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The cover photograph shows examples of each type of 'front-line' television camera tube which has been used since the BBC Television Service started in 1936.
The types shown are:

1. Iconoscope (Standard Emitron).
2. Image Iconoscope (Pye Photicon).
3. Cathode-potential-stabilised Orthicon (CPS *Emitron*).
4. 3-in. Image Orthicon (EEV).
5. 4½-in. Image Orthicon (EEV).
6. Plumbicon (Philips).

Contents

The major contributions are preceded by individual lists of contents.

Editorial

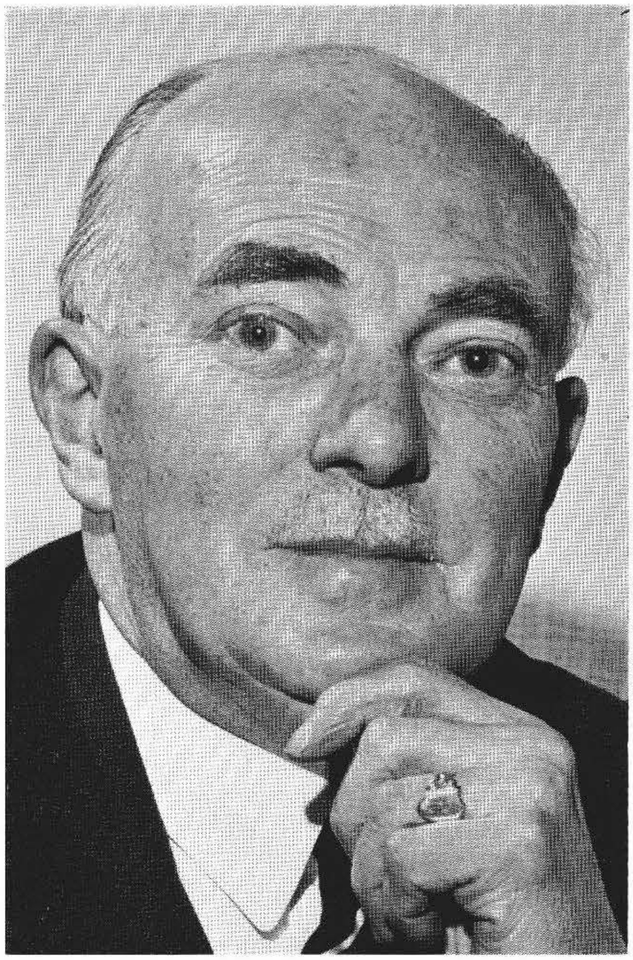
From Iconoscope to Plumbicon

Some of us may remember the days when the iconoscope picture tube (or the Standard Emitron, as the version used in this country was named) was the only means of producing live television pictures. Given enough lighting and much dedicated care in readjusting the shading ('tilt and bend') controls whenever the picture content changed, this tube could produce pictures not far short of the best which present-day cameras can offer – but few would now deny that it had serious snags. The same was true, in varying degrees, of other camera tubes introduced before and soon after the war, until the introduction of the image orthicon marked a new phase in television picture production. The shading controls in this tube were preset, it did not need keystone correction, and it was free from the exasperating 'peel-off'. The shape and dimensions were manageable, and in spite of the extremely thin target it was so robust that, when cameras were dropped, it was often other components and not the pick-up tube which were damaged. Above all, the image orthicon was much more sensitive than any of its predecessors. True, it had its snags, but they were snags which 'quality control' as well as the camera crews could learn to live with. Black haloes around highlights are nowhere near as objectionable as peeling-off and although the characteristic image orthicon 'edge' effect should strictly be considered a defect, it gives the impression of improved definition on many receivers. The $4\frac{1}{2}$-in. image orthicon, which was developed in Britain, was the first live picture source to give an overall quality assessment comparable with the best obtainable from a flying-spot telecine or slide scanner.

In Britain at any rate, the image orthicon has dominated television programme-making (apart from filmed programmes) for many years, but the end of this dominance may be in sight. All three networks are now transmitting most of their programmes in colour, using three-tube or four-tube plumbicon colour cameras. Unlike countries where colour services were started some years ago, we have no image orthicon colour cameras which have been in use since the pre-plumbicon days.

Many of us will view the passing of the image orthicon with the same feelings of nostalgic regret which railway enthusiasts feel for the passing of the steam locomotive.

Lord Jackson of Burnley



With the death on 17 February of Lord Jackson of Burnley, F.R.S., the BBC lost a valued counsellor, who was generous of his time and warm-hearted in his encouragement. As Chairman of the BBC Engineering Advisory Committee since 1965 he did a great deal to strengthen the links between the BBC and the Universities and Industry in the field of scientific research.

His interest in broadcasting extended into a wider field through his chairmanship, since 1962, of the Television Advisory Committee. He guided the Committee through the complex discussions that led it to recommend to the Postmaster General that BBC-1 and the ITA services should be duplicated on uhf on the 625-line standard and that colour should be introduced on 625 lines only. As BBC-2 was already being transmitted on 625 lines in the uhf bands, the adoption by the Government of the recommendations meant that, as the construction of uhf stations advanced, viewers would be able to receive all three programmes on a single-standard receiver with a single aerial.

Lord Jackson exemplified his own view that men trained in disciplines of science and engineering should bring their professional understanding to bear on the problems of the day in the public interest.

A graduate of Manchester University, Willis Jackson held, in the course of a very active life, a series of appointments in the academic world and in industry. After teaching at Oxford and elsewhere, he was appointed Professor of Electro-technology in the Victoria University of Manchester and later Professor of Electrical Engineering at Imperial College. He returned to industry for some years as Director of Research and Education in AEI (Manchester), and went back to Imperial College in 1961. He continued to hold the chair of Professor of Electrical Engineering and was appointed Pro Rector of the college in 1967.

In addition to his research work on the applications of solid state physics, he served on many important committees concerned with the training of engineers and technical teachers and with the deployment of man-power resources. He was President of the Institution of Electrical Engineers in 1959–60 and President of the British Association for the Advancement of Science in 1966–7. He was knighted in 1958 and created a Life Peer in 1967.

Throughout his career he took a warm interest in the welfare of members of his profession and gave a great deal of encouragement to young men entering it. He devoted himself unsparingly to all the tasks that he took up and carried them through with a modesty and humour that endeared him to a wide circle of friends.

New Teaching Techniques at the Engineering Training Centre

H. Henderson*

Contents

1 Introduction

The BBC Engineering Training Centre is situated two miles west of Evesham, Worcestershire. A large mansion, once the home of the Duc d'Orléans – who was a pretender to the throne of France – is used as headquarters and nearby, on the wooded slopes of Tunnel Hill, are monochrome and colour television studios, control rooms, recording suites, laboratories, and a restaurant block. Accommodation is provided for up to 250 trainees.

Students of all ages from BBC studios and transmitters throughout the UK attend the Training Centre for courses on a wide variety of subjects including colour television, stereo broadcasting, pulse code modulation, and line-store standards conversion in addition to basic training in electronic engineering. With these BBC staff are engineers from many parts of the world sent by their own broadcasting organisations to study in this country.

Of the 1500 students who pass through the Training Centre each year approximately 250 are young men taking their first course before joining an operational unit. Most of these come directly from sixth forms, not usually with A-levels, but with interests which fit them for a career in broadcast engineering. Some have by their own efforts acquired a good knowledge of electronics but most have no more than a fair grasp of O-level Physics and Mathematics.

One of the aims of this first course (the sixteen-week Technical Assistants' Course) is to provide the a.c. theory and electronics these young men need. In addition they are taught to work safely with high voltages and delicate apparatus, and their syllabus covers the complex equipment, systems, and procedures used in creating radio and television programmes. This article concentrates on the techniques used in teaching basic a.c. and electronic theory.

2 Programmed Learning

The most important of the techniques employed is that of Programmed Learning (PL). PL means many things to many people and arouses a surprising amount of emotion amongst teachers. This may be because it is associated with teaching machines and the idea of teaching by machine is anathema to many of them.

A learning program is essentially one enabling a student to teach himself from carefully-prepared material, to check his own progress and ideally to proceed at his own best rate. He is motivated by success at discovering step by step that he knows and understands what he has just learned.

The preparation of PL material is referred to as 'writing a program': it is a time-consuming operation requiring about fifty hours of preparation for each hour of student study time. This is why the BBC has chosen to prepare programs on basic theory: the 'life expectation' of such programs may be greater than for those on equipments and techniques which are ever-changing.

The techniques of program writing are dealt with in a number of textbooks and special courses are run for program writers. It is a skilled activity requiring a good understanding of the technique as well as of the material to be programmed and of the type of person who will study it (the 'target population'). When a program has been prepared it is subjected to a process of validation by trying it out on several groups of the target population to eliminate ambiguities, steps which are too large and, of course, errors.

2.1 PL Books

In a typical program, a piece of information is given and is followed by a question and a number of possible answers, only one of which is correct. The student reads the item of information and selects the answer he thinks to be right. Turning over the page (BBC programs are usually in book form rather than presented by machine) the student either finds he has chosen the right answer, is given a verbal pat on the back and taken on to new work, or discovers that he has selected the wrong answer. For incorrect answers the programme suggests where the student may have gone wrong, gives more

* Head of Engineering Training Department.

information and asks the student to re-read the relevant text and tackle the questions again.

Figs. 1a and 1b (see pages 6 and 7) show the two sides of typical programmed text.

2.2 Programmed Lectures

2.2.1 The Nature of Programmed Lecturing

PL books are an effective way of learning in which the student proceeds at his own pace. In courses as at present organised specific periods need to be allocated for such work: some students finish their program early and need to be provided with other work; others do not complete their PL book and may need to take it away for completion in their own time. However, the preparation of these books is time-consuming.

The programmed lecture consists of a teaching program, conducted with a group of students (twenty to thirty in number) in which demonstration and other visual aids, together with appropriate recapitulation are employed by the lecturer in charge. All students must, of course, proceed through the lecture at the rate determined by the lecturer, but recapitulation when required is determined from moment to moment by the lecturer himself and tailored to meet the needs of the individuals in the class. This treatment demands a continuous contact between the lecturer and each student in his class, which is achieved in a *feedback* classroom.

Teaching programs for the feedback classroom are prepared more economically than programmed text, but, of course, need an experienced lecturer to conduct them. In consequence, programs can be provided for subjects which may need frequent up-dating. Alteration to a feedback classroom teaching program is relatively simple compared with modifications to a programmed text.

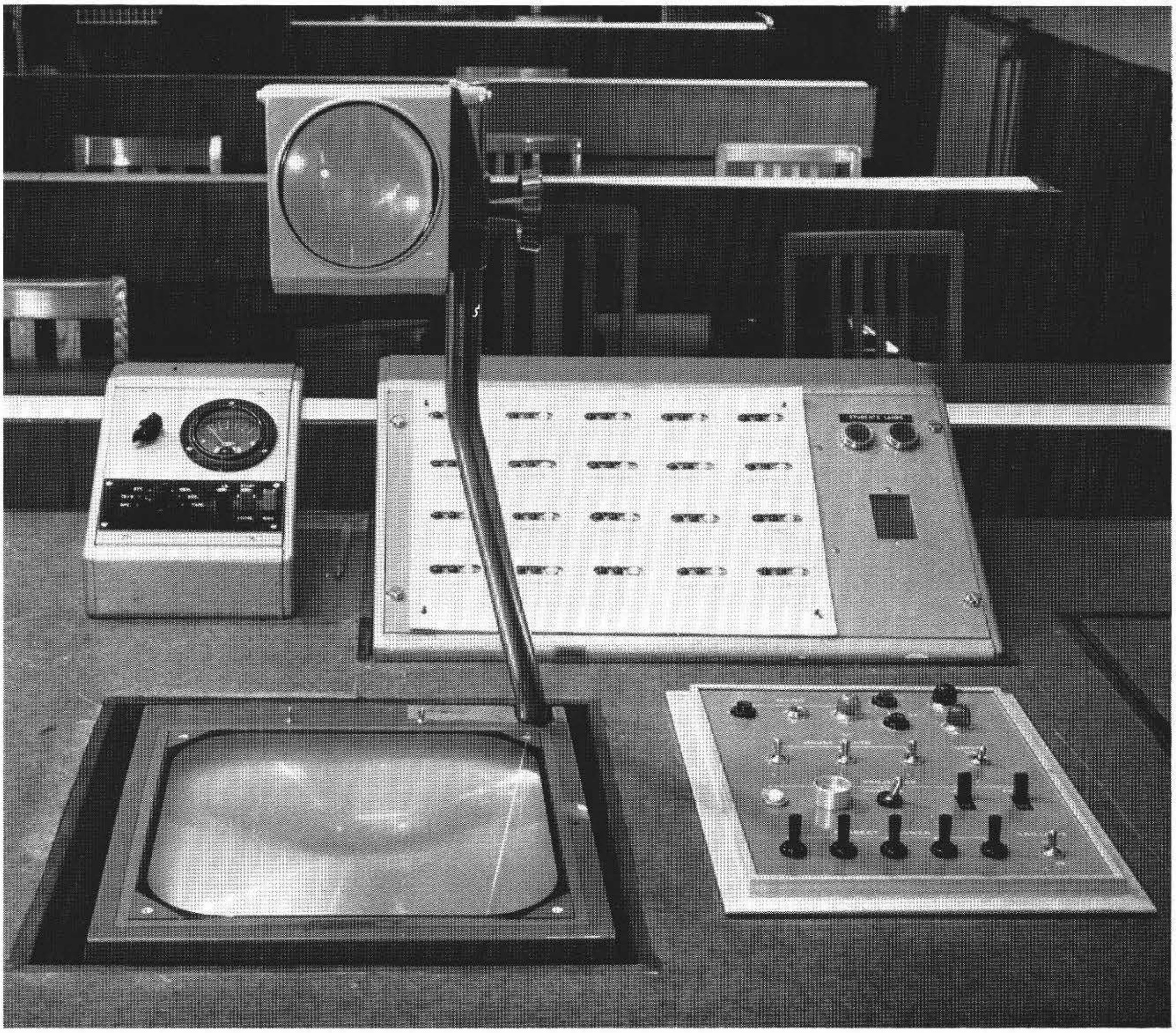


Fig. 2 Equipment on the lecturer's desk in a feedback classroom enables a record of each student's answers to be kept

The sheet superimposed on the pilot lamp panel

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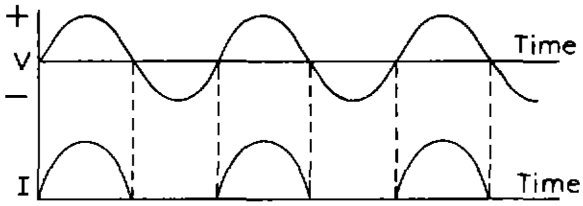
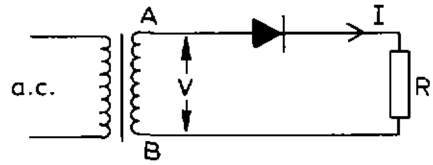
Frame 4

Here are the correct graphs again:-

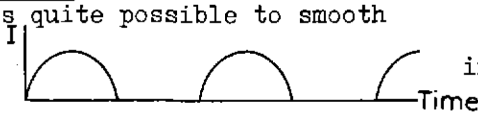




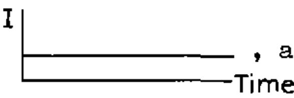
The upper graph shows the voltage across the transformer secondary, A-B, while lower graph shows the current I flowing through R. Current can only flow when A is positive and B is negative so we get pulsations of current. We can describe the current I as "RECTIFIED a.c.", but we can hardly call it "d.c." because this term implies a steady current. However we can obtain d.c. if we SMOOTH the pulsations.

Preliminary note about smoothing

Later we'll deal in detail with the way a pulsating current is smoothed to give good d.c. At this stage we'll simply say that it's quite possible to smooth

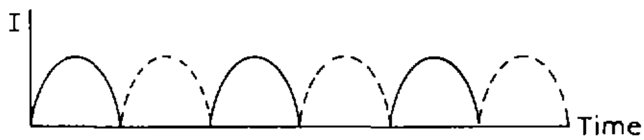


into



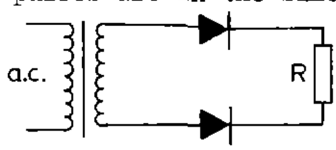
, and indeed this is done in all type of power supplier.

At the moment then, we have pulses of current with large gaps between them. The smoothing process will be made easier if we can arrange for additional pulses to occur in the gaps - like this:-



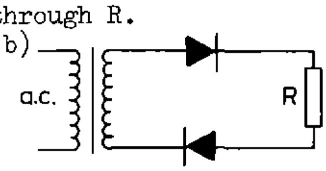
Which one of the following circuits do you think could do this? Trace carefully the currents through each rectifier and see whether the pulses are in the same direction through R.

(a)



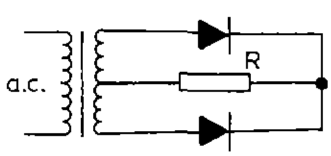
(Frame 4a)

(b)



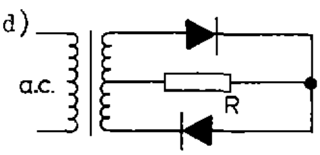
(Frame 4b)

(c)



(Frame 4c)

(d)



(Frame 4d)

(e) I don't know (Frame 4e)

(a) Questions

Fig. 1 Typical pages in a Programmed Learning Book. In practice the questions and answers would not be displayed on facing pages

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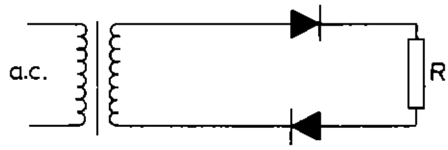
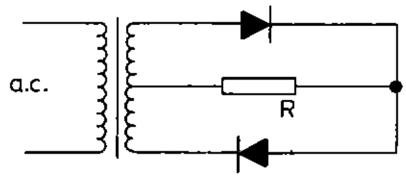
4a
A
If you chose this circuit you ought to have realised that whichever way a current tries to go through R there is a rectifier in the wrong direction, so no current flows at all. Go back to Frame 4 and try again.

4b
B
If the upper part of the transformer is positive then a current will flow through R, but what happens when the lower part of the secondary is positive? Will any current flow? The answer's NO! We're looking for an arrangement that will let current flow in R in the same direction regardless of which half, positive or negative, of the a.c. we're dealing with. Go back to Frame 4 and try again.



4c
C
The arrangement you've chosen is quite correct. Well done. Go on to Frame 5.

4d
D
The circuit you've chosen looks nearly right but in fact it won't do what we want. What is the current path when the upper end of the transformer secondary is positive, and will any current flow when the lower end is positive? In the first case there'll be a short circuit! Try again.



4e
E
If you don't know how to tackle this question imagine that first the upper end of the transformer secondary is positive and see where current can flow. Then imagine the lower end is positive and repeat the process. Go back to Frame 4 and do this for each of the circuits in turn.

(b) Answers

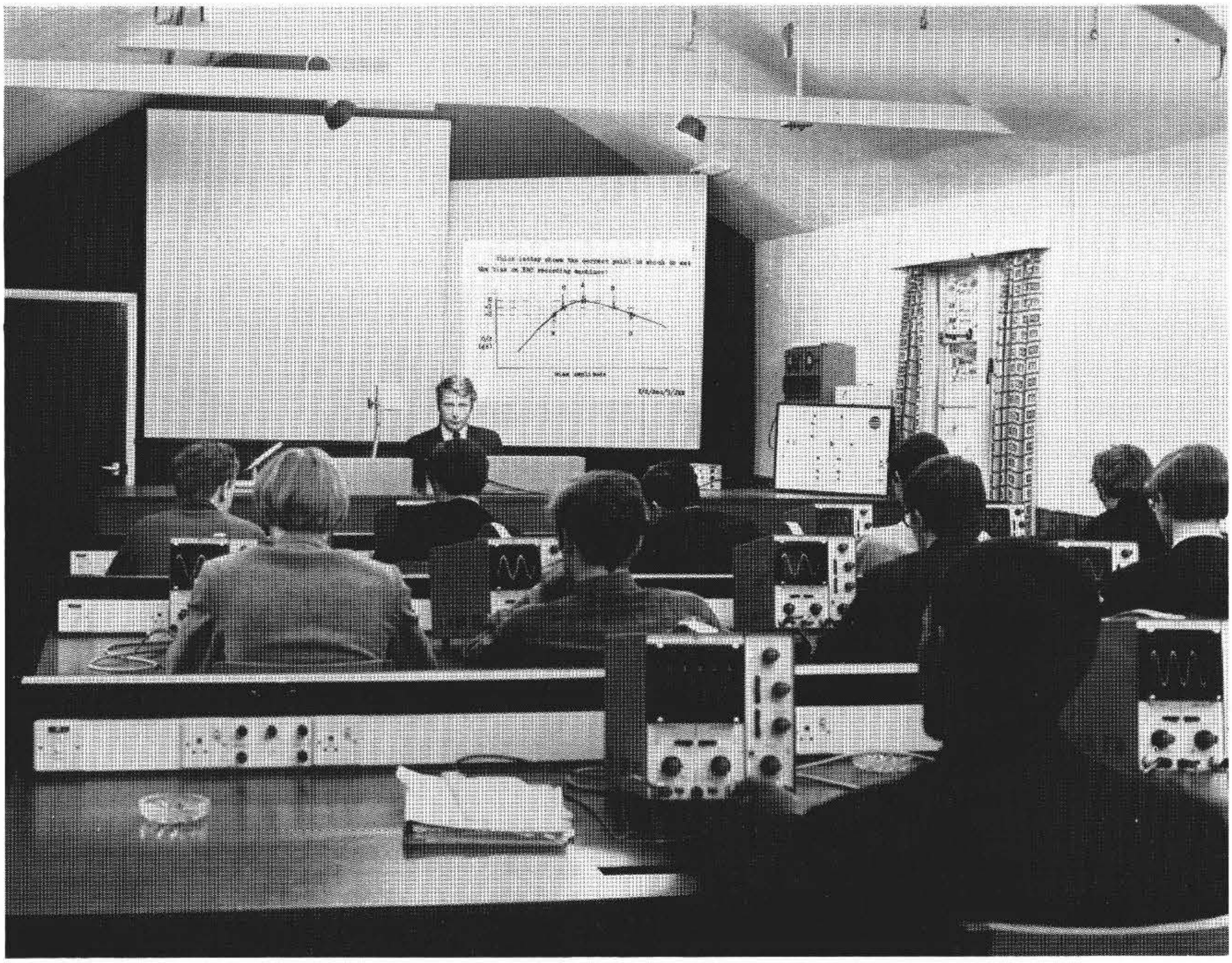


Fig. 3 'Student's-eye' view of a feedback classroom. The slot concealing the recessed panel of answering buttons can be seen to the left of the student nearest to the camera

2.2.2 Layout of Feedback Classrooms

In a feedback classroom, a programmed lesson in theoretical or practical work is conducted by a lecturer who is effectively in communication with each of his twenty students. The accompanying photographs show the equipment on the lecturer's desk (Fig. 2) and that available to each student (Fig. 3). The basic idea was developed by Mr K. Holling at Chesterfield College of Technology.

As in a PL text, the diagrams, pictures, and explanatory text associated with a lecture are presented to the student, this time by an automatic slide projector controlled remotely by the lecturer. On the slide, a question is asked and multiple-choice answers a, b, c, d, and e can be given. The student chooses his answer and presses the appropriate button on a recessed panel before him, which other students cannot see. A sixth button, which operates a white lamp on the lecturer's panel, enables the student to indicate that he does not know the answer. This facility is, however, being dropped from later designs, as experience has shown that a student, rather than admit that he doesn't know the answer, will select an answer he thinks may be the right one. With five choices the score a student can achieve by chance is adequately low. When the student's answer is correct, green lights glow on the lecturer's panel and on the student's desk: if one of the wrong answers is given, red lamps light up in both positions. The lecturer can switch any of the five answering buttons to operate the green lights, and this enables him to find out, if need be, which particular wrong answer any student has given. Before doing this, he will, of course, switch off the lights on the students' desks to avoid confusing any of them by flashing the red light after they have been signalled that their answer is correct.

In this way, the student receives the immediate confirmation necessary to the learning process. Particularly effective has been the development of practical work conducted in a feedback classroom, where each student has his own equipment and is continuously required to make measurements and observations, signalling his results by means of his five push buttons.

Students in such an environment work with enthusiasm and involvement. Everyone feels he is expected to respond to each new question. Those who are in difficulty are readily (yet privately) observed and can be helped at the time or in a later tutorial session. This private and discerning link between the lecturer and each student in the class makes the feed-back classroom particularly useful where students are diffident or represent a wide range of seniority within the organisation.

3 The Student's Day

No student can spend seven hours each day in front of a programmed book. He must have variety and must work with equipment. In fact students at the BBC Engineering Training Centre spend only about one hour each day during the first six weeks of their course with PL texts. Another hour each day is spent in the feedback classroom.

Reference has been made to tutorial activity and this is, of course, where the lecturer has his greatest value. The student has learned basic facts from PL; now he must discuss the interrelation of these facts in practical situations with an experienced tutor.

The day might also include a period in the workshop, demonstrations, and practice in lifting and handling heavy equipment as well as an early introduction to the operation of professional broadcasting equipment.

Thus each day is broken into a number of activities including a maximum of two hours devoted to Programmed Learning by text and in a feedback classroom.

4 Other Teaching Aids

Much consolidation remains to be done in the areas in which PL is to be employed. In other areas members of the lecturing staff are thinking anew of presentation methods. Automatic slide projectors are permanently installed in classrooms and these facilities encourage the accumulation of useful collections of slides for illustration of lecture topics. If a lecturer cannot be assured that a projector is always available for his lecture, and in a room which allows it to be effective, he is unlikely to prepare suitable slides for fear that he will find himself giving the lecture without the benefit of projection facilities: moreover a projector carried from classroom to classroom is liable to be damaged.

Similarly overhead projectors and screens should be permanently installed in a number of classrooms. The first use of such devices is often as a more effective form of blackboard. Then as their great potential becomes apparent, the transparencies used with this projector are prepared with greater and greater forethought. Superimpositions in colour, dynamic demonstrations, and built-up block diagrams replace the earlier chinagraph on acetate sheet. There is scope for development in this form of teaching aid.

Demonstrations can have great teaching impact if carefully designed. Some are complicated to prepare and must be prepared for each occasion. Others such as demonstrations of microphone directivity patterns are difficult to arrange before a large audience. The use of closed-circuit television and video tape recordings may provide a solution to these difficulties although they sometimes introduce other difficulties.

Experience with the use of video tape for teaching senior courses about microphone directivity patterns has shown that whilst the technical content of the taped program was good the presentation on a television screen has contrasted unfavourably with domestic television programmes, thus the attention of students who were themselves engaged in television production was unduly distracted. After borrowing the services of a professional producer, such distractions were eliminated and the technical training content was able to emerge.

5 Indirect Gains Resulting from the Introduction of New Techniques

The earliest gain from the introduction of PL techniques arose from the joint study which members of the lecturing staff had to give to the preparation of the material to be programmed. Searching questions on the prime objectives of the course needed to be answered.

Such questions arise in any course but often tend to be overlooked in the hurly-burly of normal teaching activity. 'Am I teaching this because it is a favourite subject of mine or because the students must understand this in order to deal effectively with some later aspect of their work?' 'Is it essential knowledge or just "nice to know"?' 'Is it already out of date?' Such questions *must* be answered before the heavy investment in program writing can begin.

Another early gain arose from the emphasis which PL places on measurement of progress. Even in a conventional lecture the lecturer who has been required to think about program writing ponders on the effectiveness of his instruction and devises tests to measure this. A consequence of the introduction of PL has thus been the adoption of continuous assessment on a senior course run for post-graduates in which no PL is employed. In general, lecturing staff associated with PL preparation have an increased awareness of the problems of communication in formal teaching.

The introduction of PL and other techniques has led to an examination of the way in which BBC courses are organised. In June (1970) there will begin a regular monthly intake of young men (about twenty-five each month) who will progress through a new three-month training course. The new teaching techniques will be applied during the first two months of each course in two separate areas each solely concerned with a month's theoretical and practical tuition. Thus in each area twenty-five sets of equipment, specially prepared slides, and texts will be in use in every month of the year to achieve a high degree of efficiency. Further, less experienced lecturers (for example men attached to the teaching staff for a year) can conduct much of this learning activity once the complicated preparations have been completed.

This steady flow of trained men into the Operational Broadcasting Service will make on-station familiarisation easier to organise and, in fact, achieves a significant saving in salaries over the previous method of thrice-yearly recruitment and subsequent training. It will also provide a more even load at the Training Centre and allow a better distribution of other training courses with a consequent reduction in the standing training reserve which Operational Departments must hold if they are to release men to attend training courses.

6 Conclusion

Much work lies ahead in developing and expanding these new techniques and considerable help and encouragement is received from industrial, service, and academic establishments who are thinking along similar lines. These techniques have great application in schools and universities and could significantly increase the efficiency of the national education system. Not least amongst the gains would be that first vital step of defining clearly the objectives of the exercise.

The Design and Production of Prototype Thin- and Thick-Film Circuits

D. J. Walker*

Contents

1 Introduction

The technological advances which have taken place in monolithic integrated circuit production have led to a very rapid increase in their use for system design, particularly in applications such as digital circuitry where a standardised circuit approach can be used. The more varied and specialised needs of the linear system designer, however, do not lend themselves to this standardised approach, and the production of relatively small quantities of special silicon integrated circuits is not an economic proposition.

Hybrid thin- or thick-film circuits can offer many of the advantages of the monolithic devices with the added incentives of quick and economic production of relatively small quantities of units having very stable and accurate components. Furthermore, experimental and prototype circuits can be made in a laboratory using a minimum of specialised equipment. Before considering the use of hybrid circuits, however, it is well to be aware of the fundamental differences between the two major types, thin-film and thick-film, their limitations, and to some extent the manufacturing methods employed.

2 Thin-Film Circuits

Thin-film circuits are normally manufactured using a vacuum deposition process. The substrate is generally 0·5 mm thick borosilicate glass, as this has good thermal properties and a high surface finish. Resistive components are formed by depositing a film of metal, usually nichrome or tantalum, approximately 75×10^{-8}cm in thickness, resulting in a sheet resistivity of the order 100 to 300 ohms/square.† The conductor pattern, usually gold or aluminium, is then deposited with a thickness of about 20×10^{-7}cm.

The temperature coefficient of the resistive layer is a function of film thickness, and a value of the order of +80 p.p.m. is typical for 300 ohm/square nichrome. Deposited resistors normally have a best tolerance of ±5%, and greater accuracy than this is obtained by individual adjustment using a diamond scriber or micro engraving techniques (this is not as expensive as it sounds). Accuracies of ±0·1% are readily obtained in this way and resistors on the same substrate have closely matched thermal characteristics. Resistor dissipation can be 1 watt per square centimetre for ambient temperatures of up to 55°C.

Capacitors may be formed by successive deposition of the conductor and a dielectric material such as silicon dioxide or tantalum pentoxide. This process can, however, be expensive particularly if only one or two capacitors are required and a better alternative in this instance is for subminiature leadless ceramic chip capacitors or tantalum electrolytics to be bonded to the substrate as discrete components.

Despite a considerable amount of research effort, the production of active elements by vacuum deposition, as opposed to conventional diffusion, is not a practical proposition and these must be added to the passive circuit as discrete devices. The encapsulation most suitable for this use is the L.I.D. (Leadless Inverted Device) in which a semiconductor chip is mounted on a ceramic carrier 1·9 mm × 0·8 mm. The carrier has gold pads which are soldered or ultrasonically bonded to the conductive film (Fig. 1). A wide range of transistor types

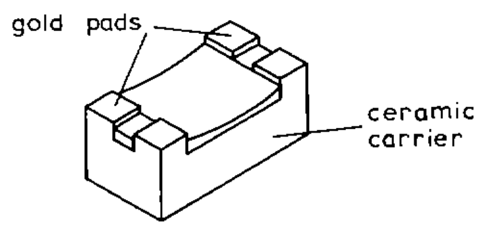


Fig. 1 L.I.D. transistor

can be obtained in this form and in addition, various diodes and zener diodes are becoming available. Another suitable packaging format is the sub-miniature plastic transistor with tape leadouts.

* Mr Walker is with Engineering Designs Department.
† See Appendix 1.

3 Thick-Film Circuits

The conductor and resistor pattern on a thick film circuit is produced by a printing process very similar to that of silk screen lithography.

The 'ink' used is a paste formed from a mixture of glass powder and precious metals including gold, silver, palladium, and other platinum group metals in an organic binder. A platinum/gold or silver mixture is used with the glass for conductor patterns, whilst various combinations of the metals mentioned are used for resistor inks.

After the printing operation the circuit is fired in a furnace at a temperature between 600°C and 1100°C and during this process the glass and metals fuse to form cermet tracks approximately $1{\cdot}5 \times 10^{-6}$cm in thickness. Because of this firing process a ceramic-type substrate material such as 96 per cent alumina is used.

The sheet resistivity of the resistor track is controlled by the proportions of the various metals present in the ink, and values in the range 50ohm/square to 15K-ohm/square with temperature coefficients in the range ±300p.p.m./°C are typical.

The printed resistors have a tolerance of approximately ±15% to ±25% according to area, but this can be readily reduced to ±1% using automatic air abrasion adjustment. Many manufacturers consider this adjustment to be an essential part of the manufacturing process and offer ±1% to ±5% as standard tolerances. Permissible resistor dissipation is in excess of 2 Watts per square centimetre.

Whilst it is possible to make capacitors by successive printing of conductor and dielectric inks the yield rate is poor and added subminiature chip capacitors are by far the most satisfactory solution.

Most of the transistor and diode packages suitable for thin-film circuits can also be used on thick-films, and under some circumstances standard plastic encapsulated transistors are also suitable.

Comparison Table

Resistors	*Thin-Film*	*Thick-Film*
Range	10ohm to 100K-ohm	10ohm to 1M-ohm
Type	Metal film	Palladium/silver cermet
Tolerance	±5% *unadjusted ±0·1% adjusted	±20% *unadjusted ±1% adjusted
T.C.R.	+80p.p.m./°C	±300p.p.m./°C
Stability	Extremely good	normal 'high stability'
Capacitor types	Deposited 12–1000pf, ceramic chip, tantalum electrolytic	ceramic chip, tantalum electrolytic
Active devices	L.I.D. Subminiature plastic Flat pack integrated circuits	L.I.D. Subminiature plastic Flat pack integrated circuits Some T018 type, plastic.

* These are approximate figures only and are given for comparison.

4 Comparison of Thin- and Thick-Film Circuits

The advantages to be gained from the use of film circuits depend to a large extent on the number of modules which are required. If the quantity of units required is less than 100, the tooling charges made for the preparation of artwork and masks will increase the unit cost considerably. Thin-film techniques offer the most economic solution and may be used where it is necessary to obtain a reduction in size with greater reliability compared to conventional components. The technique may also be used to produce closely matched, highly accurate resistor networks. It should be noted that the adjustment of resistors need not necessarily be carried out in order to achieve a particular resistance value; it can also be done to obtain a specified output condition in order to allow for tolerances in other components (e.g. to account for variations in f.e.t. switch resistance in a digital-to-analogue converter using current summing, or to give a specified frequency response in an active filter). In this way it is possible to achieve a degree of accuracy and lifetime stability difficult to obtain by any other method. Thick-film circuits are used in such quantities only for very specialised applications such as uhf micro-strip circuits, which make use of high-dielectric-constant substrate materials to reduce the physical dimensions of tuned lines.

Thin film circuits will again offer the most economic solution when quantities of up to 1000 units are required. The inclusion of a tooling charge does mean that the unit cost will be somewhat higher than the equivalent conventional circuitry but this must be weighed against the advantages of reduced size, higher reliability, and more stable circuitry. Thick-film circuits may be used if a higher component packing density is required.

For larger quantities, say above 1000 units, thick film circuits with their simplified processing and easy adaption to large quantity production methods can be cheaper than the equivalent conventional counterpart. This is particularly true if wiring, testing, and overhead costs in relation to conventional components are taken into account. It should be realised that the unit that is received from the manufacturer is a tested, working circuit, produced in clean conditions with rigorous inspection at every stage, and with each component measured and within tolerance.

Thin-film circuits can also be produced more cheaply than the equivalent conventional counterparts and are used where greater component stability is desirable.

5 Production of Film Circuits in Small Quantities

There are very few restrictions which are placed on the designer when designing a circuit to be manufactured in thin- or thick-film form. There are, however, a few rules which, if sensibly applied, can facilitate manufacture and result in a better end product. It is of advantage to avoid the use of capacitors, particularly of high value, or to ensure that a suitable value, tolerance and temperature coefficient is available in chip ceramic form. Resistor values should be restricted to the range 50 ohms to 30 K-ohms if possible and the effect of tolerances on circuit performance should be carefully considered. It is not necessary to round off to preferred values or call for a higher power dissipation than is required.

It is quite satisfactory when circuit design is complete to hand this to a manufacturer who will then produce a layout for the customer's approval, make the necessary masks, and manufacture pre-production samples. However, in many instances in addition to the use of normal breadboarding techniques it is desirable for the circuit to be proved in film form before a manufacturer is involved. A method of making experimental quantities of thin-film circuits using the minimum amount of specialised equipment has been developed in the BBC Designs Department. The process is basically an extension of printed circuit techniques and uses as the base material a commercially available pre-coated glass substrate, coated on one side with 300 ohm/square nichrome which has been subsequently covered by a layer of gold.

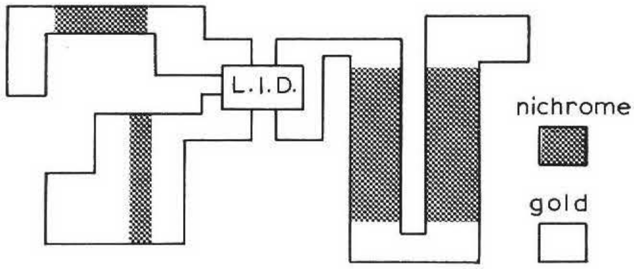
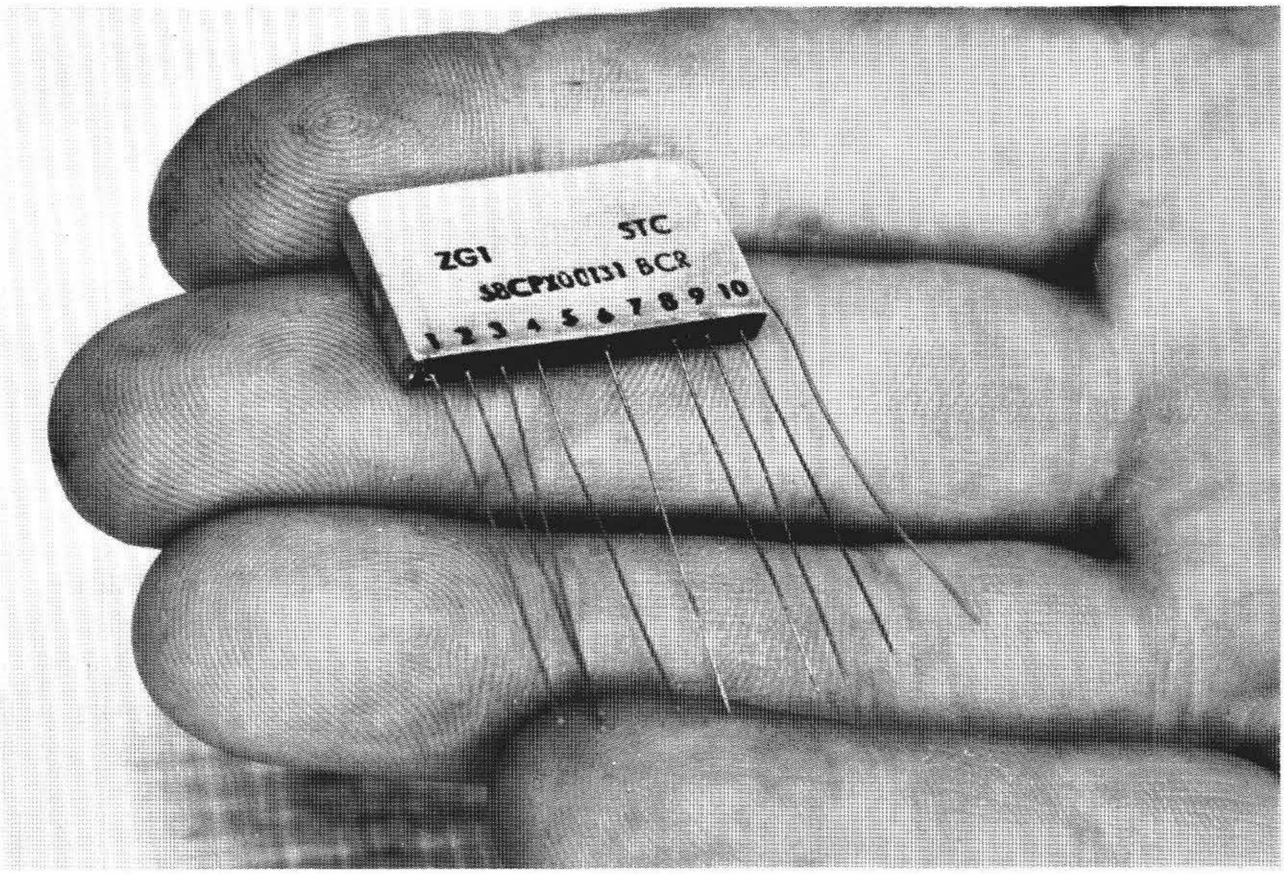


Fig. 2 Film circuit layout



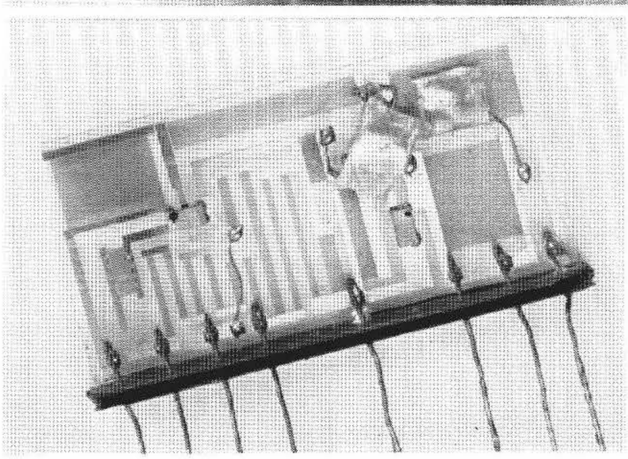


Fig. 3 Thin-film video amplifier module before and after encapsulation

Basically the process is as follows. The resistor dimensions are calculated from knowledge of the value and dissipation required, and the sheet resistivity of the nichrome film. Two master diagrams are produced, one having a pattern representing both conductor and resistor areas ('circuit master') and the other having just the conductor pattern ('conductor master'). These can be made using draughting tape on plastic sheet in the same way as printed circuit masters, and are photographically reduced to produce the working negatives. A representation of a section of circuit is shown in Figure 2 and this indicates a way of laying out the resistors which considerably reduces inaccuracy due to negative misregistration.

The substrate is cleaned and coated with Kodak Photo Resist (K.P.R.) which is initially air dried and then baked at a temperature of 60°C. The resist is exposed to ultra-violet light with the circuit negative in place, developed, and rinsed, leaving conductor and resistor areas resistant to etchant. The unwanted gold is then removed using etchant 1 (Appendix 2) and the nichrome with etchant 2. The old resist is stripped off, new resist applied as before, exposed with the conductor negative in position, and developed. The gold is then removed from the resistor areas with etchant 1. After the resist is removed and the substrate cleaned, discrete elements (transistors, capacitors, lead out wires, etc.) can be fixed in place using low melting point solder. It is preferable to use temperature-controlled equipment especially designed for this application although a temperature-controlled miniature iron has been used with some success. It is advisable to preheat the substrate to 100°C before soldering in order to reduce thermal shock.

Two BBC designs are currently being manufactured as film circuits, a thin-film video amplifier* (Fig. 3), and a thick-film version of the shunt switch element in the solid state video switching matrix MA2/501. The thick-film circuit contains a two-transistor feedback amplifier having a grounded base input stage which maintains an input impedance of less than 1 ohm over the whole of the video frequency range. Included on this circuit are two transistors, four resistors, and a capacitor. In quantity production it will be cheaper than its conventional counterpart.

* ST&C Type 58CPX00131BCR.

6 Appendix 1: Determination of resistor values

$$\text{Resistance} = \frac{\rho . L}{W . T}$$

where ρ is the bulk resistivity.
L is the length of conducting path
W is the width of conducting path
T is the thickness of conducting path

when $\frac{L}{W} = 1$ i.e. a square of any size

$$R = \frac{\rho}{T} = \text{sheet resistivity in ohms/square}$$

$$\text{thus } R = \frac{L}{W} \times \text{sheet resistivity.}$$

For a film having a sheet resistivity of 300 ohms/square a rectangular patch twice as long as its width has a resistance of $\frac{2}{1} \times 300 = 600$ ohms, etc.

7 Appendix 2: Composition of etchants

Etchant 1 (for etching gold)

Potassium Iodide	50 g
Iodine	15 g
Water to	250 cc

The potassium iodide is dissolved in a quantity of water, the iodine dissolved in this solution and water added to make 250 cc.

Etchant 2 (for etching nichrome)

Ceric Sulphate	25 g
Nitric Acid (70% HNO_3)	38 cc
Water	25 cc

The nitric acid is diluted with some of the water before the ceric sulphate is added.

A Receiver for Measuring the Radiated Harmonics of uhf Television Transmissions

N. H. C. Gilchrist*

Introductory note from the Superintendent Engineer, Television and vhf Radio Transmitters.

The C.C.I.R. XIth Plenary Assembly held at Oslo in 1966 considered the question of protecting the frequencies used by Radio Astronomers and recommended that '... administrations should afford all practicable protection to the frequencies used by radio astronomers'. This particularly referred to the band 1400–1427 MHz and also allows observations of the natural OH line emissions: taken as 1665·4 and 1667·4 MHz. The C.C.I.R. will make general recommendations based on user experience as to what figure is eventually acceptable, practicable, and economical to achieve. The receiver method described below will enable power levels to be measured which should in general ensure that meaningful checks can be made down to better than −100 dB relative to peak (sync) power.

Summary: This article describes a receiver which, as part of the work to assess possible interference to radio astronomy arising from Band IV/V television stations, was developed for measuring low-level harmonic radiation in the frequency range 1 to 2 GHz. The receiver employs a simple commutator detector which, with the fundamental signal square-wave modulated, enables input signals of about 0·5 μV to be measured. The complete receiving equipment (including a horn aerial) enables harmonic field strengths of about 8 dB (μV/m) to be measured, in the presence of fundamental field strengths up to about 120 dB (μV/m).

The complete equipment is readily transportable, it can easily be tuned over the frequency band and it is simple to operate. The article discusses features of the design which could be changed if the sensitivity needed to be further improved, but, as the article points out, this can only be achieved at the expense of simplicity and versatility.

Contents

1 Introduction

At all transmitting stations, it is important to ensure that no out-of-band radiation exceeds a level which could cause interference to other radio services. The maximum permissible level of such radiation from uhf television stations is not at present fully specified in precise terms but equipment is normally installed which ensures that no individual out-of-band spectral component exceeds −60dB relative to peak sync power. This limit is found to be generally acceptable but there are exceptions, the most important of which applies when the radiation occurs in a radio-astronomy band. In these parts of the radio spectrum the out-of-band radiation may require to be reduced to a much lower level.

The only radio-astronomy bands considered in this report are those containing the hydrogen (H) and hydroxyl (OH) 'line' emissions. Although these line emissions occur only at specific frequencies, it is necessary to provide bands of frequencies for measurement purposes in order to allow for Doppler shift. The hydrogen-line band is generally defined as 1,400 to 1,427 MHz and, in that band, the Post Office in consultation with the radio astronomers have specified interference power equivalent to steady signals of −41 dB (μV/m) and −34 dB (μV/m) as being harmful at the radio observatories at Jodrell Bank and Cambridge respectively. The hydroxyl-line band is not so clearly defined but, for the purpose of this article, it is taken to be 1645 to 1675 MHz. Harmful levels of interference in this band also have not been clearly defined, but it is likely that they are of the same order as those in the H-line band. Bearing in mind that the peak sync field strength of uhf television transmissions may exceed 70dB (μV/m) at the radio observatories, it is clear that spurious emissions in these bands should ideally be suppressed below −100dB relative to peak sync power.

Radiation arising from second and third harmonics of the

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carrier frequencies, and from intermodulation between the carriers, will occur in one or both of these two radio-astronomy bands at 49 of the 61 main uhf television stations now being planned or implemented. It is reasonable to suppose that a similar proportion of relay stations could also cause interference in this way.

The question of interference to radio astronomy therefore constitutes an important factor when planning uhf stations. It is possible to specify high-power filters which would provide adequate suppression but, first, the level at which the interfering signals are generated in the transmitters must be known because this determines the attenuation required. In addition, the level at which they may be regenerated in the aerials and feeders must be known because this determines the attenuation (and thus the maximum degree of suppression) that it is technically feasible to provide.

It was therefore desirable to attempt to measure low-level signals (e.g. 1 to 10 μV) in the frequency range 1 to 2 GHz, in the presence of high-level signals (e.g. 100 mV to 1 V) in Bands IV and V. Because no suitable receiver was commercially available, the receiver described in this report was developed.

2 The Receiver

2.1 General Description

The receiver comprises an Air Ministry receiver type 1294, modified to improve its sensitivity and noise factor. This receiver has a cavity-tuned crystal mixer, without an r.f. stage, a 13·5 MHz i.f. amplifier with an overall bandwidth of 250 kHz, and an envelope detector followed by a two-stage video amplifier. Three modifications were made to the basic receiver:

(i) Because valves were no longer available for the original local oscillator, an external local oscillator was provided. Arrangements were made to monitor its frequency using a digital frequency counter.

(ii) A cascode i.f. pre-amplifier was installed to increase the i.f. gain and improve the noise factor.

(iii) A mixer-diode current meter was fitted; this was necessary as it was found that there was a restricted range of diode current (200 to 300 μA) within which the optimum noise performance was obtained.

The noise factor of the modified receiver is 5 dB at 1 GHz rising to 9 dB at 1·7 GHz (measured at optimum mixer-diode current). In this condition the open-circuit input signal from a 50 ohm source at room temperature which produces a signal output power equal to the noise output power is 1·6 μV at 1 GHz, increasing to 2·6 μV at 1·7 GHz.

2.2 Method of Detection

In order to improve the overall sensitivity of the receiver without undue complication a commutator detector was provided subsequent to the envelope detector of the basic receiver. The principle of the commutator detector is as follows. The received signal is square-wave modulated at a low frequency; the output of the receiver envelope detector thus comprises a square wave with noise superimposed. This output is fed into the commutator detector together with a reference signal consisting of the (noise-free) square-wave modulation. The detector compares the receiver outputs during the two half-cycles of the reference signal and produces a d.c. output proportional to the difference between the two conditions.

It is possible to operate a very simple commutator detector working on this principle by square-wave modulating the signal incoming from the aerial, but this method has three rather serious drawbacks. Firstly, the modulator in the aerial feeder introduces loss. Secondly, the action of the modulator causes transients at the receiver input which are synchronous with the reference signal. The receiver therefore gives an output indication when no input signal is applied. Finally, it

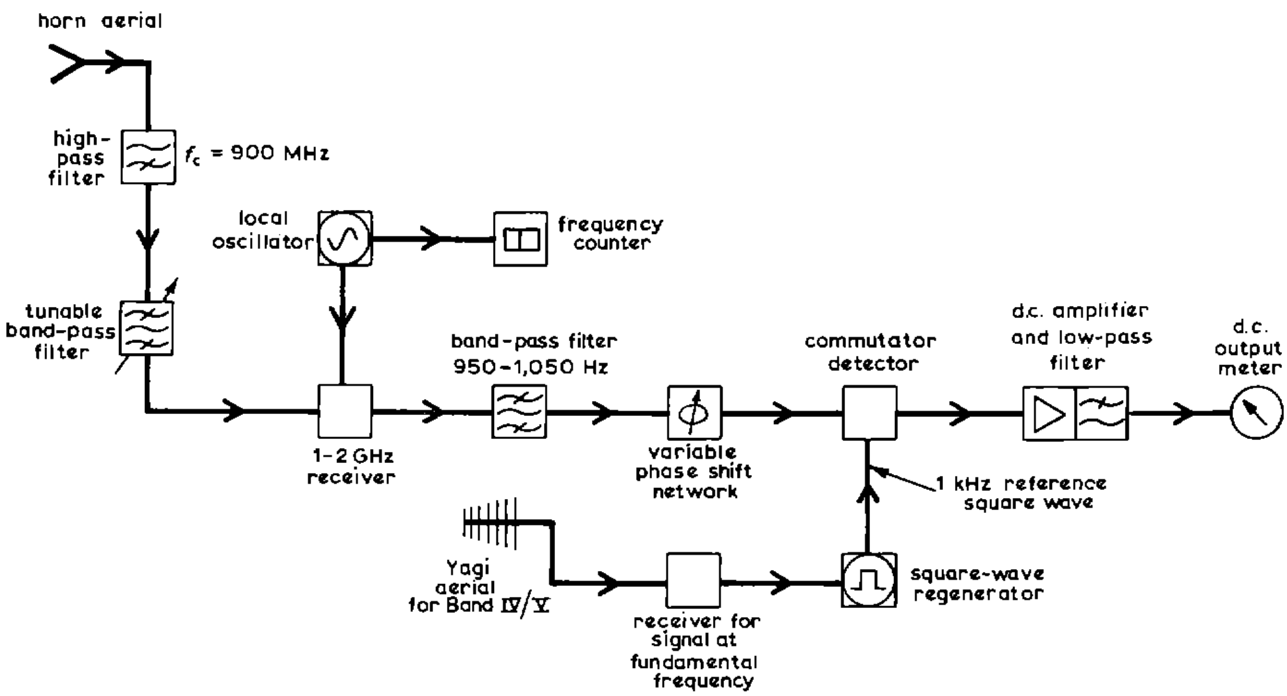

Fig. 1 Block diagram of receiving system

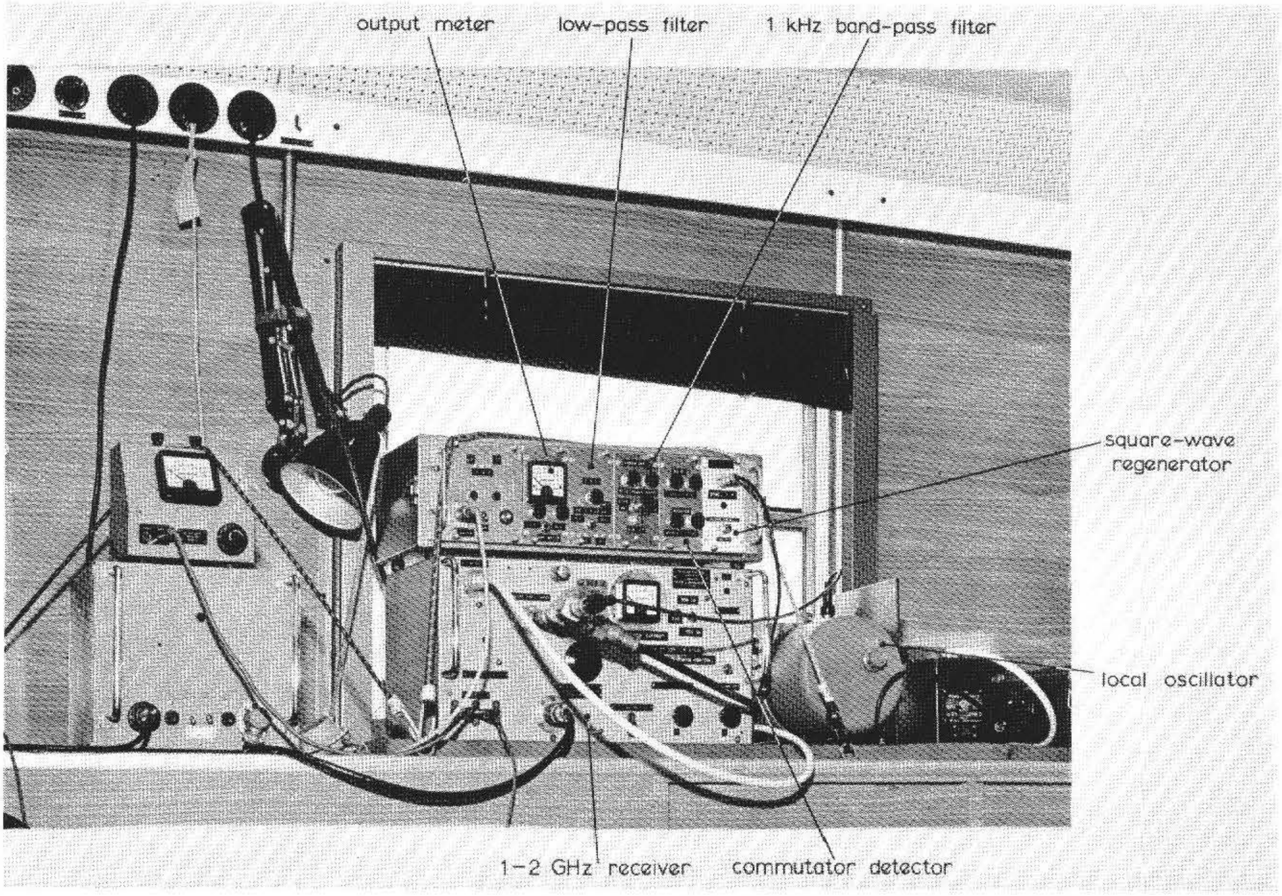

Fig. 2 The receiver and commutator detector assembly

should be noted that the only noise which a system of this type can discriminate against is that which arises in the receiver; any interfering signals picked up by the aerial will be treated as wanted signal by the device.

In view of these drawbacks an alternative system was adopted whereby the transmitter was 100 per cent modulated by a 1 kHz square wave; a reference signal for synchronising the commutator detector was derived from a separate receiver tuned to the fundamental signal whilst the measuring receiver was tuned to the required harmonic frequency. In order to provide a noise-free reference signal, the reference square wave was regenerated locally using a trigger circuit.

A block diagram of the complete system is shown in Fig. 1. High-pass filters in the aerial feeder prevented harmonic generation in the receiver from the high-level fundamental signal. The tunable bandpass filter also helped in this respect, but its prime function was to permit the identification of the received harmonic signal. The receiver was sufficiently well screened for it to operate satisfactorily in a high ambient fundamental field strength (but not sufficiently well for it to be used in transmitter buildings). The receiver output signal was fed to the commutator detector through a 1 kHz band-pass filter and a variable phase-shift network; the latter enabled the signal phase to be adjusted to give maximum sensitivity. The commutator detector consisted of a differential amplifier, using a pair of transistors; the signal output from the main receiver was fed co-phased to the two emitters, the reference signal driving the two bases in antiphase. Thus the two transistors were alternately switched on and off by the reference signal synchronously with the modulation on the received signal. Outputs were taken from the two collectors to a d.c. differential amplifier driving a d.c. output meter via a low-pass filter.

Under no-signal conditions, there is no deflection of the output meter since the square waves at the two collectors of the commutator detector are equal in amplitude. When a synchronous signal is present, however, an unbalance occurs which causes a reading on the meter. Residual noise fluctuations are largely removed by the low-pass filter which should, ideally, have a long time-constant. In practice the time-constant that could be usefully employed was limited by drift in the d.c. amplifier.

For use, the complete equipment was mounted in a mobile laboratory as shown in Fig. 2. Fig. 3 shows the arrangement of aerials on the vehicle. The small dish aerial was used to radiate a low-level signal to test the equipment.

3 Performance

3.1 General

The sensitivity of the complete equipment was measured at both 1·4 GHz and 1·7 GHz in preparation for harmonic measurements which were to be made at approximately those frequencies.

3.2 The Receiver

Figs. 4a and 4b show the variation of output indication as a function of input signal level for the receiver, with and without the commutator detector, at 1·4GHz and 1·7GHz respectively.

The minimum measurable input signals were found in practice to be as follows:

	at 1·4GHz	*at 1·7GHz*
Using envelope detector (main receiver) only	2·5 μV	2·0 μV
Using envelope detector and commutator detector	0·5 μV	0·4 μV

These figures show that the commutator detector improved the sensitivity by 14dB at both frequencies.

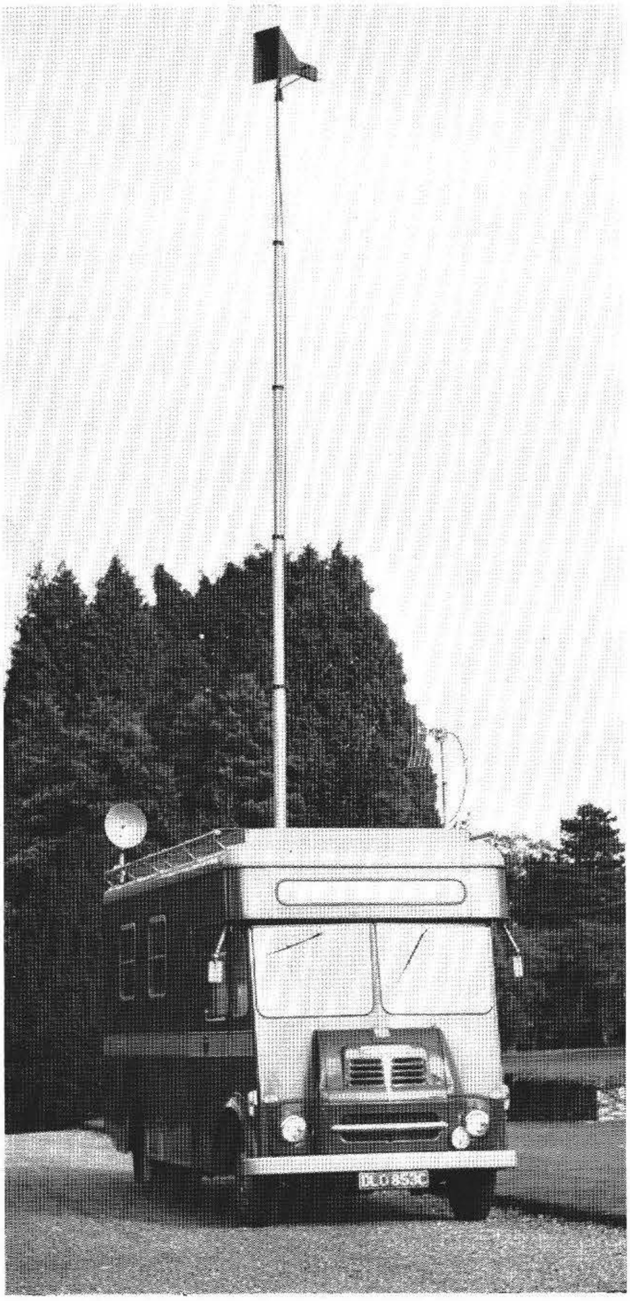

Fig. 3 The measuring vehicle, showing the telescopic mast fully extended with the horn aerial in position

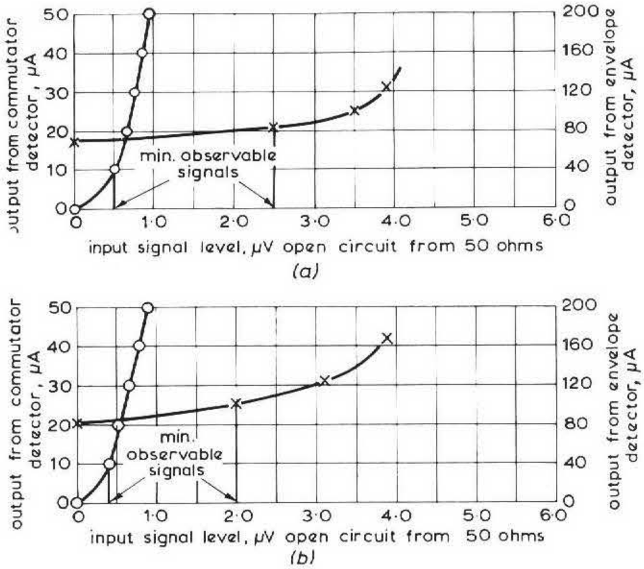

Fig. 4 Output indication as a function of input signal level (a) at 1,4 GHz (b) at 1,7 GHz

X Output from envelope detector

O Output from commutator detector

3.3 The Aerial

The aerial used with the harmonic measuring receiver comprised a rectangular brass horn with an aperture 300mm by 450mm. A photograph of the aerial is shown in Fig. 5. Relative to a $\lambda/2$ dipole, its intrinsic gain was 12dB at 1·4GHz and 14dB at 1·7GHz. The feeder loss was 3dB and 4dB respectively at the two measuring frequencies. The net gain of the aerial system was thus 9dB at 1·4GHz and 10dB at 1·7GHz.

3.4 Maximum Overall Sensitivity of Measurement

The results given in Sections 3.2 and 3.3 enable the maximum sensitivity of the overall arrangement to be specified as follows:

	at 1·4GHz	*at 1·7GHz*
Minimum measurable signal at receiver input terminals from 50ohm source	0·5 μV	0·4 μV
Equivalent field strength assuming a half-wave dipole	7·3 μV/m 17dB(μV/m)	7·1 μV/m 17dB(μV/m)
Net aerial gain	9dB	10dB
Minimum measurable field strength (using the aerial described in Section 3.3)	2·5 μV/m 8dB(μV/m)	2·2 μV/m 7dB(μV/m)

3.5 Further Improvements in Sensitivity

There are three ways by which the measurement sensitivity could be further increased if required:

Fig. 5 The horn aerial

(i) The aerial gain could be increased by increasing its aperture; to increase the gain by 3 dB the aperture area would need to be doubled. When measuring, however, it is important to be able to explore the field over a range of aerial heights from 3 to 6 metres a.g.l. This would be less convenient if the aerial dimensions were increased above those specified in Section 3.3.

(ii) An r.f. stage could be added to the receiver. This would improve the noise factor (and hence the sensitivity) by a few decibels but, in view of the extra complication and the possible extra difficulty of tuning, it would appear not to be worthwhile.

(iii) The noise bandwidth of the receiver could be reduced at any of four parts of the receiver:

(a) Before the mixer. The tunable bandpass filter associated with the equipment has a noise bandwidth of about 30 MHz. It would not be practicable to reduce this further nor, in view of the narrower i.f. bandwidth (see below), would it be helpful except insofar that it may reduce noise or interference at the image channel response of the receiver.

(b) Before the envelope detector. The sensitivity could be increased by 10 dB if the i.f. bandwidth were decreased to 25 kHz but this would require a local oscillator having greatly increased frequency stability. The receiver would probably be less easy to tune from one harmonic to another and the general facility of measurement would be reduced.

(c) Before the commutator detector. A band-pass filter with noise bandwidth about 100 Hz was fitted between the envelope detector and the commutator detector (see Fig. 1). Because of the narrow bandwidth of the low-pass filter subsequent to the commutator detector (see below), no improvement would be obtained by further decreasing the bandwidth of this band-pass filter.

(d) Before the output meter. A low-pass filter with noise bandwidth of about 0·3 Hz was fitted between the commutator detector and the output meter. The sensitivity could be increased by further decreasing this bandwidth but the drift with time in the d.c. amplifier would then also need to be reduced. It also follows that the measurement time would be increased; this would be inconvenient when, for example, tuning the receiver or moving the aerial to explore a field pattern.

It is evident that any increase of sensitivity would entail a reduction in the general facility of measurement. If, nevertheless, increased sensitivity is demanded, it would appear best to decrease the i.f. bandwidth and to accept the need for a crystal-controlled local oscillator.

4 Conclusions

The receiver described in this report enables the field strength of harmonic radiations in the frequency range 1 to 2 GHz to be measured down to about 10 dB above 1 μV/m in the presence of fields at the fundamental frequency up to about 120 dB above 1 μV/m. It is of straightforward construction, simple to operate, and may readily be retuned from one harmonic to another. The aerial is reasonably small and manœuvrable and it may be moved to explore the radiated field as a function of aerial height or position.

Experience has shown that the whole can readily be used to measure harmonic levels down to -100 dB relative to peak sync power but only in situations near to the transmitter where the fundamental field strength is large. This is a serious restriction because it can generally only be used in the sidelobe pattern of the v.r.p. of the transmitting aerial and cannot be used in the distant field. If measurements in the distant field are later shown to be necessary a more sophisticated receiver along the lines discussed in Section 3.5 will also be necessary. It is evident that such a receiver would be less convenient to use and special arrangements may also be required to manœuvre a larger aerial.

A non-synchronous method of demodulating amplitude-modulated signals with asymmetrical sidebands

K. Hacking*

Summary: The theoretical requirements for demodulating a.m. signals with asymmetrical sidebands are outlined. A novel form of demodulator is described which makes use of the variation in the phase of the modulated carrier to derive a signal that can correct the output of a normal envelope detector for non-linear distortion. Several schematic arrangements are given together with the circuit used in an experimental demodulator. The limitations and advantages of the method are discussed and compared with those of a conventional synchronous demodulator.

Contents

1 Introduction

It is well known that the envelope of an amplitude-modulated carrier wave is not a faithful representation of the original modulating signal if the symmetry of the upper and lower sidebands is perturbed. In order to preserve envelope fidelity, it is necessary that the amplitude of the upper and lower side-frequencies are equal and that their phases are precisely anti-symmetrical with respect to the carrier phase. Moreover, the carrier must at all times be of sufficient amplitude to prevent overmodulation. The output of a conventional envelope detector is therefore a distorted version of the original modulating signal whenever sideband asymmetry exists at the input to the detector. Furthermore, the kind of distortion introduced (sometimes called quadrature distortion) is non-linear so that harmonics and intermodulation products of the original signal spectrum appear as spurious components in the demodulated version. Sideband asymmetry at the input to the detector may arise due to propagation phenomena, mistuning in the receiver, or be deliberately introduced at the transmitter to conserve bandwidth as, for example, in a vestigial-sideband television transmission.

There are several methods of demodulating amplitude-modulated signals with asymmetrical sidebands which avoid the non-linear distortion associated with envelope detection. The most common of these, and certainly the most elegant and complete solution conceptually, is synchronous demodulation. This well-known method, which is discussed in more detail later, relies on providing in the receiver a periodic signal which is in precise synchronism with the incoming carrier. This requirement is not easy to implement in many applications, e.g. a domestic m.f. radio receiver, without a substantial increase in cost. Another approach is to use an envelope detector and to apply to its output a correction which cancels the non-linear distortion components. Voelcker[1] has shown that for a single-sideband transmission the correction signal required can, under certain constraints, be derived entirely from the (distorted) output of an envelope detector. One constraint is that the carrier component is sufficiently strong, so that overmodulation does not occur, and another is that the absolute carrier level at the input to the envelope detector is constant, which implies precise a.g.c. circuitry. The proposed new method,[2] which is the subject of this article, is similar in approach but the required correction signal is derived from the variation in the *phase* of the modulated carrier. The potential advantage of this approach, over classical synchronous demodulation, is that the need to provide a regenerated carrier in the receiver is removed, and this could lead to a cheaper and simpler form of demodulator. There are, however, important limitations of the method which arise due to phase ambiguity and these may restrict its general application.

* *Mr Hacking is with Research Department.*

2 Theoretical Basis of the Method

The effect of sideband asymmetry on the carrier wave can be seen by comparing an example of single-sideband modulation (which is clearly an extreme form of sideband asymmetry) with that of a symmetrical double-sideband transmission. In the latter case, a single component of the modulating signal gives rise to a symmetrical pair of side-frequencies which in the conventional vector representation of Fig. 1a are shown by a pair of counter-rotating vectors, BC and BD, with respect to the carrier component AB. The resultant BE of the

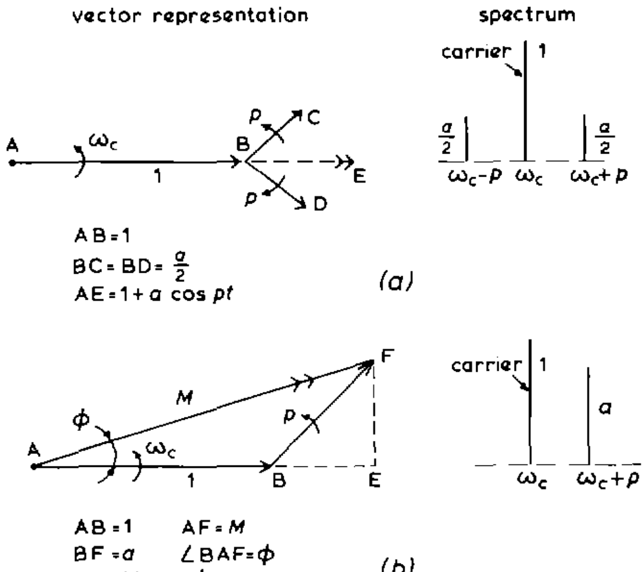

Fig. 1 Representation of single-tone amplitude modulation:
(a) Double-sideband (sideband symmetry)
(b) Single-sideband (sideband asymmetry)

side-frequencies is always in phase with the carrier so that AE represents the instantaneous amplitude of the carrier wave. Hence, from Fig. 1a, an envelope detector would give an output $1+a\cos pt$, where $a\cos pt$ is the original modulation component and the carrier is of unit amplitude. Therefore, apart from a constant d.c. term, the envelope is an undistorted version of the original signal. Consider now Fig. 1b which shows a single-sideband arrangement having the same modulation depth. The spectrum now has a single side-frequency of amplitude a and angular frequency (ω_c+p) where ω_c is the carrier frequency. The vector representation for this case shows that the resultant, AF, is not always in phase with the carrier, the phase difference being a function of the amplitude and phase of the modulating signal. The instantaneous amplitude M of the carrier wave is not linearly related to the original modulation $a\cos pt$, as it was in the symmetrical double-sideband situation. However, it is easily seen from the vector geometry that the required output is in fact $M(t)\cos\phi(t)$, where the M and ϕ are written as functions of t to emphasise their time dependence.

The above examples relate to single modulation components for simplicity, but it can be shown that a general expression for a modulated carrier-wave $E(t)$ is given by

$$E(t)=M(t)\cos[\omega_c t+\phi(t)] \qquad (1)$$

and that the correctly demodulated output (in the sense that it is free from non-linear distortion) is $M(t)\cos\phi(t)$. That this is correct is shown by considering the output from an ideal synchronous detector. Let the input to the synchronous detector be given by Equation (1), and this be multiplied in the detector by the regenerated carrier $2\cos\omega_c t$. The product is

$$2E(t)\cos\omega_c t=2M(t)\cos\omega_c t.\cos[\omega_c t+\phi(t)]$$
$$=M(t)\Big\{\cos\phi(t)-\cos[2\omega_c t+\phi(t)]\Big\}$$

which, after passing the signal through a low-pass filter, leaves $M(t)\cos\phi(t)$ as the demodulated output. Thus the required undistorted output is obtained if the function $M(t)\cos\phi(t)$ can be accurately generated from the carrier-wave.

3 Approximate Non-Synchronous Methods

The $M(t)$ factor is easily obtained from the output of an envelope detector, but the $\cos\phi(t)$ factor is more difficult to obtain. The method suggested here is to extract the phase modulation component $\phi(t)$ from the incoming carrier wave and then synthesise the cosine function by approximate means. For example, one basic arrangement of the method is shown in Fig. 2a. The output from an intermediate-frequency amplifier, $M(t)\cos[\omega_i t+\phi(t)]$, where ω_i is the intermediate frequency, is split into two paths. In one path the signal is passed through a symmetrical limiter to remove the amplitude modulation. The limiter output $\cos[\omega_i t+\phi(t)]$ is then fed to a frequency discriminator to obtain $\omega_i+\mathrm{d}/\mathrm{d}t\{\phi(t)\}$. The d.c. component is removed and the signal passed through an integrator to obtain a signal proportional to $\phi(t)$. The phase component is then squared and multiplied by the output from an envelope detector in the alternative path to obtain a correction signal $\beta M(t)\phi^2(t)$, where β is an arbitrary gain factor. Finally, the correction signal is subtracted from the output of the envelope detector to give $M(t)-\beta M(t)\phi^2(t)$ as the final output of the demodulator. Setting $\beta=\frac{1}{2}$, the output is seen to be a first order approximation to the function $M(t)\cos\phi(t)$ for, by series expansion,

$$M(t)\cos\phi(t)=M(t)\left\{1-\frac{\phi^2(t)}{2}+\frac{\phi^4(t)}{24}-\cdots\right\} \qquad (2)$$

For $\phi(t)$ equal to $\pi/2$ radians, the error in the correction signal due to the approximation is $(\pi^2/8):1$, which is 22% approximately. However, in practice, the value of β would be adjusted to spread the error more uniformly. For example, with $\beta=1/2{\cdot}2$ the correction signal would be in error by not more than $\pm10\%$ over the range $0\leqslant\phi(t)\leqslant\pi/2$ which corresponds to the range of modulation depths 0 to 100%.

To illustrate the correction process, Fig. 3 shows the waveforms at the points A to F in Fig. 2a when the input signal comprises a carrier and a single side-frequency component (90% of the carrier amplitude).

Alternative circuit arrangements are shown in Figs. 2b and 2c. In Fig. 2b the correction is effected before envelope detection. This may be advantageous if the envelope detector is imperfect, introducing its own distortions. In Fig. 2c a triple-input linear multiplier is used to form the product $M(t)\,(1-\phi(t)/\sqrt{2})\,(1+\phi(t)/\sqrt{2})$ which is equivalent to the output of the other arrangements.

4 Limitations

Apart from small errors due to the approximation to the cosine function, the performance of the non-synchronous demodulator described above is similar in every respect to that of a conventional synchronous demodulator except that it will not deal with overmodulation. The difficulty is that when the input signal runs into overmodulation there is a carrier-phase jump of π radians on entering and leaving the overmodulation excursion. This gives rise to a pulse at the output of the frequency discriminator which perturbs the correction signal and therefore the final output. The mechanism is similar to that which produces the well-known 'plops' in f.m. demodulators subjected to interference with a peak value exceeding the carrier amplitude. Fig. 4 illustrates the

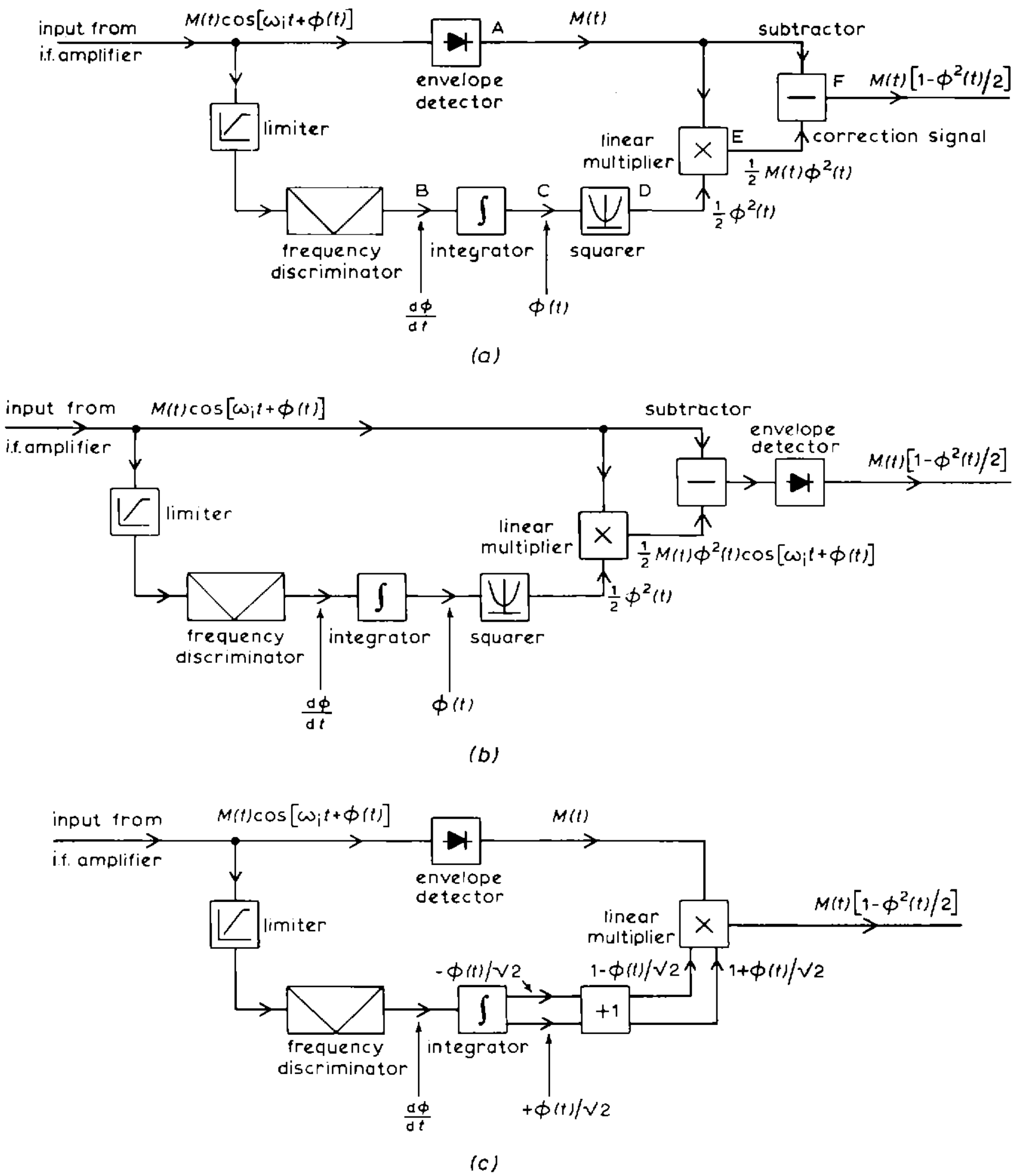

Fig. 2 Non-synchronous detector arrangements

effect of overmodulation for the particular case of a single-sideband input signal comprising a carrier and one side-frequency which progressively increases in amplitude until it exceeds that of the carrier and then subsides. In Fig. 4 the voltage across the load at the output of the frequency discriminator is shown as a function of time. Until the side-frequency component exceeds that of the carrier the mean voltage per cycle of modulation is constant, its value corresponding to the carrier frequency. During the overmodulation period, however, the mean voltage (although still constant in this particular example) has a different value, in fact that corresponding to the side-frequency. In this manner, the ideal output from the discriminator becomes marred by the addition of a rectangular pulse whose height is proportional to the difference in frequency between the carrier and that of the side component, and whose width is proportional to the

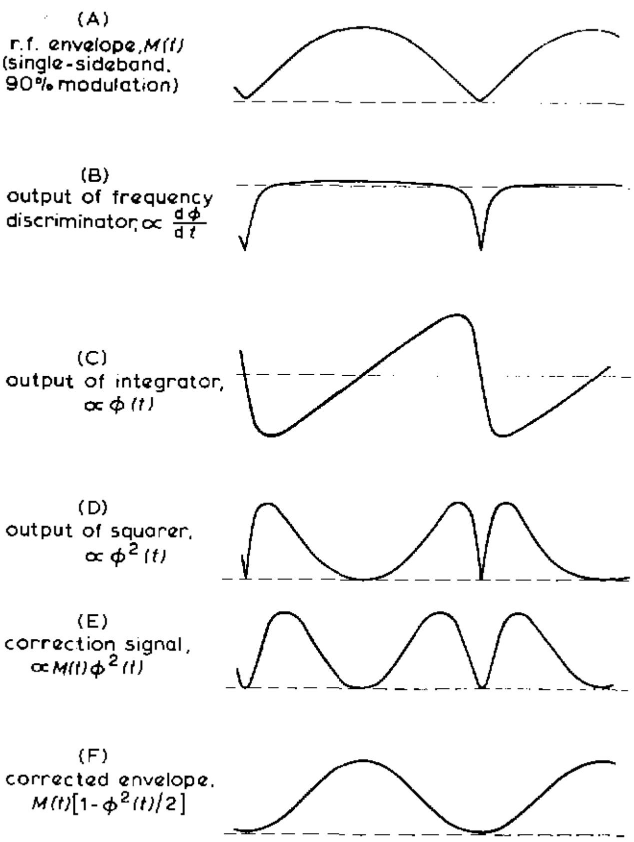

Fig. 3 *Waveforms at various points in detector chain (see Fig. 2(a)), for single-tone modulation and single-sideband transmission*

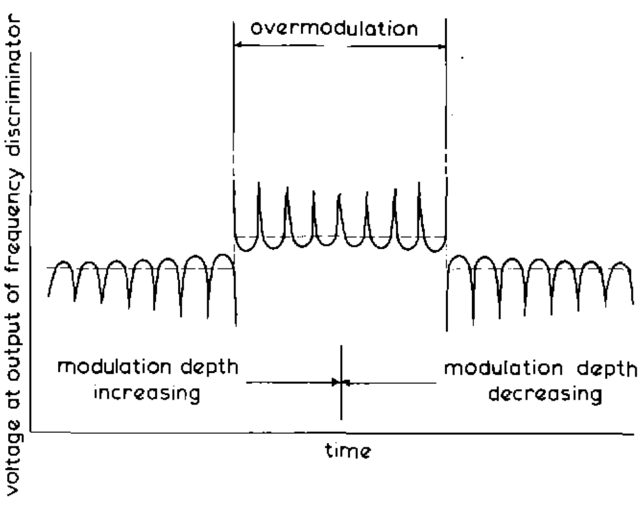

Fig. 4 The effect of overmodulation on the frequency-discriminator output

length of time during which the amplitude of the side component exceeds that of the carrier. In the single-sideband transmission of a complex modulating signal, having a band of side-frequencies, the added pulse during the overmodulation period will, of course, be of irregular shape and its average height will depend on the mean frequency of the sideband components. The pulse does not occur during overmodulation in a double-sideband transmission which has sideband symmetry because the mean frequency of the sideband components is always that of the carrier.

Another limitation of the non-synchronous method described is that it is difficult to implement an integrator which performs well at very low frequencies with the result that the envelope distortion is less effectively corrected for these components.

Symmetrical limiting is somewhat difficult to achieve if the intermediate frequency is high, although in the demodulator described it should be remembered that phase errors which occur due to the limiter deficiencies affect only the correction signal.

Linear multiplication presents no problem with the advent of stable integrated circuits.

5 Experimental Demodulator

An experimental demodulator was constructed to demonstrate the principle of the asynchronous method and help to assess the feasibility of this approach. Fig. 5 shows the circuit used, which is based largely on the schematic arrangement shown in Fig. 2c. It was built up mainly from general-purpose integrated circuits. Briefly, limiting and frequency discrimination are achieved by one integrated circuit (PA. 189) and an external tuned circuit adjusted to resonate in the region of 1 MHz. The filtered output from the discriminator is d.c. coupled to a single p.n.p. transistor stage with capacitive (collector-to-base) feedback which forms the integrator. The four-quadrant linear-multiplier follows closely the design proposed by Gilbert[9] in which if two bipolar-signals X and Y, say, form the inputs, the collector current in one of the output pair of transistors is proportional to $(1-XY)I_E$, where I_E is the common emitter current. Hence if we arrange that $X=Y=\phi(t)/\sqrt{2}$ and also that I_E is proportional to $M(t)$ we obtain the required output current, namely $M(t)(1-\phi(t)^2/2)$. In fact, the transistor (T in Fig. 5) in the emitter circuit was biased both to perform envelope detection and to provide a third input feed to the multiplier, the rectified carrier components being filtered off at the final output.

A diode chain was used to provide the bias feeds to the multiplier transistors. It was found that the integrator began to deteriorate in performance below about 300 Hz.

The demodulator was tested by injecting either single- or double-sideband modulated carriers at 1 MHz by direct feed from low-power modulators. The circuit parameters were adjusted to obtain optimum correction of single-sideband distortion using an audio-tone generator as the modulating source. Sound programmes (both music and speech) were also used for brief listening tests, and it could be demonstrated that reproduction when using the non-synchronous demodulator was almost indistinguishable between single-sideband and double-sideband inputs. The impulse-type noise arising from overmodulation peaks, which can arise if the maximum modulation depth is not strictly limited to less than 100%, was found to be subjectively disturbing.

The intrinsic advantage of the non-synchronous method, over the synchronous approach, is its insensitivity to the frequency of the carrier. In the experimental demodulator the carrier frequency could be varied by at least ±2 kHz without significant change in the demodulated output.

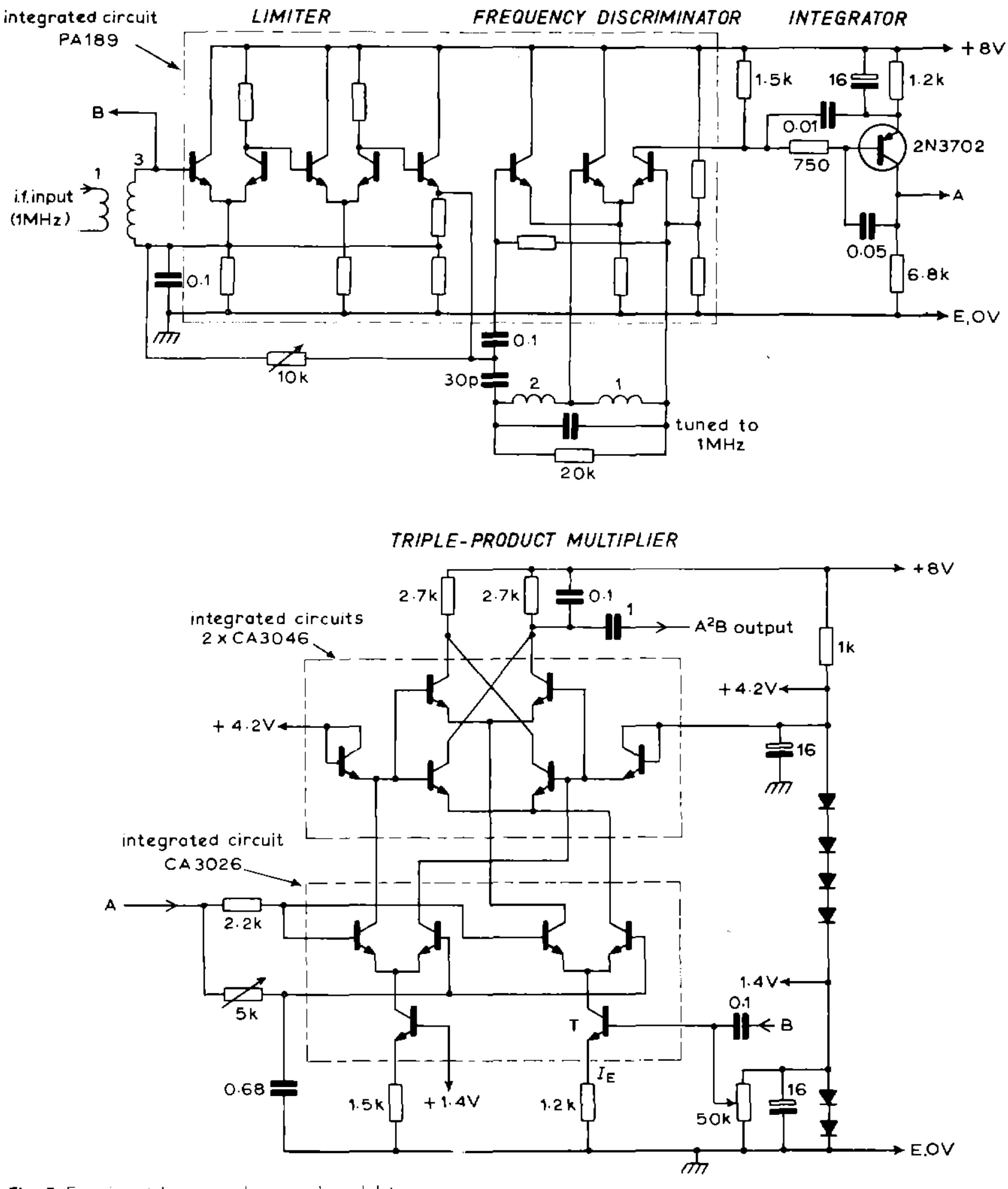

Fig. 5 Experimental non-synchronous demodulator

6 Possible Applications

The new demodulator can be regarded as an improved form of envelope detector because it will substantially eliminate non-linear distortion due to sideband asymmetry. The restricted range of modulation depths (i.e. not exceeding 100%) is also similar to that of a conventional envelope detector except that the limitation on overmodulation is more abruptly defined due to the effects of carrier-phase jumping. Hence the new demodulator could, in theory, be employed to advantage for any communication application which uses a conventional envelope detector in the receiver, providing the transmissions are not likely to significantly violate the modulation-depth limitation when the r.f. signal spectrum is simultaneously asymmetrical.

Consider the potential advantages, for example, if the proposed demodulator were incorporated in a standard transistor portable for m.f. reception of normal double-sideband transmissions. The selectivity of the r.f. plus i.f. circuits of these receivers is typically that shown in Fig. 6a. It is clear

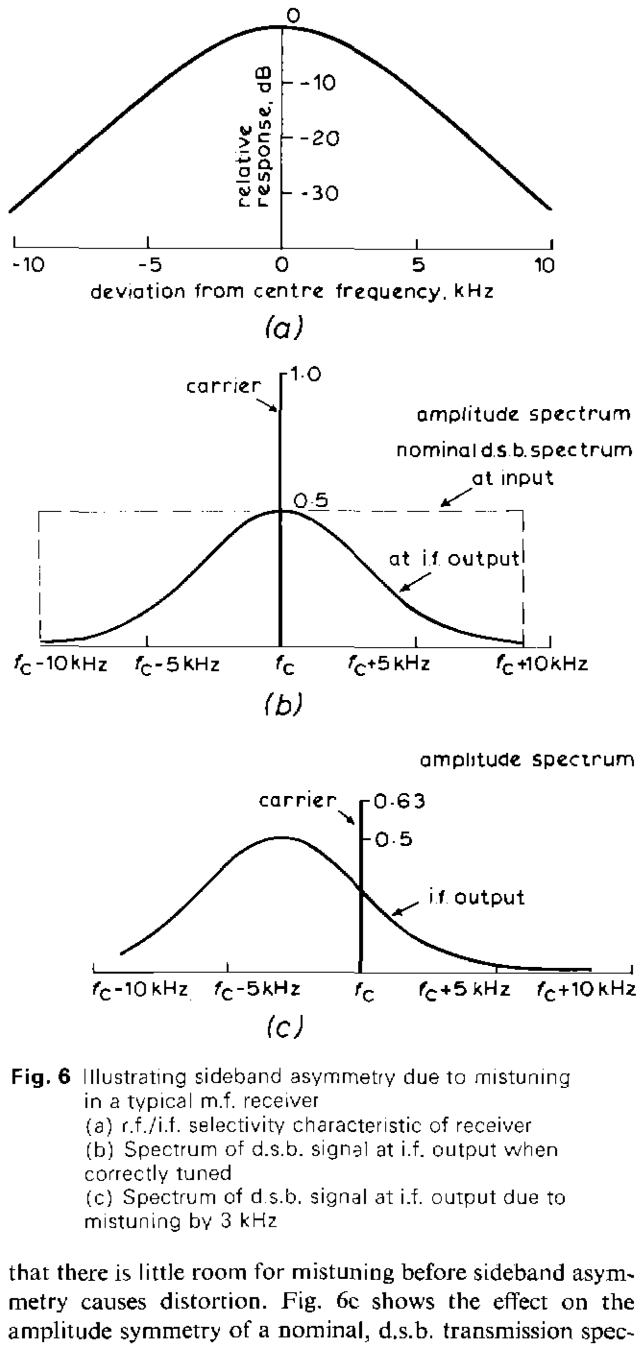

Fig. 6 Illustrating sideband asymmetry due to mistuning in a typical m.f. receiver
(a) r.f./i.f. selectivity characteristic of receiver
(b) Spectrum of d.s.b. signal at i.f. output when correctly tuned
(c) Spectrum of d.s.b. signal at i.f. output due to mistuning by 3 kHz

that there is little room for mistuning before sideband asymmetry causes distortion. Fig. 6c shows the effect on the amplitude symmetry of a nominal, d.s.b. transmission spectrum when the receiver is mistuned by 3 kHz. Not only is the carrier amplitude reduced (by about 4 dB), but the spectrum is severely asymmetrical and, followed by a conventional envelope detector, would probably result in objectionable non-linear distortion. Using the new demodulator this distortion would not arise, so that the receiver would have in effect a substantially greater acceptable tuning range. Moreover, advantage could be taken of this when there is interference from an adjacent channel because one is able to tune away from the interfering station. In the example shown in Fig. 6, approximately 10 dB greater rejection of the adjacent-channel interference would be obtained by the 3 kHz mistuning.

Colour television is another possible field of application. Here saturation errors arise due to quadrature distortion when conventional envelope detection is used. The problem becomes more serious for rebroadcast transmissions because the quadrature distortion occurs at least twice, once in each rebroadcast receiver (there may be two or more rebroadcast links in tandem) and again in the domestic receiver. Synchronous detection is the ideal solution but this requires considerable circuit complexity. Moreover it may fail if the carrier frequency is out of tolerance. The proposed demodulator does not involve expensive components and may be a useful alternative approach. For this application, the video frequencies most in need of correction are centred around the chrominance subcarrier; this eases the design of the integrator in the demodulator. Nevertheless, it may be advantageous to correct over a wider range of video frequencies in order to reduce quadrature distortion associated with luminance detail.

In principle, the demodulator should aid the reception of a.m. transmissions which are subject to selective fading, since any sideband asymmetry introduced will not lead to non-linear distortion, provided that sufficient carrier is maintained at the input to the demodulator to prevent overmodulation.

7 Conclusions

The principle of a new form of demodulator has been demonstrated, and its limitations and possible applications discussed. In order to assess the feasibility of the method more fully, experimental versions of the demodulator might be incorporated in both an a.m. radio receiver and a colour television monitor (or receiver). In the radio application the effects of noise, interference etc. could be determined realistically. In the television application, further development of the demodulator circuitry would be necessary to deal with the higher intermediate frequency. Implementation would, in fact, have been somewhat difficult with components available hitherto, but progress in integrated circuits should make development easier. Consideration should also be given to the possibility of using the proposed demodulator for h.f. rebroadcast receptions of d.s.b. transmissions at overseas transmitters. It could have practical advantages over a synchronous detector, but it would probably require enhancement of the carrier by i.f. selectivity (the resulting loss of high-frequency response being corrected at a.f.) so as to reduce the frequency of occurrence of overmodulation due to selective fading of the carrier.*

8 References

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3. Gilbert, B. 1968. A precise four-quadrant multiplier with sub-nanosecond response. *IEEE Journal of Solid-State Circuits*, 1968, SC—3, 4, pp. 365–73.

* Suggested by Dr G. J. Phillips.

Satellite Broadcasting Service Areas

J. W. Head*

Summary: The coverage area provided by a transmitter on a geostationary satellite emitting a right-circular conical beam is mainly considered. For sufficiently small beamwidth, the service area is nearly an ellipse in the tangent plane to the Earth at the target point, and the axial lengths and directions are calculated. In general, the service-area boundary is best specified by plotting latitude against longitude, and formulæ for specifying this boundary for arbitrary beamwidth are given and discussed. In particular, the 'tulip' shape of some of the *diagrams of Fig. 2 (reproduced from reference 1) is explained.*

Contents

1 Introduction

If a satellite is to be geostationary, it must be above a point on the Equator, and at a distance of about 42 200km from the centre of the Earth. The beam emanating from the satellite will generally be assumed to be a right-circular cone (the case of a general conical beam is discussed in outline in Part 5). This cone will meet the Earth's surface in a curve, as illustrated in Fig. 1. For sufficiently small beamwidth, the curve

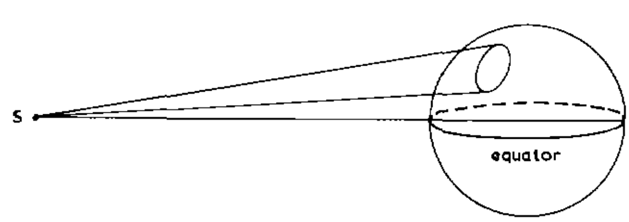

Fig. 1 Illustration of a typical service area (for narrow beamwidth)

is approximately an ellipse in the tangent plane to the Earth at the target point. For larger beamwidth, the curve will cease to lie in one plane, although it will remain a closed curve unless the beamwidth is so great that part of the cone of radiation from the satellite lies outside the tangent cone from the satellite to the surface of the Earth.

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In general, it is more convenient to specify the boundary of the service area by plotting latitude against longitude. Fig. 2 (reproduced from Reference 1) indicates the general nature of such curves.

The object of this article is to provide formulæ by means of which the boundary of the service area can be determined whatever the beamwidth. The notation, based mainly upon the geometry of Fig. 3, is explained in Part 2. The case when the beamwidth is small in considered in Part 3; axial ratios and directions are tabulated for ten-degree intervals of the latitude of the target point, and of the difference between the longitudes of the target point and the subsatellite point.

If latitude is plotted against longitude, it can be shown that the small-beamwidth service-area boundaries remain ellipses, but as the beamwidth increases, they become non-elliptical closed curves until the cone of radiation from the satellite meets the tangent cone from the satellite to the Earth. For still greater beamwidth, we have the 'tulip' shape of some of the curves in Fig. 2. To determine a typical point of such a curve, it is necessary to find first the co-ordinates of a geometrically-specified point Q_n on the boundary of the service area and in the plane through the target point P_1 normal to the line SP_1 joining P_1 to the satellite S. Thereafter we have to determine whether the line SQ_n meets the Earth's surface, and, if so, the co-ordinates of P_n, the point of intersection. Formulæ determining the co-ordinates of Q_n, and the co-ordinates of P_n in terms of those of Q_n, are given in Part 4. Part 6 gives conclusions.

2 Notation and the Geometrical Considerations Underlying it

Most of the points which are relevant to the determination of the service area are in the plane SCP_1 (the plane of the paper in Fig. 3), where S is the satellite, P_1 the target point, and C the centre of the Earth. The co-ordinate axes (not shown in Fig. 3) are such that Cz is the axis of the Earth's rotation, and Cx is the intersection of the Equator and the Greenwich meridian; for a right-handed system of axes, latitude North,

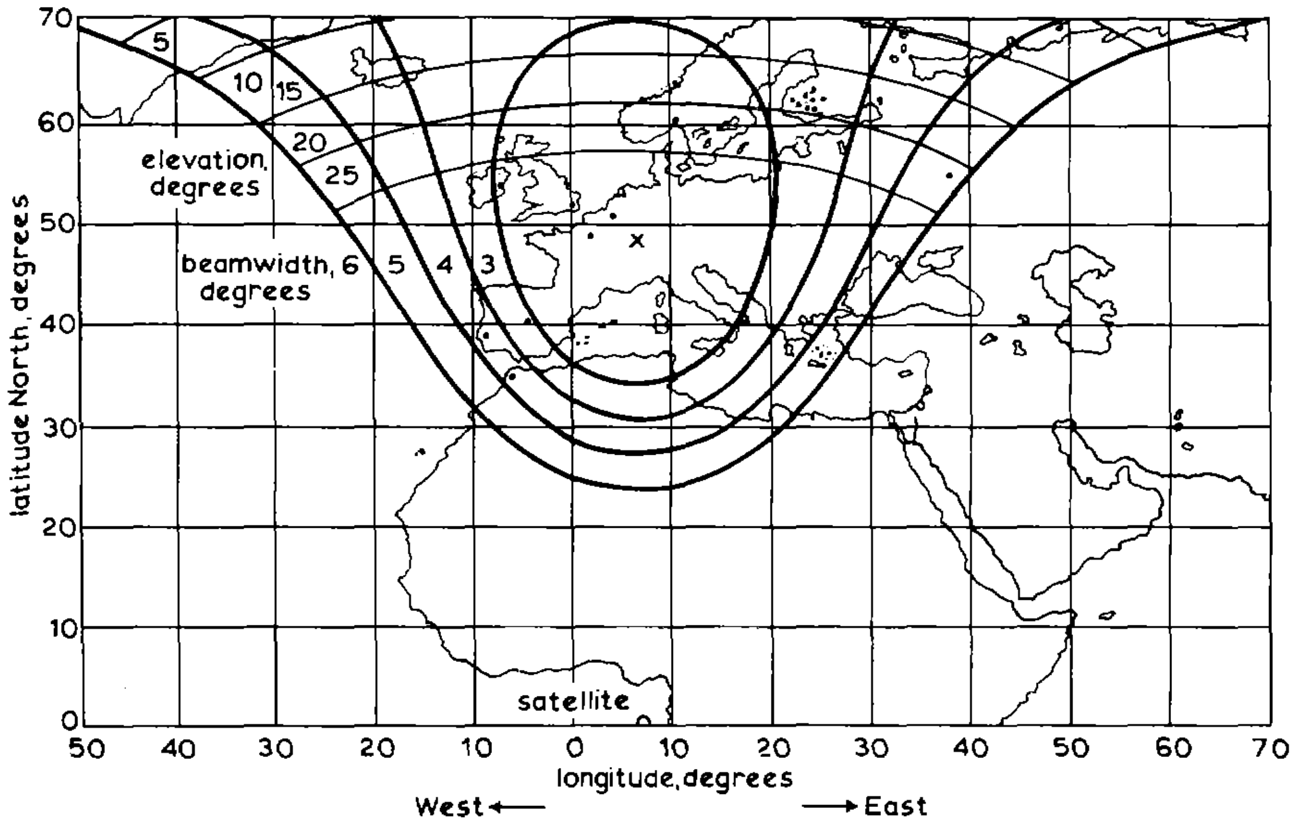

Fig. 2 Coverage of Europe with the satellite at 8°E, a target point near Strasbourg, and varying beamwidth. The scales of latitude and longitude are linear, so that the map is not a standard projection

and longitude East of Greenwich, must be taken as positive. We shall use the following symbols:

C The centre of the Earth (and origin of co-ordinates)
D The distance CS, taken as 26300 miles (42200 km) in numerical work
K_1 The point on the Equator due South of P_1
L_1M_1 Arbitrary points in the plane SCP_1 such that $L_1P_1M_1$ is the tangent to the Earth at P_1 in that plane
P_t The target point
P_n $(n\neq 1)$ A point where the line SQ_n meets the Earth's surface
Q_2, Q_3 Points in the plane SCP_1, in the plane through P_1 normal to SP_1 and on the boundary of the service area
Q_n $(n\neq 1,2,3)$ Any other point on the boundary of the service area and in the plane through P_1 normal to SP_1
r The distance Q_nP_1 $(=SP_1\tan\frac{1}{2}\alpha)$
R The radius of the Earth, taken as 3960 miles (6375 km) in numerical work
s The subsatellite point
S The position of the satellite (co-ordinates $(D\cos\phi_s, D\sin\phi_s, 0)$)
x_n $(=R\cos\theta_n\cos\phi_n)$ x – co-ordinate of P_n
y_n $(=R\cos\theta_n\sin\phi_n)$ y – co-ordinate of P_n
z_n $(=R\sin\theta_n)$ z – co-ordinate of P_n
$\left.\begin{matrix}x_1+X_n\\ y_1+Y_n\\ z_1+Z_n\end{matrix}\right\}$ Co-ordinates of Q_n (See Equations 9, 10, & 12)

α The beamwidth
γ_n The angle between CP_1 produced and P_1Q_n
θ_n The latitude of P_n
ξ_n The angle $Q_nP_1Q_2$
ϕ_n The longitude of P_n
ϕ_s The longitude of s
ψ_1 The angle SCP_1
χ_1 The elevation of S as seen from P_t
ω_1 The bearing (West or East of South) of s from P_1

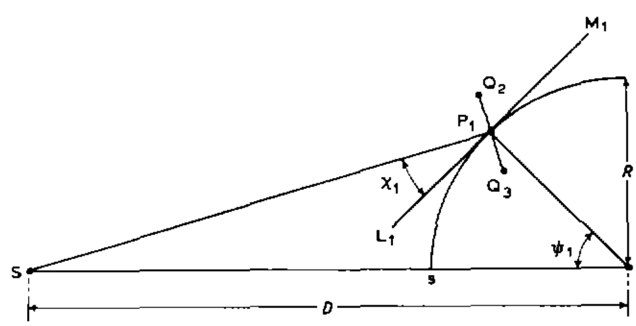

Fig. 3 Relevant points associated with the plane formed by the satellite, the target point, and the centre of the Earth

3 Axial Lengths and Directions for the Narrow-Beam Case

From the geometry of Fig. 3 the following relations can be established:

$$\sin\chi_1 = \frac{D\cos\psi_1 - R}{SP_1}\,;\ \cos\chi_1 = D\sin\psi_1/SP_1 \quad (1)$$

$$\cos\gamma_n = \cos\chi_1 \cos\xi_n = D\sin\psi_1 \cos\xi_n/SP_1 \quad (2)$$

$$SP_1^2 = D^2 + R^2 - 2DR\cos\psi_1 \quad (3)$$

$$\cos\psi_1 = \cos\theta_1 \cos(\phi_1 - \phi_s) \quad (4)$$

and from the spherical triangle P_1sK_1, right-angled at K_1

$$\sin\omega_1 = \sin(\phi_s - \phi_1)/\sin\psi_1 \quad (5)$$

For the narrow-beam case, we can regard the beam from S as a right-circular cylinder of radius r, cut by the tangent plane to the Earth (in which the elliptical service-area boundary lies) at an angle $(\pi/2 - \chi_1)$ with the plane through P_1 normal to SP_1. Hence the major axis of the ellipse is in the direction $L_1P_1M_1$ and has length $r\operatorname{cosec}\chi_1$, and the minor axis (in the direction through P_1 perpendicular to the plane SCP_1) has length r. The ratio $\operatorname{cosec}\chi_1$ of the axial lengths can be deduced directly from the first of (1), and is tabulated in Table 1, while the direction of the major axis is determined by ω_1 in Equation (5) and is tabulated in Table 2.

4 Determination of a Typical Point of the Boundary of the Service Area in the General Case

When the beamwidth is not necessarily small, we seek to determine the latitude θ_n and longitude ϕ_n of a typical point P_n of the boundary of the service area, and for this purpose it is convenient first to find the co-ordinates of the point Q_n in which SP_n meets the plane through P_1 normal to SP_1. We assume that Q_n is specified in terms of r (which is determined by the beamwidth and the position of the target point P_1, and is thus a known constant) and the angle ξ_n or $Q_nP_1Q_2$. Then since $P_1Q_n = r$

$$X_n^2 + Y_n^2 + Z_n^2 = r^2 \quad (6)$$

TABLE 1
Axial Ratio

θ_1 / $\phi_s - \phi_1$	0°	10°	20°	30°	40°	50°	60°	70°	80°
0°	1	1·021	1·090	1·220	1·446	1·850	2·673	5·009	42·68
10°	1·021	1·043	1·113	1·246	1·477	1·891	2·736	5·153	48·21
20°	1·090	1·113	1·188	1·330	1·577	2·022	2·940	5·633	78·25
30°	1·220	1·246	1·330	1·489	1·769	2·277	3·344	6·634	—
40°	1·446	1·477	1·577	1·769	2·108	2·736	4·098	8·718	—
50°	1·850	1·891	2·022	2·277	2·736	3·609	5·633	14·12	—
60°	2·673	2·736	2·940	3·344	4·098	5·633	9·789	48·21	—
70°	5·009	5·153	5·633	6·634	8·718	14·12	48·21	—	—
80°	42·68	48·21	78·25	—	—	—	—	—	—

TABLE 2
Direction of Major Axis. Angle (ω_1) from a North-South Line is Tabulated

θ_1 / $\phi_s - \phi_1$	0°	10°	20°	30°	40°	50°	60°	70°	80°
0°	—	0°	0°	0°	0°	0°	0°	0°	0°
10°	90°	45°27′	27°16½′	19°25½′	15°20′	12°58′	11°30½′	10°38′	10°9′
20°	90°	64°30′	46°47′	36°3′	29°31′	25°25′	22°48′	21°10′	20°17′
30°	90°	73°16′	59°22′	49°6′	41°56′	37°0′	33°41½′	31°34′	—
40°	90°	78°18½′	67°49′	59°13′	52°33′	47°36′	44°6′	41°46′	—
50°	90°	81°42½′	73°59′	67°14½′	61°39½′	57°16′	54°0′	51°45′	—
60°	90°	84°16′	78°50′	73°54′	69°38½′	66°9′	63°26′	61°31′	—
70°	90°	86°22½′	82°54′	79°41′	76°50′	74°25′	72°30′	—	—
80°	90°	88°14′	86°33′	—	—	—	—	—	—

Since P_1Q_n is perpendicular to SP_1,

$$(D\cos\phi_s - R\cos\theta_1\cos\phi_1)X_n + (D\sin\phi_s - R\cos\theta_1\sin\phi_1)Y_n - R\sin\theta_1 Z_n = 0 \qquad (7)$$

and since the angle between CP_1 produced and P_1Q_n is γ_n specified by equation (2), it can also be shown that

$$\cos\theta_1\cos\phi_1 X_n + \cos\theta_1\sin\phi_1 Y_n + \sin\theta_1 Z_n = rD\sin\psi_1\cos\xi_n/SP_1 \qquad (8)$$

In general, (that is, if $\phi_1 \neq \phi_s$) (6), (7), and (8) can be reduced to

$$Z_n^2\sin^2\psi_1 - 2rZ_n\cos\xi_n\sin\theta_1\sin\psi_1[D - R\cos\psi_1]/SP_1 + r^2\{\sin^2\theta_1 - \sin^2\psi_1\sin^2\xi_n - (R^2\sin^2\theta_1\sin^2\psi_1\cos^2\xi_n/SP_1^2)\} = 0 \qquad (9)$$

and X_n, Y_n are then obtained from

$$\left.\begin{aligned} X_n \sin(\phi_1 - \phi_s) &= B\sin\phi_1 - A\sin\phi_s \\ Y_n \sin(\phi_1 - \phi_s) &= A\cos\phi_s - B\cos\phi_1 \end{aligned}\right\} \qquad (10)$$

where

$$\left.\begin{aligned} A &= \{(rD\sin\psi_1\cos\xi_n)/SP_1 - Z_n\sin\theta_1\}/\cos\theta_1 \\ B &= (rR\sin\psi_1\cos\xi_n)/SP_1 \end{aligned}\right\} \qquad (11)$$

Given r, ξ_n [as well as θ_1, $(\phi_1 - \phi_s)$ and hence ψ_1] (9) gives two values of Z_n as a multiple of r, and then (10) and (11) give the corresponding values of X_n and Y_n as multiples of r. Equations (9), (10), and (11) are unaltered if ξ_n is replaced by $-\xi_n$: this explains why a quadratic equation for Z_n is to be expected. If $\phi_1 = \phi_s$, the corresponding solution is explicit, namely

$$\left.\begin{aligned} X_n &= r[\pm\sin\phi_1\sin\xi_n + (R/SP_1)\sin\theta_1\cos\phi_1\cos\xi_n] \\ Y_n &= r[\mp\cos\phi_1\sin\xi_n + (R/SP_1)\sin\theta_1\sin\phi_1\cos\xi_n] \\ Z_n &= r\cos\xi_n(D - R\cos\theta_1)/SP_1 \end{aligned}\right\} \qquad (12)$$

At this stage we are in a position to determine explicitly the co-ordinates of Q_n, which are

$$(x_1 + X_n,\ y_1 + Y_n,\ z_1 + Z_n),$$

when r and ξ_n are given; it remains to determine the latitude and longitude of the point P_n in which SQ_n meets the Earth's surface. It is sufficient to determine the co-ordinates (x_n, y_n, z_n) of P_n; it can be shown that these satisfy the equations

$$\frac{x_n - x_1 - X_n}{D\cos\phi_s - x_1 - X_n} = \frac{y_n - y_1 - Y_n}{D\sin\phi_s - y_1 - Y_n} = \frac{z_1 + Z_n - z_n}{z_1 + Z_n} = \eta, \text{ say} \qquad (13)$$

where

$$\eta^2[SP_1^2 + r^2 - (2rRD\sin\psi_1\cos\xi_n/SP_1)] - 2\eta[DR\cos\psi_1 + (rRD\sin\psi_1\cos\xi_n/SP_1) - R^2 - r^2] + r^2 = 0 \qquad (14)$$

Equation (14) (of which only the smaller root is relevant) only involves ξ_n, r and known quantities, so η can be determined without knowing the position of Q_n explicitly. Once η is known, the last of equations (13) gives z_n and hence θ_n directly. Either of the remaining equations (13) then gives x_n or y_n and hence ϕ_n.

We are therefore now able to plot the latitude θ_n against the longitude ϕ_n in the general case, using a linear scale for both quantities. The geographical significance of the results is best appreciated by appropriately distorting the map of the world in the relevant region, since it is only major geographical features with which we are concerned.

If the beamwidth α (and therefore r) is sufficiently small, the line SQ_2 in Fig. 3 will meet the Earth's surface at a point P_2 (not shown in Fig. 3) on the great circle SP_1 produced, but if the beamwidth is too large, the line SQ_2 will fail to meet the Earth. The critical situation occurs when the line SQ_2P_2 touches the Earth at P_2, so that the angle P_2SC is $\sin^{-1}(R/D)$ or 8°40′. If angle P_2SC exceeds this value, there must be a value of ξ_n such that angle P_nSC is

$$\sin^{-1}(R/D)$$

instead, and the point P_n represents the theoretical extremity of the 'tulip' in such a case. At this point P_n, the elevation of the satellite is zero, so that service, though theoretically possible, will be poor. In practice, the service area should be regarded as terminated when the elevation of the satellite has a minimum value of at least a few degrees: the 'tulip'-shaped curves of Fig. 2 are for this reason terminated before the theoretical extremity is reached.

5 Modification for a Conical Beam of Elliptical Cross-Section

If the beam emitted from the satellite is an arbitrary cone, meeting the plane through P_1 normal to SP_1 in an ellipse with centre P_1, the only difference will be that the distance r to a point on the service-area boundary in the plane through P_1 normal to SP_1 will no longer be constant and equal to $SP_1\tan\frac{1}{2}\alpha$, but will instead be given by a formula of the form

$$\frac{1}{r^2} = \frac{1}{r_1^2}\cos^2(\xi_n - \xi_0) + \frac{1}{r_2^2}\sin^2(\xi_n - \xi_0) \qquad (15)$$

where $r_1 = SP_1\tan\frac{1}{2}\alpha_1$ is the distance to a point of the service-area boundary in the direction of maximum beamwidth α_1, ξ_0 is the value of ξ_n in this direction, and $r_2 = SP_1\tan\frac{1}{2}\alpha_2$ is the distance to a point of the service-area boundary in the perpendicular direction of minimum beamwidth α_2. In equation (15) there are three distinct essential parameters, r_1, r_2, and ξ_0, whereas when $r_1 = r_2$ the value of ξ_0 is immaterial and we only have one parameter.

Qualitatively, we can expect the same general features as before: nearly elliptical service areas for sufficiently small beamwidth, and 'tulip'-shaped service areas for larger beamwidth. But as the orientation of the beam now has to be considered, there is no longer necessarily symmetry of the service area boundaries with respect to the plane SP_1C of Fig. 1, and, as already noted, there are three significant parameters to consider instead of only one. No detailed quantitative investigation will therefore be undertaken here, although all the mathematical machinery for such an investigation is available in the formulæ already derived if a practical requirement should arise.

6 Conclusions

For a right-circular beam of sufficiently small beamwidth, the service-area boundary is approximately an ellipse in the tangent plane to the Earth, centred at the point towards which the beam is directed. The axial lengths and directions of this ellipse are determined by simple formulæ based on the geometry of Fig. 3. For larger beamwidths the service area boundaries have the 'tulip' shape so noticeable in Fig. 2 (taken from Reference 1); formulæ are given for determining the position of an arbitrary point of such a service area, and in particular, for determining the theoretical extremity of the 'tulip'. The case of a conical beam which is not right-circular is qualitatively similar and could be fully investigated by means of the formulæ given here, but has not been attempted because it is qualitatively much more complicated, since three significant parameters are involved instead of one.

7 Reference

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Contributors to this issue

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David Walker came to the BBC in 1961 after studying at Willesden Technical College and working for four years at the GPO Research Station at Dollis Hill. He joined Designs Department as a technical assistant in the Television Transmission Section, and since being promoted engineer in 1964 has worked in a number of other Sections in that department including Special Systems and Television Studios, which was where his interest in film circuits was first aroused.

At present he is a senior engineer in the Sound Transmission Section and is concerned with the design of equipment for the Sound-in-Syncs system.

Neil Gilchrist joined the BBC in 1965 after graduating from Manchester University and worked in Designs Department as a Graduate Trainee. Early in 1967 he moved to Research Department to take up an appointment as an Engineer in Transmission Section.

In addition to the measurement of low-level harmonic signals, he has worked on the design of a number of aerials for radio-camera applications and has been investigating the ability of transmitting feeders to withstand high-voltage pulses of very short duration.

Kenneth Hacking joined the BBC Research Department in 1955, after working in an industrial research laboratory mainly concerned with thin films and vacuum deposition techniques. The early part of his career in Research Department was with Optics Section dealing with the performance of lenses for television, assessment of image quality, and colorimetric analysis in colour television systems.

He was transferred in 1965 to Transmission Section, Radio Group, and worked on prediction problems associated with the propagation of uhf/vhf signals over hills. Later he moved to Radio Frequency Systems Section for a short period to study the possible application of less-conventional modulation systems for m.f. sound broadcasting.

At present he is with the Physics Section of Studio Group.

John Waldegrave Head has been mathematical consultant to the BBC Research Department since 1950. Born in 1912, he was educated at Marlborough and Trinity College, Cambridge. At first he taught mathematics at Canford School and later at Malvern College. Subsequently he worked in the Vibration Department of de Havilland's, at T.R.E. Malvern (now Royal Radar Establishment), and was editor of the *Wireless Engineer* Abstracts at the National Physical Laboratory, Teddington.

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