, so that while, for a first approximation—maybe for production testing—the circular standing wave detector would be admirable, it would be viewed with some suspicion by some people for precise measurements. One of the best known impedance measurement techniques is Feenberg's (approximate) method, in which one moves shorting plungers various distances and takes a large series of impedance measurements. The solution is a laborious process, and in the long run its accuracy depends significantly on the accuracy of the Smith chart.

As an application of this procedure I will describe a problem in which I was interested a few years ago. Having only measuring equipment in one size of waveguide, I wished to find the impedance of three taper transitions from this to another size of waveguide. The technique used was to couple pairs of transitions back to back, terminating with a good matched load and measuring the input v.s.w.r. With each combination of transition, v.s.w.r. measurements were made using several different lengths of the larger guide to separate them (see Fig. 5), the length being approximately $\frac{1}{16}$, $\frac{1}{8}$ and $\frac{1}{4}$ of the appropriate guide wavelength.

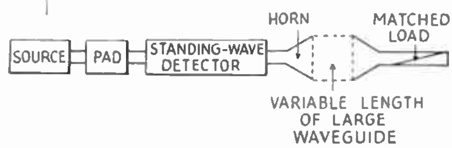


Fig. 5.—Measurement of impedance of waveguide transitions.

In measuring the standing wave ratio due to two mismatches r_1 and r_2, the worst s.w.r. is $r_1 r_2$ and the best is r_1/r_2 or r_2/r_1. The general form of standing wave patterns is a more or less smooth curve of roughly the shape illustrated in Fig. 6. The number of sections which could be interposed (three) gave seven points only, and these were chosen to be about $\frac{1}{16}\lambda$ apart, as indicated in Fig. 7. The estimated curve gives the maximum and minimum s.w.r.'s, from which the values of r_1 and r_2 were deduced and converted to voltage reflection coefficients.

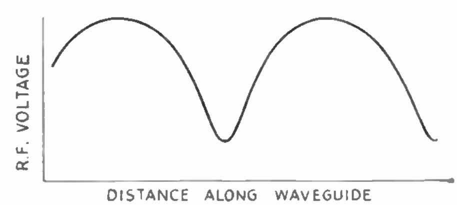


Fig. 6.—Typical standing wave pattern.

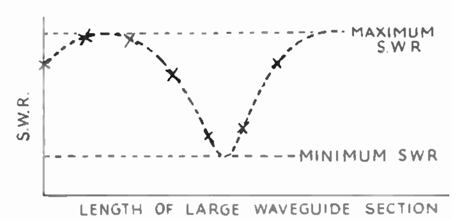


Fig. 7.—Experimental determination of standing wave ratio.

For each mismatch there were, in effect, four determinations and in each case these coefficients agreed to one part in the fourth place in the reflection coefficient. The method thus has considerable accuracy although based on only seven points.

A variant of this method which takes a lot of the labour out of the measurements has been evolved by D. R. Persson. Suppose, for example, we have a coaxial waveguide transition. We cannot get a perfectly matched load, but we do want to know what sort of match the transition would present to the waveguide if the coaxial line were matched. The equipment required consists of just this imperfect load in conjunction with a standing wave detector. We adjust the position of the load in the coaxial line to get the worst possible standing wave ratio (in practice the process is rapidly convergent and takes only a very short time). The reading and position of the standing wave detector in the position of the minimum v_1 are noted and it is then moved to the maximum v_2. Leaving the probe in that position the load is readjusted to obtain a minimum in this position, v_3. We then revert to the first position and measure v_4; it does not follow that the new reading in that position is necessarily going to be a maximum of the new distribution.

Using s.w.r.'s less than unity, we have for the case of the worst s.w.r., $v_1/v_2 = r_1 r_2$ where r_1 refers to the transition and r_2 to the load. The best s.w.r. will be r_1/r_2 or r_2/r_1 depending on which is the larger. Thus we have r_1/r_2 or $r_2/r_1 = v_3/v_4$ or v_4/v_3 depending on whether v_3 is less than or greater than v_4.

Hence one of the desired values r_1 and r_2 is given by $\sqrt{\dfrac{v_1 v_3}{v_2 v_4}}$ and the other by $\sqrt{\dfrac{v_1 v_4}{v_2 v_3}}$. To determine which is which we note that if the coaxial load had been perfect, i.e. $r_2 = 1$, then v_1 and v_2 would have been equal respectively to v_4 and v_3 since movement of the load would not affect the readings in this case. In this case the first of the above expressions becomes unity and hence gives r_2.

$$\text{Thus } r_2 = \sqrt{\frac{v_1 v_3}{v_2 v_4}} \quad : \quad r_1 = \sqrt{\frac{v_1 v_4}{v_2 v_3}} .$$

The phase of the mismatch is obtained from the fact that the position of the minimum v_1 is the same as that which would be obtained if the coaxial load were perfect.

It has been assumed that the values v_1, v_2, v_3 and v_4 represent r.f. voltages. If as will usually be the case, the measurement is made using a square law (e.g. crystal) detector then the square roots in the above should be replaced by fourth roots.

A complete theoretical justification is too long to give here, but the method can be rigorously justified. Unlike, for example, Feenberg's method which is only valid for small mismatches, this method is valid for all orders of mismatch, is highly accurate, and is certainly very much simpler than most comparable methods.

B. E. Kingdon (Graduate)*: The "three probe line" technique of impedance measurement has been successfully applied to coaxial lines in the v.h.f. band†, and is quite capable of adaption to waveguides (Fig. 8). At frequencies

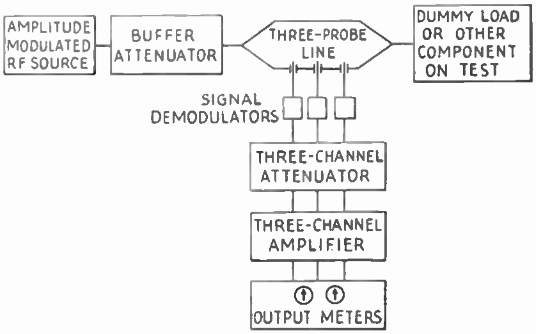


Fig. 8.—Layout of "three-probe line" test circuit.

of the order 100 Mc/s, the probes may be conveniently spaced one-eighth of a wavelength apart. Adjustment of the penetration of each probe allows equal sensitivities to be obtained, while amplitude modulation of the source enables a sufficiently small degree of penetration to be used, where subsequent amplification of a high standard is necessary.

The signal from each probe is demodulated by a valve-detector and fed into one section of an accurately calibrated three-channel attenuator unit. This is coupled to a three-

* Atomic Energy Research Establishment.

† A. Bloch, F. J. Fisher, and G. J. Hunt, "New equipment for impedance matching and measurement at very high frequencies," *Proc. Instn Elect. Engrs*, **100**, Part III, pp. 93-99, March 1953.

channel amplifier, the outputs of which are taken via bridge-meter rectifiers to two centre-zero meters, connected so as to compare the signals from each pair of probes.

When making a measurement, the probe sensitivities are brought initially to equality, the line being terminated in a matched load. The component or line under test is then substituted for the load, and the attenuators are adjusted to give zero deflection on each output meter. Comparison of the attenuator settings on a specially constructed diagram enables the modulus and argument of the reflection coefficient present in the line to be accurately determined.

C. W. Miller (Associate Member)* described an automatic plotting device for standing-wave ratios in which a slotted guide was effectively bent into a semi-circle so that the probe could move with rotary motion. The probe is mounted in a drum and driven round, triggering the time base on a cathode ray oscillograph to give a continuous plot.

FREQUENCY MEASUREMENT

B. Rogal: More emphasis is placed every day on producing equipment which will cover very wide bands. As far as the instrument using a waveguide is concerned (such as a standing wave indicator), it is usually limited to the actual band for which the guide has been primarily designed. This does not apply to more general instruments such as wavemeters. In such cases successful attempts have been made to obtain wide-band coverage of the order of 2½ to 1. There is of course a price to pay in such a wavemeter, namely the resolution and ultimate accuracy. However, when the wavemeters are to be used in investigating new techniques involving, for example, travelling wave tubes, the wide-band aspect is extremely important. One of the other requirements of such wavemeters is that they must not respond to spurious resonances.

Wide-band wavemeters operating in the E_{010}-hybrid mode, originally due to Dr. Essen, have been extended to cover bands right up to 40,000 Mc/s. The main response curve is fairly linear, linearity being better if no coupling loops project into the cavity.

The lower frequency bands use coaxial low-pass filters, so that the responses to frequencies above the desired range are eliminated. The nature of the spurious responses can be best appreciated with the aid of Fig. 9. This shows the main desired response in the hybrid mode, the undesired responses, the action of low-pass filter, and the configuration of the cavity and short-circuiting lines round the tuning plunger. At least one spurious resonance will occur when the length of the tuning plunger, in combination with a small reactance jx, will produce effectively a $\lambda/4$ line. Other resonances will be due to waveguide modes being excited in the coaxial line when the insertion of the plunger into the cavity is considerable. The identification or suppression of unwanted resonances in the cavity itself is not easy. The best solution is to insert low-pass filters in the input connector to the cavity, so that the undesired responses are strongly attenuated.

When the cavity is fed directly from the

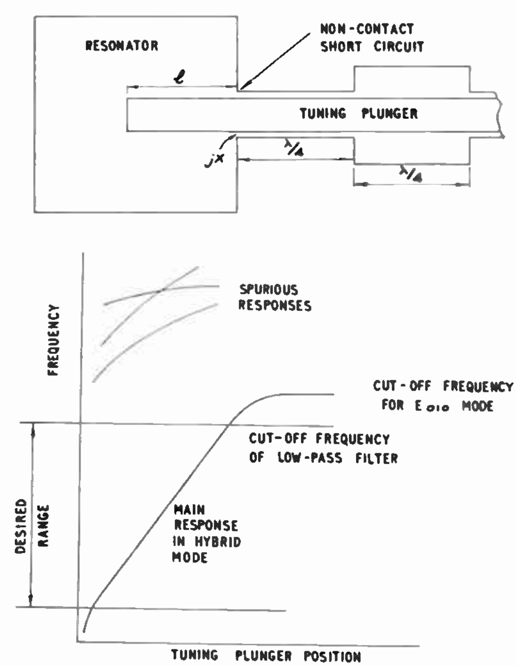


Fig. 9.—Cavity resonator type of frequency meter and typical calibration.

* Metropolitan-Vickers Electrical Co. Ltd.

waveguide, as is the case in the Q-band wavemeter (23–41 kMc/s) a frequency resolution of 10 Mc/s is obtained and this is adequate for wide-band work.

In this resonator the solution to eliminate spurious responses is a proper choice of the ratio of the tuning plunger to the cavity diameter in such a way that the unwanted responses occur at frequencies much higher than the required upper frequency limits.

MEASUREMENTS ON COMPLETE SYSTEMS

B. Rogal: Mr. Wareham mentioned the echo boxes which are used to test overall performance of the radar system and one of the requirements is that the Q of the cavity be measured to an accuracy of better than one per cent. Higher discrimination than this is needed, so that the cavities can be produced with good uniformity. An instrument has been developed, giving this degree of accuracy and as far as it is known it is the most reliable means of measuring the Q's of echo boxes.

This instrument measures the exponential decay rate of free oscillations set up in the resonator, and starts with two pulses decaying exponentially, one due to the resonator, the other due to the standard capacitor discharging into a resistor. Two pulses are generated (and later integrated), the widths of which are proportional to the rates of decay of the initial pulses. For actual connections of the cavity to measure the Q, two methods are used. There is a provision for a simple transmission measurement and a second for measurement of resonators with one coupling aperture.

J. W. Sutherland: There are two types of noise source in general use. In one the source is matched to the guide only when the tube is excited, and is inconvenient to use, because the noise source must be removed or switched with a wave guide switch and a dummy load substituted to get a reliable reading. There have also been developed noise sources matched to the wave guide, both hot and cold. In order to be accurate this type of noise source must have a high through attenuation when the tube is struck, and a low through attenuation when the tube is not struck.

There is an expression, to give the effective noise temperature, T_e, in terms of the hot loss L_t the cold loss L_c and electron noise temperature, T_d:

$$\frac{T_e - 1}{T_d - 1} = 1 - \frac{L_c}{L_t}$$

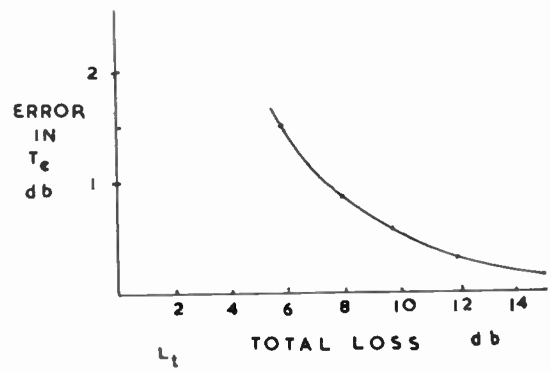


Fig. 10.—Errors in calculation of effective noise temperature.

If L_t tends to zero then the effective noise temperature tends to the value T_d. Fig. 10 shows how important it is to make L_t large in order to keep the error in T_e small.

MEASUREMENTS ON PARTICLE ACCELERATORS

C. W. Miller discussed the problem of plotting the standing wave in microwave guides used in particle accelerators. Here a normal E_0 waveguide has irises at intervals along the guide which slow down the velocity of propagation below c. It might be required to plot the pattern with a short circuit at the end of the guide and there might possibly be only four irises or corrugations which would only provide four points. The technique adopted is, therefore, to move the short circuit and hence the standing wave pattern. If the wavelength in the guide is not integrally related to the iris spacing, a large number of points at different frequencies is plotted and it is found that with a large number of corrugations the basic pattern repeats itself.

Measurement of the precise values of the electric and magnetic fields within corrugated waveguides is often necessary and the perturbation technique affords an important method. A small metallic bead is suspended in the guide and moved along, the fractional change in resonant frequency being representative of the energy stored in the bead and the total in the whole cavity. The fields can then be calculated. The method is also applicable to cavities used for accelerating protons which are loaded with drift tubes.

B. E. Kingdon: If I may first elaborate Mr. Miller's remarks: In basic principle, the fields inside the cavity are locally modified or distorted by the introduction of a metal or dielectric bead, having dimensions very much smaller than those of the cavity. The changes in field thus produced give rise to a change in resonant frequency of the cavity, which can be measured, according to the waveband involved, by heterodyne methods or by observing on an oscilloscope the shift of the resonance-curve relative to a frequency-graduated time-base. The value of the fractional change in angular frequency ($\delta\omega$ / ω_0) can then be used to calculate the value of field strength present at the point of perturbation, if the bead is calibrated in a "standard" cavity of known characteristics.

A spherical metal bead will perturb both the electric and magnetic field components present, and will give $\delta\omega/\omega_0$ proportional to $a^3(\frac{1}{2}\mu H^2 - \varepsilon E^2)$,* whereas a similar dielectric bead will affect only the electric field, so that the term in H will be absent from the expression. Thus the effect of the magnetic component can be eliminated, and by substitution both the electric and magnetic components present at the point of perturbation can be evaluated.

For perturbation by a spherical bead in an electric field, the value of $\delta\omega$ / ω_0 depends directly on the factor $(K-1)/(K+2)$, where K is the relative permittivity of the perturbing body material, and it is convenient to use a dielectric such as distrene, which is fairly easily machined and holds its nominal value of K (about 2·45–2·65) over all frequencies up to 25,000 Mc/s. In regions where only the electric component is present, however, or the magnetic effect is negligible so that no possibility of ambiguity exists, one can obtain an enhanced value of $\delta\omega/\omega_0$ by using a metal bead ($K \rightarrow \infty$).

The direction of the field may be obtained by the use of perturbing beads in the form of

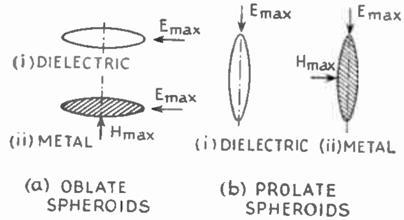


Fig. 11.—Illustrating various forms of perturbing body.

prolate and oblate spheroids (Fig. 11). A maximum perturbation of an electric field will be obtained when the direction of field is along the major axis of a metal or dielectric spheroid of either shape, whereas the magnetic perturbation will be greatest when the direction is parallel to the minor axis of a metal spheroid. Where the direction of field is known, however, or the magnitude is alone required, the use of a spherical bead greatly simplifies matters.

The determination of the cavity Q-factor is intimately related to all this, both because of the techniques involved, and also since its value is used, together with that of the fractional frequency change, in evaluating certain other parameters in accelerator design. One can measure Q in various ways, such as with a frequency-modulated oscillator, or, indirectly, by comparison of the cavity decay-time against that of a known CR circuit, as Mr. Rogal has suggested.

We may take as illustration one of the types of cavity used in heavy particle acceleration, to which Mr. Miller referred (Fig. 12). This form of re-entrant structure produces an electric field which is principally confined to the longitudinal axis XX′ of the resonator, while the azimuthal magnetic field is zero on the axis and rises to a maximum at a radius approaching that of the outer wall.

* Where μ and ε are respectively the permeability and permittivity of the medium enclosed by the cavity, H and E are the magnetic and electric field strength respectively, and a is the radius of the bead.

Perturbation by a metal or dielectric bead along the axis XX′ enables the electric field strength responsible for the acceleration to be determined for a given stored energy in the cavity, and additional knowledge of the Q-factor allows one to evaluate the shunt impedance and acceleration efficiency of the structure. The series impedance and attenuation may also be found from dispersion and Q measurements made on a chain of these cavities, coupled together as in an actual accelerator design.

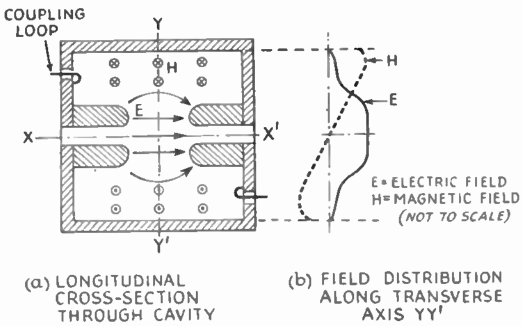


Fig. 12.—Mode of operation of typical proton accelerator cavity.

One great advantage in accelerator work is that all the measurements may be carried out on small scale models, so that an accelerating structure of, say, 3 ft or 4 ft diameter at 200 Mc/s can be reduced to something of the order 2½ in. diameter at 3,000 Mc/s, and the results suitably assessed. Where greater accuracy is desired, a somewhat larger model at, say, 1,000 Mc/s could be made. This method helps in saving a great deal of time and money, and the perturbation technique is really charming to use.

DIELECTRIC MEASUREMENTS

Dr. J. Brown: A field of measurements which is of considerable practical importance is that of the electrical constants of dielectrics, and now also of the ferrite class of magnetic materials. Methods involving samples of the material placed in waveguides or cavity resonators give consistent values for the dielectric constant and loss tangent of dielectrics and are being developed to give the corresponding magnetic constants of the ferrites.

In some recent work at Imperial College* an alternative method for measuring dielectric constant has been examined. A prism of the material is placed in a parallel plate spectrometer as indicated in Fig. 13 and the deviation, D, of the beam caused by travelling through the prism is measured. The dielectric constant can then be calculated from Snell's Law. A series of measurements carried out in this way gave the dielectric constant for polystyrene some four per cent. lower than the value obtained in a waveguide or a resonator. After a long investigation of the possible causes of the discrepancy, it was found that good agreement between the prism measurement and the others could be obtained if the prism surfaces were coated with tin foil before being placed between the parallel plates. No explanation as to why this should make an appreciable difference has yet been found. There remains however the possibility that similar discrepancies may arise in cavity measurements, particularly when relatively large areas of dielectric are in contact with conduction surfaces, and it would be interesting to hear if any evidence of this has been found.

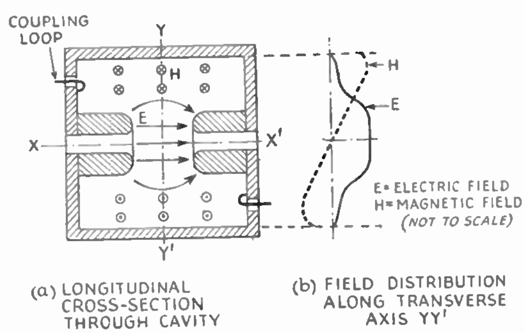


Fig. 13.—Measurement of dielectric constant using parallel plate spectrometer.

* P. H. Sollom and J. Brown, "A centimetre-wave parallel plate spectrometer," *Proc. Instn Elect. Engrs*, **103**, Part B, May 1956.

MECHANICAL CONSTRUCTION

B. E. Kingdon: At X-band wavelengths it is convenient to make slotted-section standing-wave indicators by machining them from the solid in two halves, joined along the centre-lines of the broad faces of the guide. Contact is thus made at a point of zero transverse wall current. The inside rectangle in each section can be milled and ground very accurately, and the slot cut after the parts are assembled. Although this method of construction would be wasteful at S-band, it would provide a solution to a number of problems.

We have had standing-wave indicators fabricated from brass plate, and silver-plated after being accurately assembled. Unless the material is thoroughly weathered and annealed, within six months these components are quite useless, and yet they might cost £100 or more each.

Silver-plating can also be a source of trouble where moving parts are concerned. It does tend to rub unevenly, and one gets erroneous results from standing-wave tests made on silver-plated surfaces with a simple probe and carriage arrangement. Naturally, this also applies to varnishing.

I. A. Harris: One frequently finds choke couplings for the waveguides on laboratory measuring instruments. The precise characteristics of such couplings are not well known. It is evident that, among other things, there is appreciable loss on account of the high frequency resistance in the half wavelength short-circuited line formed by the choke joint, particularly at frequencies as high as 10,000 Mc/s. Therefore it is becoming increasingly popular to grind the flush coupling faces accurately flat (and square with the waveguide axis) and to clamp pairs of these tightly to form the junctions, dispensing with the choke couplings in the important parts of a measuring set up.

L. C. Walters pointed out the tendency to obtain unwanted resonances with choke couplings.

E. M. Wareham: The couplings are certainly the vital links when you are trying to do measurements in waveguide, but the main difficulty is not to get the faces absolutely flat, nor is it necessary to pull them together very hard. It is just necessary to make them touch, and it is surprising how very rarely this happens with the kind of waveguide couplings we have. For example, the type fixed with screwed rings pull together with the backs of the coupling against the shoulders of the rings. There is a considerable risk, therefore, that the faces may not be parallel, and may only touch at one point, leaving an appreciable gap opposite.

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A VERY LOW FREQUENCY RECEIVER WITH HIGH SELECTIVITY *

by

C. S. Fowler (Associate Member) †

SUMMARY

The paper describes the design of a super-heterodyne receiver having very high selectivity, sensitivity and gain stability for use in the frequency band 6 to 36 kc/s. The high selectivity is obtained by means of controlled increase of Q-factor of the tuned circuits of the i.f. amplifiers, using mixed positive and negative feedback.

1. Introduction

The receiver was designed for use in an investigation of wave propagation at very low frequencies. It covers the frequency range 6 kc/s to 36 kc/s and its main features are very high selectivity, gain stability and sensitivity.

The method employed to obtain the required selectivity characteristics depends on the controlled increase of the Q-factor of a tuned circuit by mixed positive and negative feedback. The particular manner in which this is achieved in this receiver is thought to be new.

2. General Description

The general layout of the receiver is conventional (Fig. 1) and consists of one r.f. stage, mixer, local oscillator, two i.f. (40 kc/s) stages, infinite impedance detector and audio frequency amplifier. A stable beat frequency oscillator is provided for the reception of c.w. signals, and unwanted signals can be reduced in amplitude by means of a narrow band rejector circuit coupled to the first intermediate amplifier stage. The aerial and r.f. circuit tuning is ganged, but the aerial capacitor can be rotated over a mechanically limited arc independently of the r.f. capacitor. It is not practicable to gang the local oscillator tuning also, owing to the much smaller ratio of maximum to minimum frequencies (46–76 kc/s) required from the oscillator.

To conserve space and reduce stray r.f. fields to a minimum all coils are wound on pot-type iron dust cores, and adjacent coils are mounted so that their axes are mutually at right angles (all stages are separately screened). To reduce hum pick-up the heater winding is centre tapped to earth and the power supply unit is external to the receiver.

3. R.F. Stage

The frequency band is covered in two ranges, 6–18 and 16–36 kc/s, using a variable tuning capacitor of 1,100 pF maximum capacitance.

To keep the gain more uniform over the band the anode load of the variable-mu r.f. valve is made resistive and loosely coupled to the next tuned circuit.

When using the normal open aerial input the r.f. bandwidth varies between 500 and 1,500 c/s over each frequency range. With the direct input, which is provided for use with a tuned loop aerial, the effective bandwidth and r.f. sensitivity depend on the Q-factor of the loop.

4. Local Oscillator and Mixer Stages

In view of the high gain stability required and the narrow i.f. bandwidths used, it is necessary to ensure that the frequency drift of the local oscillator is small. With a 50 c/s bandwidth this must not exceed 1 part in 10^4 for the gain variation to be less than $\frac{1}{2}$ db.

A Franklin-type oscillator circuit was chosen, and modified to provide a more constant output over the frequency range. The modification consists of an inductance in the cathode of the phase inverting valve. The resultant increase of negative feedback with frequency counteracts the normal increase of gain obtained with this circuit. The measured variations of output are less than $\frac{1}{2}$ db over the working range of the oscillator, i.e. from 46 to 76 kc/s.

* Manuscript first received 26th January 1956 and in final form on 15th May 1956. (Paper No. 360.)

† Official communication from D.S.I.R., Radio Research Station, Slough.

U.D.C. No. 621.396.62.029.45:621.375.234.

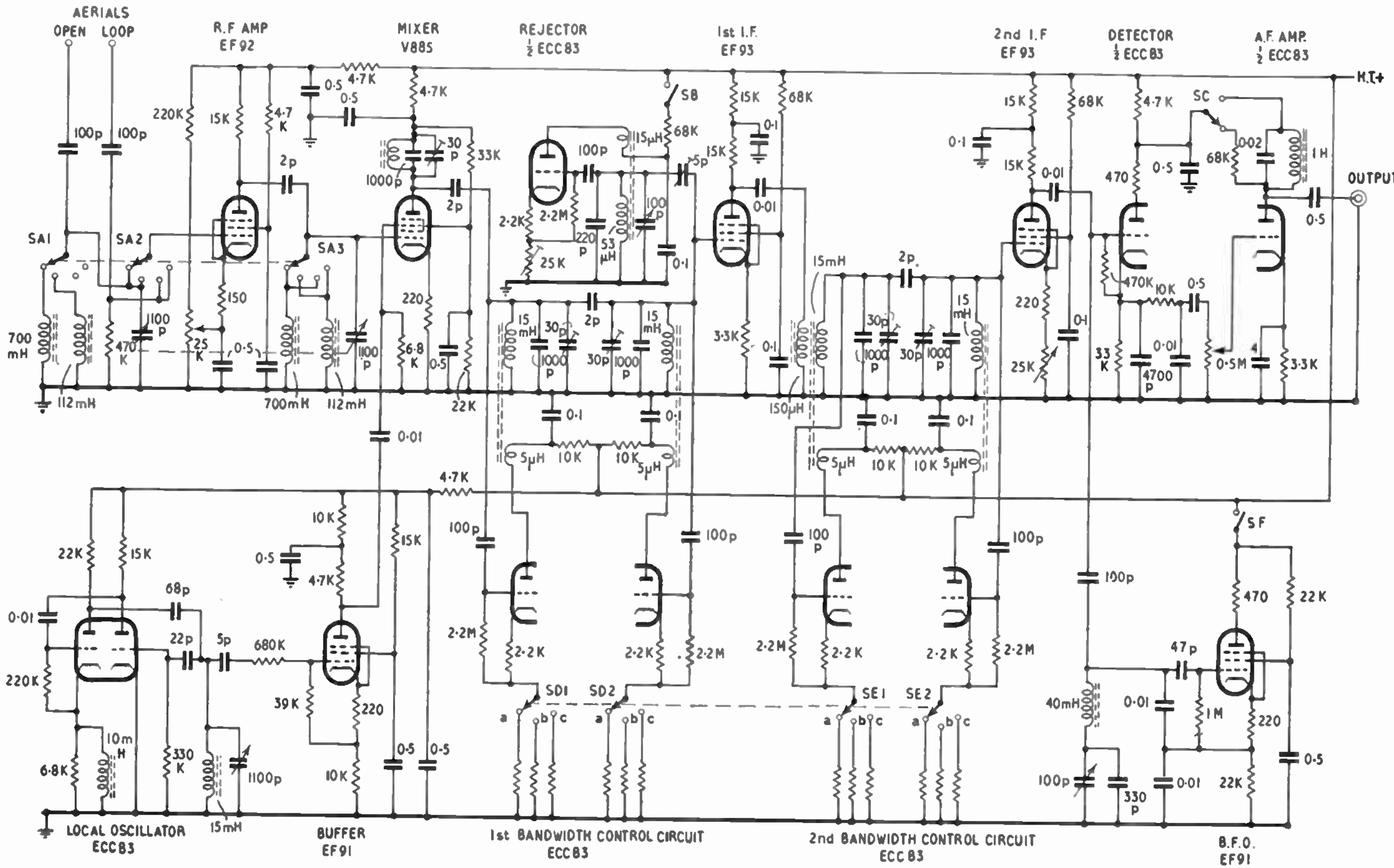

Fig. 1.—Very low frequency receiver.

5. I.F. Circuits

As shown in Fig 1 the two i.f. stages contain five top-coupled circuits tuned to 40 kc/s. The Q factor of the first circuit is lower than those of the remaining four so that its effect on the overall selectivity characteristic can be neglected. The required alternative (3 db) bandwidths of 50, 80, and 150 c/s are obtained by changing the operating conditions of the four circuits as described later. Since the 3 db bandwidth reduction factor for four similar loosely coupled circuit is 0·44, the individual bandwidths of these circuits are approximately 110, 180 and 340 cycles per second and the Q factors 350, 220 and 116 respectively.

In order to obtain the high Q-factors necessary for these narrow bandwidths an auxiliary triode is connected to each tuned circuit. The positive feedback, which would normally be just sufficient to keep the circuit in oscillation, is controlled by introducing negative feedback by means of an unbypassed cathode resistor. The higher the value of this resistor the lower the effective Q-factor of the tuned circuit. In this application the cathode resistors are switched to give the required bandwidth, but for other applications calibrated ganged variable resistors could be used.

The design of the basic i.f. circuit is conventional except that in the narrowest bandwidth condition, the coupling is adjusted to 0·707 of the critical value, as the most linear phase characteristic and the greatest selectivity are obtained under this condition. The coupling is lower than this in the wider bandwidth conditions and results in some loss of gain. However, as this loss falls within the range of the i.f. gain control it was not considered necessary to compensate for it by switching the top coupling capacitors. Again no particular care need be taken to wind high Q coils. However, where more than one circuit is used the Q factors should be matched, so that the auxiliary valves used in the feedback circuits are working under similar conditions; and for the same reason double triodes should be used in preference to individual valves.

Stable bandwidths of between 2 and 5 c/s have been obtained with one tuned circuit by this method, but instability troubles arise when two or more such circuits are combined, unless extreme care is taken with decoupling and screening.

6. Rejector Circuit

This is a single tuned circuit in which the Q-factor can be raised to the order of 2,000 using the positive and negative feedback circuit employed in the i.f. stages. It is coupled to the grid of the first i.f. valve through a 5 pF capacitor.

In operation this circuit is tuned to a few cycles below the frequency it is required to reject. In this condition it is effectively a high inductance at the rejection frequency. This inductance resonates with the coupling capacitance to form an acceptor circuit with a very low series resistance which is across the input to the i.f. valve. A rejection of 60 db is obtainable over a bandwidth of 20 c/s for any frequency within ± 3 kc/s of the intermediate frequency. Rejection over a wider bandwidth with a proportionally lower rejection ratio can be obtained by operation of the pre-set control in the cathode circuit.

7. Detector, Audio Frequency and B.F.O. Stages

The infinite impedance detector is followed by an i.f. filter, a.f. gain control and a.f. amplifier.

Alternate anode loads of either a resistance or a filter circuit tuned to 1 kc/s are provided for the audio output stage. The filter gives further discrimination against unwanted c.w. signals.

The b.f.o. circuit employs a Clapp-type oscillator,* the long-term stability of which is better than 1 part in 10^4 under typical ambient temperature variations.

8. Performance

The performance of the receiver when using the open aerial input is summarized below.

Selectivity.—The measured overall response curves at a frequency of 19 kc/s for the three bandwidths obtainable are shown in Fig. 2. The r.f. bandwidths (at -3 db) vary from 500 to 1,500 c/s on each range according to the frequency used.

Sensitivity.—The sensitivity of the receiver with various settings of the gain controls is shown in Fig. 3.

* J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability." *Proc. Inst. Radio Engrs*, **36**, p. 356, 1948.

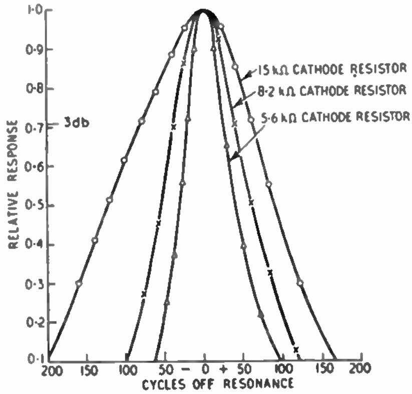

Fig. 2.—Overall selectivity curves.

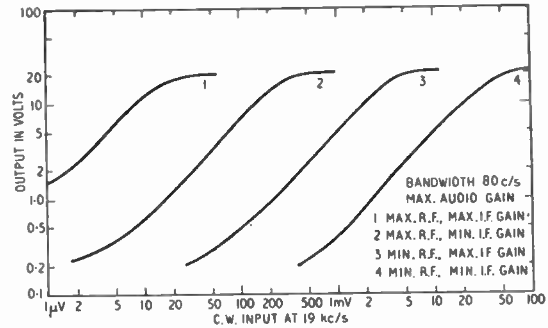

Fig. 3.—Amplification characteristics.

Signal/noise ratio.—A signal of between 2 and 4 μV is required at the input to give a (signal + noise)/noise ratio of 6 db at the output.

Image frequency suppression.—Greater than 80 db.

Gain stability.—Precise measurements made over a period of 8 hours under normal operating conditions, showed gain variations of less than $\frac{1}{2}$ db, and from recordings of field strength over periods of up to 72 hours, it was concluded that this stability is maintained for longer periods.

Linearity.—The degree of linearity between output and input is given in Fig. 3 for various settings of the gain controls.

Power consumption.—H.T.: 50 mA at 250 volts (stabilized); L.T.: 3 A at 6·3 volts.

9. Acknowledgments

The author wishes to acknowledge the suggestions of Mr. G. E. Ashwell in connection with the use of stabilized positive feedback, and the assistance of Mr. C. Medhurst in the experimental work.

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VERTICAL RADIATION AND TROPICAL BROADCASTING *

by

A. H. Dickinson (Associate Member) †

SUMMARY

National stations in the Tropics should use short-wave vertical incidence aerials consisting of 16-element binomial arrays; with 5-kW transmitters these would be capable of serving 90,000 square miles of territory. Co-channel stations should be separated by 1,500 miles. The stations require 2 or 3 frequencies in the 2·5, 3·5, 5 and 9 Mc/s bands depending on location; so far as possible the 3·5 and 5 Mc/s bands only should be used. Use of the minimum number of frequencies and frequency changes assists listeners. Stations intended to serve small, densely populated areas can operate in the medium-wave band.

1. Introduction

A general paper on High Frequency Broadcast transmission with vertical radiation was presented at the 1951 Convention.[1] The present paper puts forward the author's opinions on aerials, transmitter powers and frequencies, based on five years' experience operating a vertical incidence system in Jamaica.

Most tropical countries are more or less undeveloped so that any broadcasting system planned to serve these areas must be planned within a limited budget. In general, the population is distributed fairly evenly over large areas of country except for a few large towns.

The most economical system for these conditions is one which provides first class service in the main centres of population and second class service over the rest of the territory. The first class service areas can best be served by small medium wave transmitters and the second class areas are most economically served by means of a short-wave vertical incidence system.

2. Signal Strength

The signal strength required, in the absence of man-made static and jamming due to frequency sharing, depends entirely on atmospheric and receiver noise. In the tropics, and on the frequencies usually used for vertical incidence systems, atmospheric noise is generally the controlling factor, especially if the medium-wave band is used to cover the large towns where man-made noise might be important.

A very useful map dividing the world into noise grade areas has been published by the C.R.P.L.[2] and the author has found in practice that the noise levels given by this map are usually somewhat high. Although a noise measurement programme has not yet been undertaken, the opinion expressed is based on regular listening to a vertical incidence system together with signal strength measurements.

The noise level map is, however, a very useful guide and, any error being on the right side, has been used for all calculations of required power. Those requirements published in this paper are based on noise grade 4 and will give a safe margin of power in hand for the majority of British Colonies and Dominions.

3. Range and Frequency Sharing

All the short-wave bands are exceedingly overcrowded and any alleviation which can be obtained is to the advantage of everyone concerned. Many broadcasting authorities built short-wave stations at a time when insufficient information was available on noise levels and radiation efficiencies, so that the tropical broadcasting bands have become particularly overcrowded.

There appears to be a strong but entirely erroneous opinion held by some engineers in the tropics that broadcasting on the medium-wave bands is impractical due to high noise levels, and that a better service can be obtained for local broadcasting by using short-waves.

* Manuscript received June 8th, 1956 (Paper No. 361).

† Central Rediffusion Services Ltd., London, S.W.1.

U.D.C. No. 621.396.677.42.

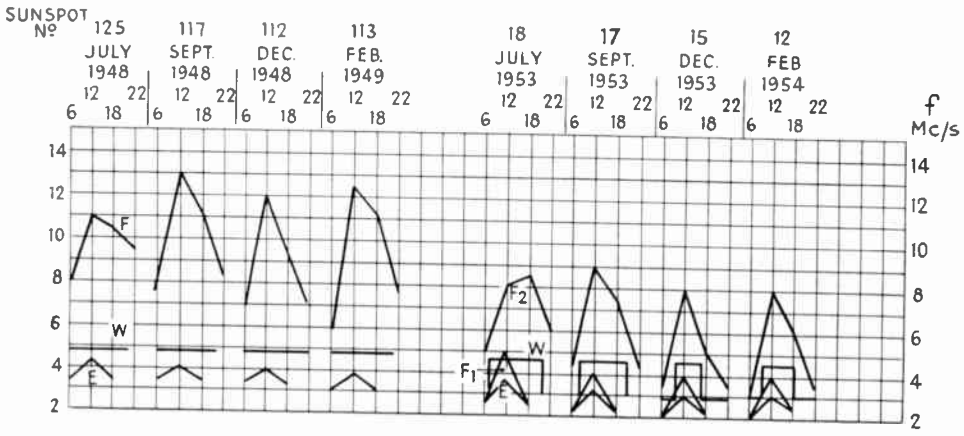

Fig. 1.—Critical frequencies and working frequencies for Jamaica.

The author has built and operated several stations in the West Indies in the medium-wave band and obtained good coverage. A signal strength of 0·5 mV/m gives a very satisfactory signal-to-noise ratio on a frequency of 700 kc/s except during local thunderstorms.

Many local service stations are being operated in the tropical broadcasting bands 2, 3 and 5 Mc/s, with vertical aerials and single dipoles, which would be better replaced by medium-wave transmitters in the range 600-1,600 kc/s. A superior local service would be obtained and the interference situation greatly alleviated.

Broadcasting authorities might well limit the use of the short-wave bands to National Stations intended to cover the largest amount of territory and limit the duplication of such stations within the national boundaries.

These National Stations could each cover a range of about 300 miles (90,000 square miles) with a power of 5 kilowatts, and stations 1,500 miles apart could operate on the same frequency if the correct types of aerial were used. Each country would thus enjoy interference-free broadcasting by using only a few frequencies in the existing tropical broadcasting bands. The British Colonies and Dominions in the Tropics might be completely covered using only about eight frequencies in each of the chosen bands.

4. Frequencies

The ideal situation is, of course, one where it is possible to use the optimum working frequency (O.W.F.) at all times and thus work

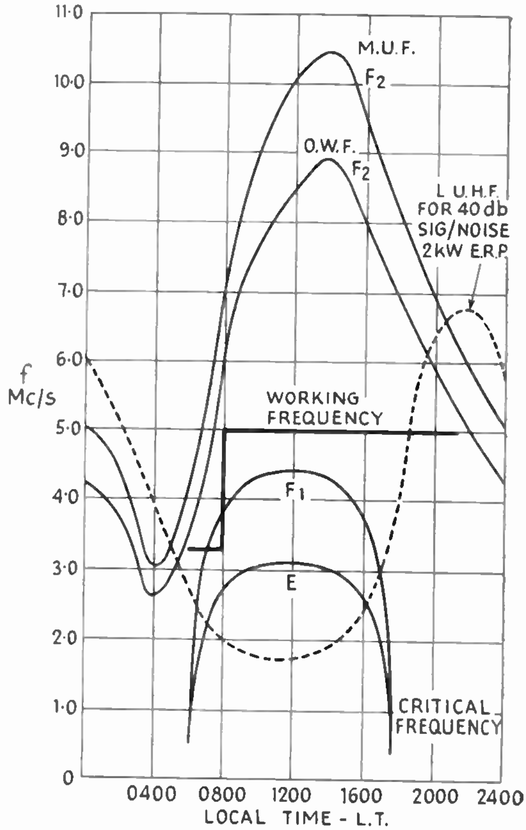

Fig. 2.—Working frequencies for West Indies, October 1954. Sunspot number 10.

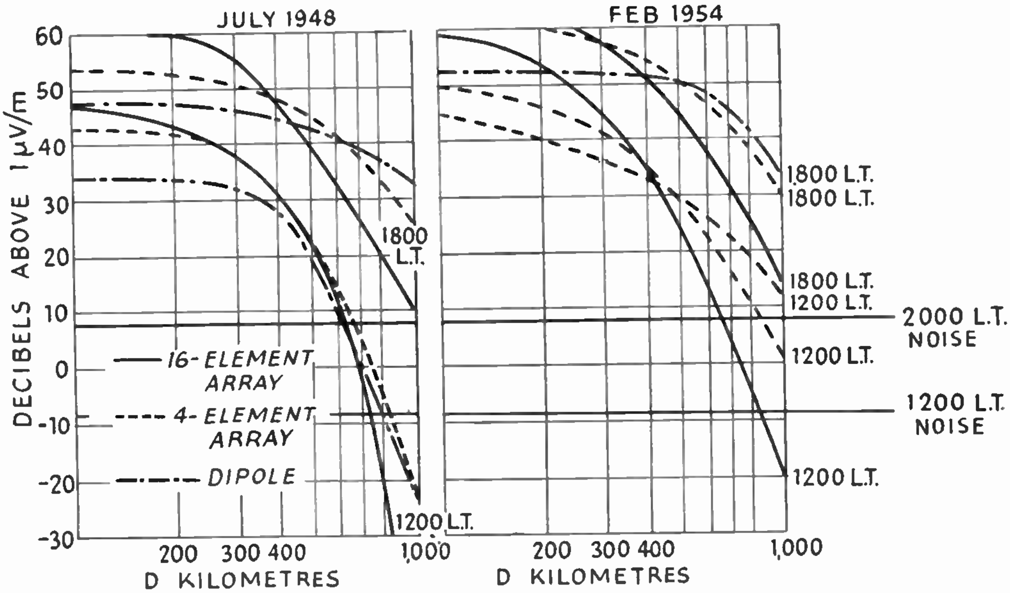

Fig. 3.—Comparative curves showing field strength against distance for various aerials using vertical incidence. Active transmission modes up to 2-hop F2. 5 kW power.

with the minimum possible power. However, there are several reasons why it is not possible to do this.

A vertical incidence broadcasting system is intended to serve listeners situated at anything from a few miles to 300 miles from the transmitter; when E and F1 transmissions are used there is a 2 : 1 difference in the maximum usable frequencies at 0 and 300 miles. F2 transmission only calls for a five or ten per cent. frequency difference for the two distances.

The variation in critical frequencies during the usual broadcasting hours over the sunspot cycle is shown in Fig. 1, the range being from 3·5 Mc/s to 13 Mc/s and is at times as great as from 5 to 12 Mc/s over a single day.

A broadcasting service is catering for receiver operators who are not technically-minded and therefore do not appreciate the reasons for frequency changes and in fact will not listen to the station if they have to be continually re-tuning their receivers. It has been found in practice that the minimum number of frequencies should be used in order to assist the listener to memorize them and the minimum number of daily changes should be made. In Jamaica two frequencies are sufficient over the major portion of the sunspot cycle with two changes during the most difficult day providing satisfactory reception. During a large part of the sunspot cycle it was possible to operate all day on only one frequency. Fig. 1 shows the frequency schedule used.

It is desirable in view of channel congestion to keep short-wave broadcasting in the lower frequency ranges. In Jamaica it was found that during sunspot maximum there was too much interference on the 9 Mc/s band, and although it appeared to be reasonably near the O.W.F. at that time, a much better service was provided by using the 5 Mc/s band.

During sunspot minimum the 3·5 Mc/s and 5 Mc/s bands together gave satisfactory signals except for a very short period in the winter of 1954 during late evenings and early mornings. At this time 2·5 Mc/s was necessary, due to 3·5 Mc/s being above the critical frequency of the F2 layer and therefore skipping a large portion of the service area.

5. Short Term Frequency Changes

The National Bureau of Standards in Washington, U.S.A., and the National Physical Laboratory in Britain produce monthly frequency predictions in advance. These provide sunspot number and critical frequencies

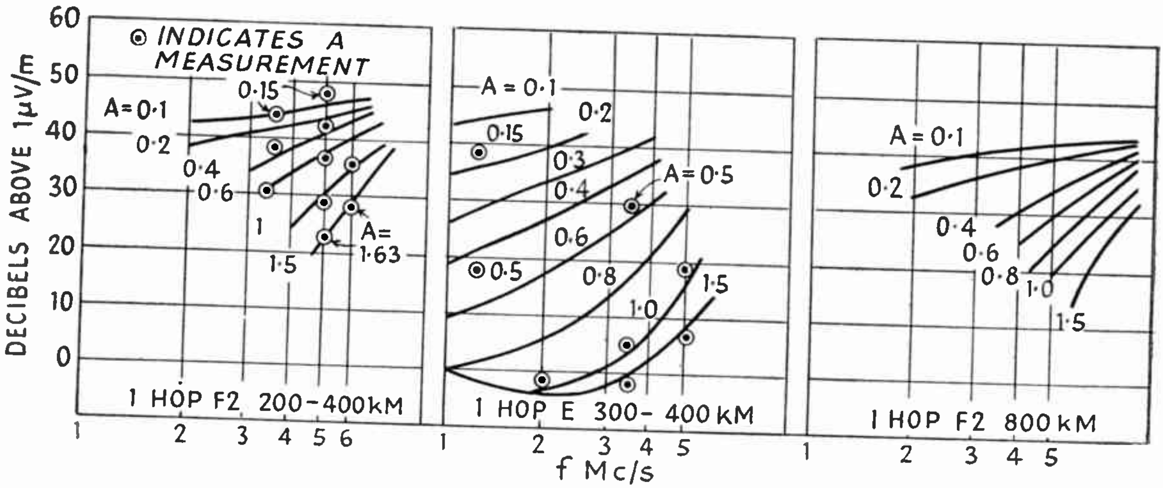

Fig. 4.—Field strength against frequency for various distances, absorption coefficients and transmission modes.

for the F2 and E layers, so that it is then possible to produce from them a set of curves showing M.U.F. and L.U.H.F. for all times of the day (Fig. 2). If this is done once a month, the correct working frequency is always known. Ionospheric predictions are now also issued by the National Physical Laboratory of India.

The M.U.F. for a vertical incidence service should be 10 per cent. below the critical frequency and the L.U.H.F. can be found from the nomogram of Fig. 5 which is based on the C.R.P.L. noise grade 4 and on the field strengths values of Figs. 3 and 4.

The O.W.F. should be as low as possible without going below the L.U.H.F. or near the E and F1 critical frequencies. The reason for using the lowest possible frequencies is that selective fading, flutter and interference from other stations is minimized when working as far below M.U.F. as possible.[3]

6. Power Requirements

The required power for a good broadcasting service is that power which with the correct aerial system will give at least a 35 decibels signal/noise ratio during most of the broadcasting time, with the noise figure based on atmospheric noise.

At the present time it is impossible during the evenings of sunspot minimum to overcome the noise due to jamming by other stations without using very large powers. Even if this is done, a power race between the various authorities achieves nothing except a very temporary advantage, and, as explained above, is unnecessary if proper frequency allocations are made and authorities use the correct frequencies, power and antennae.

The power requirement varies with the diurnal, seasonal and sunspot cycle, but in order to set up a broadcasting system it is necessary to provide the maximum power required. At certain other times the transmitter may be worked at quarter power to save running expense, although it is doubtful whether the saving in power cost is worth the labour involved.

In order to find the maximum power required, it is necessary to take into account the atmospheric noise at various times of the day and the field strength per $\sqrt{(kW)}$ at the limits of the service area and at the working frequency.

Table 1

Lowest Required Radiated Powers for 40 db signal-to-noise ratio at 300 miles

February 1954				July 1948		
f Mc/s	Absorption Coefficient	L.R.R.P. kW	Local Time	*f* Mc/s	Absorption Coefficient	L.R.R.P. kW
3·5	0·6	0·15	0800	5	1	0·06
5	1·2	1·2	1200	5	1·6	4
3·5	0·15	5	1800	5	0·5	3½
Sunspot Number 12				Sunspot Number 125		

The receiver noise can be neglected as it is generally well below atmospheric noise.

The nomograph of Fig. 5 has been prepared for finding the power requirements, based on grade 4 noise and the field strengths of Fig. 4. The grade 4 noise levels are as indicated by the curves produced by C.R.P.L. and Fig. 4 gives signal strength for various distances, absorption coefficients and modes of propagation. These curves were produced by the National Bureau of Standards from actual measurements made in the U.S.A., and checks made in Jamaica are shown in circles. These checks agree fairly well with the curves, although for the short distances, lower field strengths are shown than are indicated by the usual theoretical calculations.

Table 1, showing the lowest radiated power requirements, has been prepared from the nomograph for sunspot maximum and minimum. It will be seen that the maximum power necessary if frequencies are limited to 3·5 and 5 Mc/s is only 5 kW but the use of higher frequencies than 5 Mc/s would reduce the required power if a clear channel were obtained.

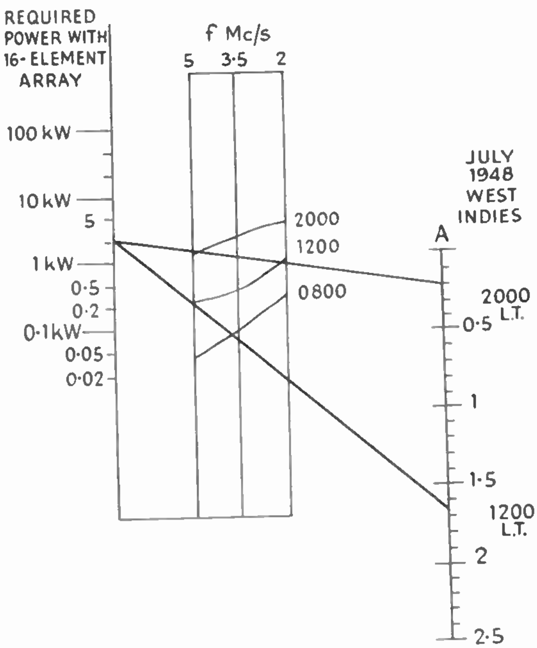

Fig. 5.—Nomograph for calculating L.U.H.F. for given power or L.R.R.P. for given frequency at various values of absorption. Based on Noise Grade 4. Range 0 to 300 miles, using 16-element array.

7. Aerials

Aerial requirements are fixed by the service area required, the distance between stations sharing frequency and the position of the transmitter in relation to its service area. In

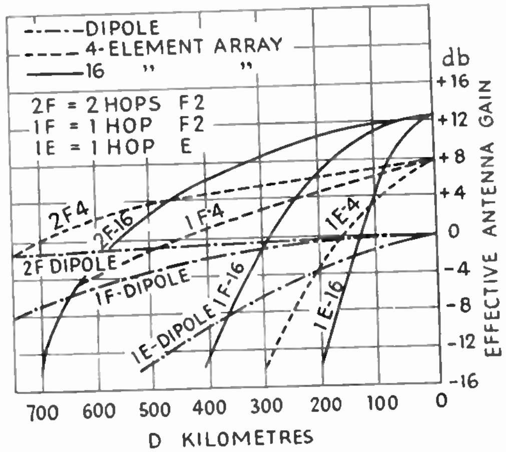

Fig. 6.—Effective radiated power gain of arrays for various modes of propagation and receiving distances.

order to obtain maximum signal strength it is necessary to concentrate as much power as possible in the vertical direction and the required lobe width will depend on the maximum distance to the outside boundary of the service area.

It has been found in practice that with a service area of 300 miles radius, it is usually possible to work with "two-hop" F2 transmission, which enables one to use a narrower lobe than would be possible using "one-hop" transmission, with the consequent advantage of increased signal strength in most of the service area and reduced signal outside the service area facilitating frequency sharing.

Simple aerials should not be used for they very often produce interference thousands of miles away, and frequency sharing becomes impossible. At one time reception reports were received in Jamaica from a listener 4,000 miles from the transmitter which at that time was using an aerial having a gain of 5 db in the vertical direction, and rather a high value of low angle radiation, due to the element height above earth being only 0·17 wavelength.

The use of low gain or wide lobe aerials will not give any consistent increase in service area over about 300 miles, especially during periods of medium or high absorption.

To illustrate the effect of the radiator on service area the set of curves shown in Fig. 3 have been prepared.

The three aerials considered are:—

(1) A halfwave horizontal dipole $\frac{1}{4}\lambda$ above earth.

(2) Four horizontal halfwave dipoles $\frac{1}{4}\lambda$ above earth, placed in two parallel sets of two, spaced half λ between centres and driven all in phase with equal currents.

(3) Sixteen elements as above but fed with currents for a binomial array and placed in four parallel sets of four.

All these aerials are described in a paper by Adorian and the present author.[1]

These curves give field strength against distance at sunspot maximum and minimum during high and low absorption times of the day. They have been calculated from the curves of Fig. 4 and Fig. 6. Fig. 4 is based on actual field strength measurements and Fig. 6 on the calculated polar diagrams of the aerials[4], on the basis of 110 km effective height for the E layer and 350 km effective height for the F2 layer and for a spherical earth. The method of calculation using this type of nomograph is explained in a National Bureau of Standards circular.[5]

A 16-element array fed with the same current in all elements has minor lobes in its radiation pattern at low angles so that it will tend to give interference at great distances. The author therefore recommends the use of a binomial array when 16 elements are used because such an array can be arranged to get rid of these minor lobes.

Examination of Fig. 3 shows that during the daytime at sunspot maximum, the practical service area is about 300 miles radius and that there is no advantage in using transmitters of more than 5 kW except perhaps in Noise Grade 5 areas. Furthermore there is nothing to be gained and much to be lost by using simple aerials.

Radiation of 5 kW by any practical vertical incidence aerial provides a minimum signal-to-noise ratio at 450 miles (700 km) of only 8 db and in order to obtain 36 db a power of 2 MW would be necessary. Even if the operating frequency were increased from 5 to 9 Mc/s the power required for a 450-mile service area would still be 150 kilowatts for 36 db. Furthermore, at this distance and at that time the advantage of using a low gain wide lobe aerial is only 6 db.

Between 300 miles and zero distance the advantage of the high gain aerial becomes progressively greater and at the former distance the signal-to-noise ratio is 30 db for the high gain aerials and only 28 db for the dipole, while at 100 miles the 16-element array is 10 db better than the dipole.

The reduction of interference caused to co-channel stations at 600 miles from the interfering transmitter is at least 18 db using the 16-element binomial array under conditions of low absorption and 33 db under conditions of medium absorption.

The advantage of the 4-element array over the dipole varies between 8 db and 3 db inside the 300 mile limit. Outside the service area the reduction of interference using this aerial varies between 14 db and 0 db.

The 16 element array is much superior to the others, being from 16 to 33 db better for shared channel working.

Table 2

Comparison between Transmitter Installations

CAPITAL COSTS (INSTALLED)

	2 (5 kW) £ Sterling	1 (75 kW) £ Sterling
Transmitters	15,000	76,000
Aerials	3,000	300
Sites	400	50
Buildings	2,000	3,000
	20,400	79,350

MAINTENANCE COSTS PER ANNUM

	2 (5 kW) £ Sterling	1 (75 kW) £ Sterling
Valves	700	2,000
Components	50	150
Labour (engineers @ £600 per annum)	3,600	3,600
Power @ 2d./unit ...	2,000	6,000
	6,350	11,750

8. "Slewing" of Arrays

Where it is not possible to site the transmitter at the centre of the service area, the lobe of the high gain aerials can be "slewed," i.e. offset from the vertical so that the boundary of the service area is further away on one side of the aerial than on the other. However, the author does not recommend that this should be done by more than a few degrees otherwise the low angle radiation may be increased sufficiently to cause objectionable inteference to co-channel stations.

9. Cost Considerations

Good engineering practice requires that maximum result for the minimum capital and maintenance cost should always be achieved. Now, as explained in Sect. 7, it is possible to obtain 36 db signal-to-noise ratio during times of highest absorption at a distance of 450 miles thus getting twice the service area recommended above by the author. To achieve this result, however, it would be necessary to use 75 kW of power and a simple dipole aerial.

It is interesting to examine (Table 2) the capital and running costs of one 75-kW transmitter installed with a dipole aerial against the costs of two 5-kW transmitters with 16-element arrays, each method of course giving equal total coverage.

10. Conclusions

It is suggested that broadcasting stations in the Tropics serving small densely populated areas should change to the medium-wave band.

National stations should use short-wave vertical incidence 16-element binomial arrays; with 5-kW transmitters these would be capable of serving 90,000 square miles of territory, and frequency sharing is possible if co-channel stations are separated by 1,500 miles.

The stations require a total of 2 or 3 frequencies in the bands $2\frac{1}{2}$, $3\frac{1}{2}$, 5 and 9 megacycles depending on location, but so far as possible the $3\frac{1}{2}$ and 5 Mc/s bands only should be used. Use of the minimum number of frequency changes assists listeners.

11. Acknowledgments

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12. References

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. . . Radio Engineering Overseas

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

534.843/5

Notes on the acoustical properties of materials. T. VOGEL. *Onde Electrique*, **36**, pp. 428-34, May 1956.

The propagation of sound within an enclosure sets a "boundary problem" which involves the observation of certain "wall" conditions. One plausible aspect of wall condition, similar to that met in thermo-kinetics, eventually leads to defining the materials in terms of specific "impedance." This impedance may be likened to the coefficient of absorption measured for different angles of incidence in plane-wave conditions, and also to the periods of reverberation. Sabine's semi-empirical formula, currently applied in acoustical engineering, actually seems to be a limit case: it is possible to get close to that limit case, but only with conditions well removed from those prevailing in reverberation chambers. For that reason, the author prefers the incidence method described in the present article, whereas the use of Sabine's formula should be restricted to preliminary work on rooms of relatively high absorption coefficients.

621.317.77:621.372.5

A simple method of accurate phase measurement of a four-terminal network.—B. CHATTERJEE. *Journal of the Institution of Telecommunication Engineers*, **2**, pp. 93-5, March 1956.

A simple method of phase measurement is described, in which the frequency change produced in a tuned oscillator is used to indicate the phase change of a given four-terminal network. The given network is introduced in the closed loop of an oscillator circuit, the phase change of which shifts the frequency of oscillation of the oscillator. A simple circuit arrangement is described and the limitations of such a method are discussed.

621.385.1

Travelling wave tubes for 4 cm. waves, research and development at the Centre National D'Etudes des Telecommunications.—A. BOBENRIETH and O. CAHEN. *Onde Electrique*, **36**, pp. 307-17, April 1956.

After recalling the P.T.T. specification for amplifier tubes for 4 cm radio links, the authors review the problems involved in producing a travelling wave tube to meet these requirements: the generation and convergence of the electronic beam, study of the helical delay line, the coupling of input and output waveguides to the amplifier. They then describe the methods adopted in the production of the model S.45 travelling wave tube, the manufacture of the helix, the cavity and the gun and the vacuum pumping method. Some performance figures are then given for the S.45 tube and for some improved tubes produced during the development.

621.385.1:621.314.7

The analogy of vacuum tube and transistor. H. BENEKING. *Archiv der Elektrischen Ubertragung*, **10**, pp. 214-21, May 1956.

It is shown that valves and transistors can be treated by the same formal method. A second Barkhausen relationship is found, which is applicable to transistors and to valves at high frequencies. Measured data can be substituted in the equivalent circuits to obtain expressions governing practical circuits. Various possible connections and the associated equivalent circuits are indicated.

621.385.832:621.397.62

The television picture tube. H. ROTHE and E. GUNDERT. *Archiv der Elektrischen Ubertragung*, **10**, pp. 188-194, May 1956.

The paper reviews those properties of the electron gun and of the focusing and deflecting systems of picture tubes which relate to modulation and to the resolving power of the picture tube. A satisfactorily reproducible half-tone presentation of the television picture is ensured, since the exponent of the control characteristic of modern picture tubes is practically independent of the tube data. The magnitude of the deflecting field depends on the tube data (anode voltage, anode current, cathode current density, maximum deflecting angle), and the deflecting coil used. The aperture error of the focusing system can be practically neglected with magnetic focusing and, in the case of a wide deflecting angle, also with electrostatic focusing. Electrostatic single lenses as well as electrostatic acceleration lenses can be constructed with low focusing voltages. In the case of single lenses, the demands on mechanical accuracy for compliance with a predetermined range of focusing voltage decrease with the distance of the lens from the cathode. Since the deflecting errors increase only slightly with wide deflecting angles, the resolving power can be improved by tubes with a wide deflecting angle.

621.396.11

Ionospheric focusing and image zones.—B. D. KHURANA. *Journal of the Institution of Telecommunication Engineers*, **2**, pp. 96-102, March 1956.

The desired target region of reception in short-wave broadcasting is covered by firing the radio waves from a suitably designed transmitting antenna into the ionosphere, at predetermined angles. Analysis shows that in addition to the target region, some extra zones on the earth's surface also come to receive an appreciably enhanced signal strength. This so-called "focusing" results from the curvature of the reflecting layers. The "first order image zones" have been determined for the regional short-wave transmitters of All India Radio, and plotted on the great-circle map, as an illustration.