Founded 1925

Incorporated
by Royal Charter 1961

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

Volume 43 No. 3 March 1973

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

The Bigger the Better?

AS modern technology—transport, communications, printing, etc.—brings the peoples of the world closer together by giving them more opportunities to meet and to exchange views—hopefully for improved understanding—so is there growth of international organizations of all kinds, particularly those of a specialized nature. Although human affairs are diverse in the extreme, there are common factors and needs for most peoples and thus there is currently a trend towards association between like groups on what may be termed occupational bases.

As an instance there has recently been announced a new grouping of European Trade Unions within the International Confederation of Free Trade Unions (ICFTU), i.e. the non-Communist body, which will effectively replace the Trade Union confederations and committees within both the E.E.C. and E.F.T.A. This will comprise 17 national union groups from 14 European contries and aims 'to promote, and represent jointly social, economic and cultural interests of trade unionists at European level' and 'will work for safeguarding and strengthening democracy in Europe'.

Engineers will already be familiar with the world wide and European confederations in their own profession—the World Federation of Engineering Organisations (WFEO) and the Fédération Européenne d'Associations Nationales d'Ingenieurs (FEANI), and similar groupings exist in other parts of the world and for other professions. What may be seen however as an expression of reservation in respect of such multi-disciplinary bodies is reported elsewhere in this issue—a 'Convention of National Electrotechnical Societies of Western Europe'.

There have clearly been valid and usually praiseworthy reasons for most international groupings of the kind referred to above. But good intentions are not enough—efforts put into the operation of these groups will only prosper to the extent that the strengths of the constituent members allow. 'Talking shops' unable to achieve anything concrete will result if there is not a sense of community of purpose. The Convention of Electrotechnical Societies may well enable progress to be made in such directions as technical collaboration in a manner which would be difficult and perhaps even irrelevant within the broader groups such as WFEO or FEANI.

In education and training the different needs of individual countries have led to apparently disparate qualifications, but closer examination often reveals that the 'end-product' is not so different: certainly not to the extent of forcing all countries and disciplines into the same strait-jacket.

Coming nearer home, the Council of Engineering Institutions is itself an example of strength being determined by the common will of its constituent members. Over many fields of activity joint aims and actions are appropriate: but the diversity of disciplines involved in the technical areas calls for much smaller often much looser groupings being made, such as the Standing Committee of Kindred Societies which co-ordinates learned society activities.

There is therefore need for a note of caution while generally welcoming international and inter-disciplinary co-operation. While in some cases unity is strength there is not an automatic argument for 'bigger being better', nor for trying to evolve a 'common' man in terms of education, training or restricted opportunity.

Contributors to this issue*

Dr. S. R. Al-Araji was born in Baghdad and educated at Kadhmia Secondary School, from which he won an Iraq Ministry of Oil Scholarship to Britain in 1963. He entered the University College of Swansea and in due course obtained the B.Sc., M.Sc., and Ph.D. degrees of the University of Wales. His Ph.D. research was concerned with innovations in radio receiver design on which he presented a paper at the 1972 IERE Conference on Radio Receivers and Associated Systems. After completing his doctorate, he participated in the Home Office joint research programme at Swansea, as a post-doctoral Fellow. He has now taken up a University post in Iraq.

Professor W. Gosling (Fellow 1968) has occupied the Chair of Electrical Engineering at the University College of Swansea since 1966 and he has recently been appointed for a three-year term as Vice Principal of the College. A Vice-President of the Institution, fuller notes on Professor Gosling's career appeared in the September 1972 and January/February 1973 issues of the Journal.

Mr. S. P. Babary received the B.Sc. degree in physics and mathematics from the University of the Punjab, Pakistan, in 1965 and the M.Sc., in physics with specialisation in electronics from the University of Karachi in 1967. He then came to England and studied at Loughborough University of Technology for the M.Sc. in Electrical Engineering which he was awarded in 1972. At present Mr. Babary is an Electronics Engineer with Pye Business Communications where he is concerned mainly with industrial electronic equipment.

Dr. G. B. Lockhart has been a Lecturer in the Department of Electrical and Electronic Engineering at the University of Leeds since 1971. (See Journal for December 1972).

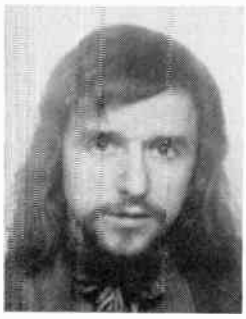
Mr. L. G. Cuthbert graduated in 1969 from Queen Mary College, University of London, with B.Sc. (Eng.) degree in electrical engineering. He then worked on digital systems as a Research Engineer at Standard Telecommunication Laboratories, Harlow, from 1969 to 1970 when he returned to Queen Mary College as a Research Student supported by S.T.L. as well as by the Science Research Council. In October 1972 he was appointed to a Lectureship in the Electrical Engineering Department at the College and he is currently working for a Ph.D. on the design of digital filters.

*See also pages 187 and 208.

Mr. P. Atkinson (Member 1962) graduated with honours in electrical engineering from Imperial College, London in 1955. He was a post-graduate apprentice and had further industrial experience with the guided weapons division of English Electric. From 1959 to 1962 he was a Lecturer in electrical engineering at North Herts Technical College and, from 1962 to 1964, a Senior Lecturer in control engineering at the College of Technology, Letchworth. In 1964, Mr. Atkinson became a Lecturer in control engineering at Reading University, a founder member of the Department of Applied Physical Sciences, and is now a Senior Lecturer. He is the author of two text-books, and many technical papers and holds several patents.

Mr. Atkinson is an active member of the Institution, currently being Chairman of the Thames Valley Section, and he serves on the Examinations Committee and the Automation and Control Systems Group Committee.

Professor P. B. Fellgett (Fellow 1972) occupies the Chair of Cybernetics and Instrument Physics in the Department of Applied Physical Sciences at the University of Reading. He has recently been appointed by the Council to chair the Instrument Technology Group Committee which is in process of formation. A fuller biographical note appeared in the September 1971 issue of the Journal.



Dr. T. G. Swann graduated at the University of Reading in cybernetics in 1966, and gained his Doctorate in control engineering in 1970. He then joined the Computing Research Laboratory of Marconi-Elliott, where he worked on a variety of topics. With the formation of GEC Computers he was employed as both engineer and microprogrammer on processor development, and is now a senior programmer.

Dr. Melapalayam Srinivasan Narasimhan received his degrees in physics from the University of Madras; and he gained M.Tech. and Ph.D. degrees in electronic engineering from the Indian Institute of Technology, Bombay. From 1970 to 1971 Dr. Narasimhan was a CSIR Senior Post-doctoral Research Fellow, working on special types of low noise feeds for large reflector antennas. In the middle of 1971, he joined the Indian Institute of Technology, Madras, as a Lecturer in the Department of Electrical Engineering. Since early 1972 he has been an Assistant Professor, and is head of a group working on research projects in microwave antennas.

UDC 621.396.9: 629.12.014.6

Coupling a position finding system to a marine automatic pilot without the use of an intermediate computer

W. H. P. CANNER,
M.R.I.N., C.Eng., M.I.E.R.E.*

SUMMARY

The coupling together of the Decca Navigator and an automatic pilot has been achieved by the Decca Navigator Company using an Omnitrac computer. The method used is highly complicated and expensive to manufacture. This paper is a feasibility study of the possibility of solving the problem in a more economical fashion for marine applications.

The proposals outlined employ a line following technique and could be adapted to suit any positional finding system using an *x, y, t* plotter as a sensing device to derive a path error signal.

* *Department of Maritime Studies, University of Wales Institute of Science and Technology, Cardiff, CF1 3NU.*

1 Introduction

Rapid advances in recent years have led to the development of several electronic aids to navigation which may, in principle, be fitted to any type of moving vehicle. The most advanced systems have appeared in aircraft and ships, and in more recent years, hovercraft have received a specialized approach resulting in the development of reliable aids designed specifically to suit the peculiarities inherent in these vehicles. Some use is also made of electronic aids in the navigation of land vehicles over desert country.

Broadly speaking, it would be fair to regard electronic navigational aids as falling into one of two categories. The first deals with locating craft position, whilst the second concerns itself with aiding the control of the craft. In the second category we have automatic steering, and it would be reasonable to regard these aids as being pilot aids. In the first category, any system which provides information of heading, speed, and more specifically position may be regarded as being a navigator's aid.

Considerable research has been carried out in coupling positional information to automatic pilots, with varied success in application. Probably, the most noteworthy has been in the automatic approach and landing of aircraft, where precise positional information in both height and locality has been geared to the automatic pilot to enable accuracies greater than those achieved in manual pilotage to be flown—coupled with reliability figures certainly comparable and often claimed to be better than those achieved in manual flying. Path guidance in level flight may also be carried out (e.g. the Smith's Flight System coupled to the Electric Pilot).

It will be appreciated that much of the labour of navigation can be removed both in the air and at sea when positional information is fed into an automatic steering device to provide path guidance. This will allow more time for the navigator to pursue his responsibilities of collision avoidance, weather observation, fuel calculations, and the general work associated with navigation. Today, as the density of traffic increases, collision avoidance is becoming more paramount.

In the marine field, the Decca Navigator has been coupled to the automatic pilot by the addition of an interface unit known as an Omnitrac computer. Unfortunately, due to its cost, many ship-owners have proved reluctant to install this system.

Finally, a word about Hovercraft. These vehicles have inherited the worst features of both sea and air navigation. They are susceptible to tide when waterborne in the terminal areas, and since this involves inshore navigation, there are added hazards from land promontories. Once airborne they are affected by wind to such an extent that the angular difference between course and track can be as much as 45°. This is due to the slow speed of operation compared to a normal aircraft flying in excess of 200 knots. They are also heavily involved in collision avoidance, since the route may lie at right angles to the shipping traffic when operating a ferry service. Economical coupling of a navigational aid and an automatic pilot could well prove particularly useful in the solution of this problem.

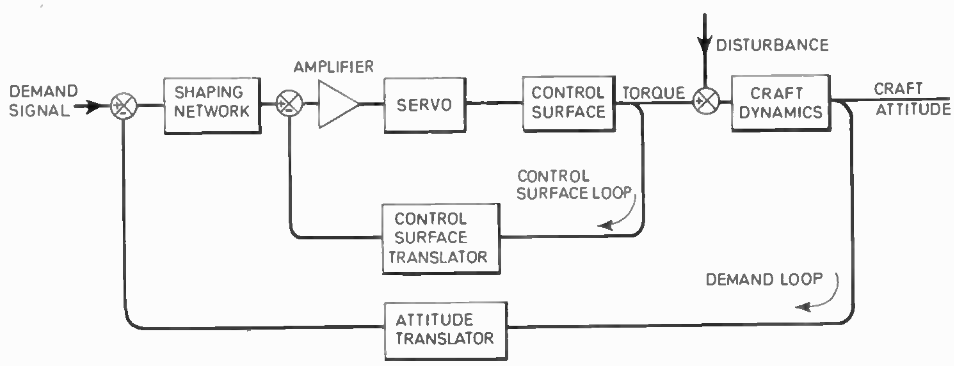


Fig. 1. Block diagram of the automatic control system used.

The author wishes to make it clear that the paper has been based on a Decca Navigator marine system and detailed investigations were originally carried out with this aid. Where references are made in the text to aircraft application they are intended in general terms only, and do not imply that the specific design is necessarily suited for in-flight control.

2 Underlying Principles

2.1 *The Decca Navigator*

The Decca Navigator is a hyperbolic system which provides three single position lines on a three-dial read-out display. Each position line represents a line along which the observer may lie, and since at least two of these position lines are provided continuously and simultaneously, the equipment provides a continuous method of fixing the observer's position. The method used to obtain each position line is essentially by phase comparison of two coherent sources of radiation of unmodulated continuous radio frequency energy in the frequency range of 70–130 kHz. The general principles of this system are well known† and will not be described here, but the author would like to lay emphasis on the visual presentation of position provided by the automatic plotting devices—namely, the flight log and the marine plotter.

In both of these devices, Decca information is translated into related movements of a roller-mounted chart, and a plotting pen, along axes lying at right angles to one another. The pen indicates the position of the craft in relation to the lattice and traces a record of craft movement. In the case of aircraft, the display head is designed for cockpit mounting and may if desired be fitted flush into the instrument panel. In ships, the display head is larger and free standing and would appear to be adaptable to the suggested application in this paper more readily than the flight log.

2.2 *Automatic Pilots*

In automatic pilots for ships or aircraft, a disturbance of craft equilibrium creates a signal which is used to initiate a chain of events that results in a control surface being displaced at a rate compatible with stability. This change in position of the control surface drives the craft back to the original attitude from which it has been disturbed. Craft dynamics alone will provide a degree of stability, but only if there is sufficient feedback of the right type will it be possible to create a perfectly stable craft able to maintain its state of equilibrium regardless of the disturbances applied to it. Apart from disturbances of this nature, however, a positive demand in certain circumstances is also required to initiate a positive change in attitude. In order to achieve this overall requirement, more than one loop will be required. Figure 1 illustrates the general case.

Consider the action starting with a positive demand signal (left of diagram) fed in to change the attitude of the craft. The demand signal passes through a suitable shaping network and thence to the amplifier. The amplifier drives the servo mechanism which is coupled to the control surface. The control surface moves, and in doing so generates a feedback signal produced by the control surface translator. (Simply, the translator could be a potentiometer with its wiper arm connected to the control surface or a tacho-generator giving an output voltage proportional to the rate of change of position of the control surface.) The feedback signal is used to back off—or damp down—the positive demand at the amplifier. The effect of this is to reduce the rate of displacement of the control surface so that the craft gradually, rather than suddenly, changes its position. It is sometimes an advantage to provide manual control over the degree of feedback in order to cope with variation in craft loading. The value of feedback should however be such as to ensure stability. The concept of $A' = A/(1-\beta A)$ for the gain of the loop will be familiar from feedback amplifier design. Here A is the feedforward path of the amplifier, servo, and control surface, whilst β is the feedback path of the control surface translator. Since β is negative, the overall gain A' will be $A/(1+\beta A)$, or the ratio of output torque to input signal will be $F_F/(1+F_F.F_B)$. Clearly, the more feedback in the system the smaller this ratio becomes.

The new position taken up by the control surface creates a torque which causes the craft to move at a rate governed by craft dynamics. As the attitude changes, it approaches the desired attitude set in by the demand signal, and feedback is continuously provided by the attitude translator in the demand loop. This effectively reduces the demand as the desired attitude is approached, until the input to the shaping network becomes zero when the attitude translator signal is equal to the attitude demand signal. Such feedback can be over-ridden if the

† O'Brien, W. J., 'Radio navigational aids', *J. Brit. Instn Radio Engrs*, 7, p. 215, 1947.

operator simply wishes to change the attitude continuously (i.e. in the case of a heading change he may wish to move in a circular path), but in many systems where the demand is provided by a sensor pre-set to some desired value the relationship between the desired attitude and the indicated attitude will be readily available, and will be used to govern craft motion.

Consider now the effect of an external disturbance attempting to upset the equilibrium of the craft. The torque so produced is again resisted by the craft dynamics, but any change in attitude which may result feeds a signal from the attitude translator to the input of the amplifier as a negative going demand. This moves the control surface in such a direction as to oppose the torque producing it. The control surface loop and the demand loop together in a well-designed system should ensure a stable craft which will ultimately take up an attitude able to counteract the disturbing force applied to it.

3 Proposed Method of Coupling

3.1 *Deriving the Path Error Signal*

The heading demand signal to an automatic pilot is derived from a gyro compass repeater. The relative position of the heading indicator and a manually adjusted pointer on the face of the repeater is used to derive d.c. voltage of some value related linearly to the angular difference between them. This signal is amplified and follows the process of feedback autopilots in general as described in Sect. 2.2.

Basically, to provide path guidance, a similar signal derived from a sensor which compares the craft's true position with a desired route, could be used to give a heading demand signal capable of bringing the craft onto the desired route and allowing it to take up a heading to maintain that route. The problem lies in relating the position of the craft to the desired route and using this signal in the correct sense as the path error signal.

The true position of the craft is frequently given by the Decca Navigator, and in aircraft the flight log is generally carried. The Decca marine automatic plotter, as the marine counterpart to the flight log, is sometimes used in ships. These devices indicate the position of the craft by the movement of a pen over the surface of a roll of paper which is overprinted with the Decca lattice.

It is proposed to use the automatic plotter as the path sensor, and to achieve this the required route should be laid down over the paper in the form of a narrow conducting strip of zero resistance accurately inserted for specific routes. The conducting surface so presented on the paper can then be used as the I-bar of an E- and I-signal generator. The E-transformer should be small and light in weight and should be mounted on the arm which carries the pen. It could be an addition to the pen if, at the same time, it is required to use the automatic plotter in its prime use, or in place of the pen, if it is felt that the E-transformer would obscure vision.

The idea behind the derivation of the path error signal is shown in Fig. 2.

With the pen, and/or its attached E-transformer, over the centre of the conducting strip forming the I-bar, the output from the secondary coils of the E-transformer will be zero. The secondary coils are wound on the outer limb of the E-piece and are in series opposition (Fig. 2(a)), so that equal amplitude signals are taken from both sides in antiphase. The net output is therefore zero and the flux linkages on both sides are equal.

Any deviation of the I-bar from the E-transformer, or vice versa, results in a fall in the amplitude of the output from the secondary coil with the weaker flux linkage. Should the E-transformer move to the left of the I-strip, the flux deteriorates in the left-hand magnetic circuit since the air gap has now widened on that side. The net output signal will be of a particular phase and amplitude dependent on the direction (governing the phase), and the distance (governing the amplitude) of the movement. It will be appreciated that the maximum amplitude is reached when the I-strip is completely displaced from one of the secondary coils (Fig. 2(b)). Should there be a deviation of the I-bar (path) wholly to one side of both secondary coils then the output will again fall to zero. This is important, since it will govern the maximum permissible change in the direction of the route. It will not be possible in this case to alter course by 90°.

A large deviation of the E-transformer from the I-strip could be caused by two factors. The first is a change in the environmental conditions surrounding the vehicle which may cause a change at a rate which cannot be compensated in time by the autopilot amplifier to prevent the E-transformer from moving completely away from

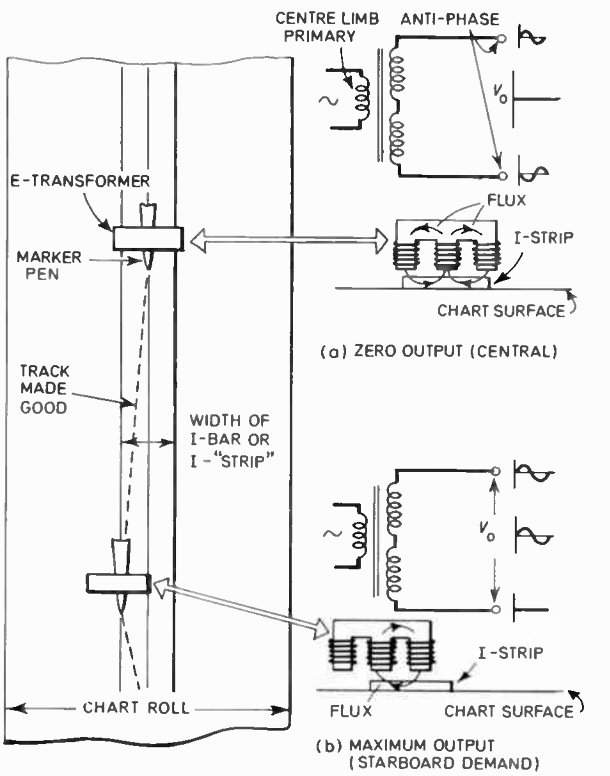


Fig. 2. Derivation of the path error signal.

the I-strip. Experimentation would reveal whether or not this is likely to occur, but if it should do so, it is suggested that it may be overcome either by widening the I-strip (within the limits of an accompanying loss of accuracy), or by using a tacho-generator attached to the shaft of the heading repeater. Such a generator will produce an output in proportion to the rate of change taking place. If this is fed to the autopilot amplifier only when its value rises above a predetermined level, then it should compensate for the fall in output voltage of the E-transformer.

The second, and more likely cause, is a change in the projected direction of the track as laid down by the I-strip which is too large to cope with. (See Fig. 3.)

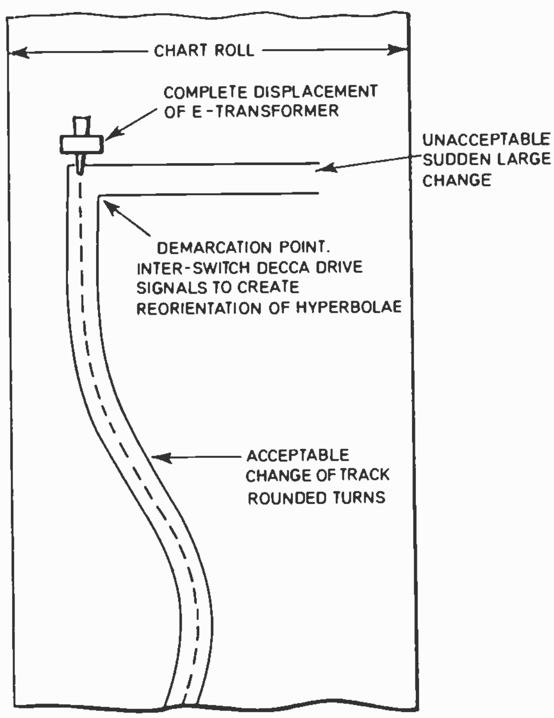


Fig. 3. Unacceptable changes in route.

The output from the E-transformer will deteriorate for changes in direction of the I-strip from 'head up' greater than 45°, so that the ideal solution is to ensure that this angular change is never exceeded. This is not to say that changes in track should not exceed 45°—a straight line projection within these limits for all conditions could be achieved by a 90° swing of the hyperbolae on the chart to ensure a plus or minus 45° head-up presentation at all times. This will necessitate switching the Decca input signals from the chart roller to the pen carriage and from the pen carriage to the chart roller as the demarcation point is reached. A pulsed output signal, used as a trigger automatically to initiate the change over, could be obtained from a second pick-up on the edge of the chart as discussed later in Sect. 3.5.

In general, whenever changes in route of angles less than 45° are involved, the I-strip should be laid down with rounded alterations to act within the limits of the vehicle's manoeuvrability. This will reduce the possibility of complete displacement of the E-transformer from the I-strip.

The output signal derived from the E-transformer may now be referred to as the path error signal. This should be amplified and subsequently fed to a phase-conscious rectifier (p.c.r.) where it would be converted to a d.c. output of polarity dependent on the direction of movement of the I-strip from the E-transformer. The amplitude of the d.c. signal will vary and will depend on the degree of displacement subject to the limitations mentioned. The width of the I-strip would be governed by the accuracy required.

For an aircraft flight system—or indeed a maritime system—a transducer which recognizes the relationship between the track made good and a desired track on an automatic plotter could be effectively designed which is not necessarily based on the E- and I-transformer proposed in this paper. The author has in fact been lately advised that the Decca Navigator Company now hold a patent for doing this in a maritime application which is based on an optical method. Another possibility would be to investigate the use of a capacitance bridge.

The d.c. signal produced by the p.c.r. provides the heading demand to reach the desired path, and can be switched into the autopilot amplifier directly in place of the signal which might otherwise be derived from the gyro or magnetic heading sensor.

To ensure stability, reference should also be continuously made to the rate of change of heading, and this can be achieved by a rate of turn gyro and integrator. The gyro rate signal is then used to back-off the heading demand created by the path error signal. The integrator feeding directly into the servo amplifier can be given a requisite amount of gain which will result in the craft gradually aligning with the track and ultimately allowing it to take up a suitable drift angle to compensate for the environmental conditions with zero tracking error as shown in Sect. 3.4.

3.2 *The Overall System*

Figure 4 illustrates the underlying principle of the complete system. Two modes of operation are envisaged, the first is the gyro mode and the second the path mode. (Switch at centre of diagram.)

An analysis of the action in the gyro mode follows closely on that given under Sect. 2.2 for a control surface in general. In this particular case however, the input signal is a heading demand (top of diagram). The heading translator gives the indicated heading so that the heading error, being a combination of these two, could be derived from the relative settings of the desired heading and an indicated heading on a gyro compass repeater as is normal. The heading error signal passes to the shaping network, and after amplification, feeds the rudder servo system. (In the case of aircraft, the required control over heading is also by the aileron servo mechanisms.) Two closed loops are used in the general system—the rudder loop providing feedback via the rudder angle translator, and the heading loop providing feedback via the heading translator to back-off the input demand. In the event of an external disturbance of equi-

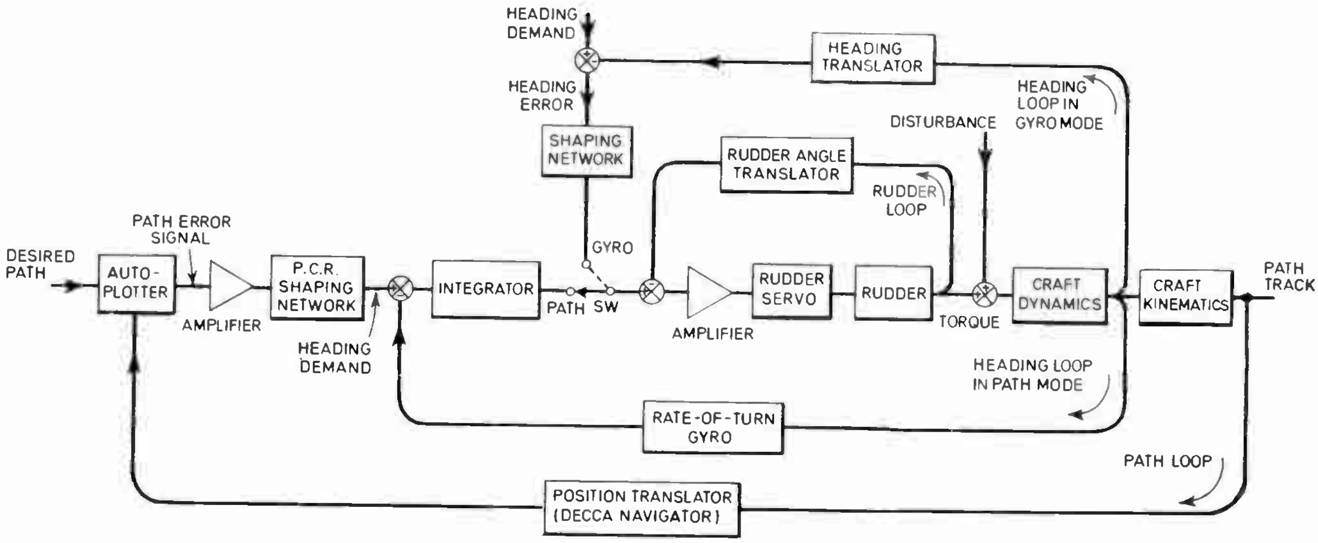


Fig. 4. Block diagram of the overall system.

librium (above right of diagram), craft dynamics provides a natural degree of feedback, but a negative heading error signal is given by the heading translator to drive the rudder (or ailerons) in such a direction as to oppose the disturbance. The two loops should operate to ensure a stable craft under all normal conditions.

If the system is switched to the path mode, there will now be three closed loops. The rudder loop remains the same as in the gyro mode. This adjusts the change of rudder angle to be compatible with the demand as in the general case. The heading loop however is now taken from a rate-of-turn gyro. If the head swings too rapidly, the rate gyro produces a strong feedback signal to back off the heading demand provided by the p.c.r. It would be possible to use the same heading loop as that used in the gyro mode, provided the gyro compass repeater was set for the desired track and continuously monitored for track changes. The suggestion of using a rate gyro is probably more flexible, and better suited to the system.

It has been suggested that the heading demand signal provided by the p.c.r. may be derived from the Decca automatic plotter. This will then have two inputs—the desired path in the form of the I-strip running through the paper, and the pen (E-transformer) position provided by the Decca Navigator. The third loop is therefore the path loop with feedback provided via the Decca Navigator—i.e. a position translator—and regulated by craft kinematics. The output from the automatic plotter as a combination of these two is the path error signal which feeds the amplifier and p.c.r., whilst the output from the p.c.r. is a heading demand to make good the required route.

It is interesting to note that the Decca Navigator does not necessarily have to be the position translator. A Doppler system, or any navigational aid capable of giving accurate position continuously could be used to feed a plotter based on the more conventional x, y, t plotter. The key lies in relating the known position to a desired path, and using this in the right sense to give the path error signal.

3.3 *The Path-error Amplifier and Phase-conscious Rectifier*

Either a class-A or a class-B amplifier in push-pull would be suitable to amplify the path-error signal. This should be coupled to the E-transformer by using a transformer of suitable turns ratio to obtain optimum matching.

The circuit diagram of Fig. 5 indicates a possible coupling arrangement. In this case a class-B amplifier in push-pull is used employing two n-p-n transistors.† Resistors RV1 and R_Z determine the forward bias preventing cross-over distortion which may otherwise be present. The emitter resistor R_e is required to improve thermal stability.

The main advantage of class-B operation, is that there is low quiescent current, and a higher degree of efficiency at full output can be obtained. It may be argued that the low quiescent current factor is unimportant since battery operation will not be used, but nevertheless there is no point in drawing current with zero input signal. The output which the class-B push-pull pair produces has a maximum theoretical efficiency of 78·5%. This refers to the a.c. output power to the d.c. input power. In practice the efficiency would probably lie in the region of 70 to 75%.

The necessity for forward biasing the transistors lies in the prevention of cross-over distortion. This occurs if the change over in current drawn from one transistor to the other is not smooth. The combined transfer characteristics of the two transistors should be a straight line. If this is not so there will be a degree of intermodulation distortion which could result in some variation of d.c. output from the phase-conscious rectifier. There is one other point with regard to cross-over distortion, and this refers to the possibility that it may reappear with extremes of temperature. This could be overcome with the use of a split load circuit.

† 'Manual of Transistor Circuits', Mullard Ltd., Mullard House, London, W.C.1.

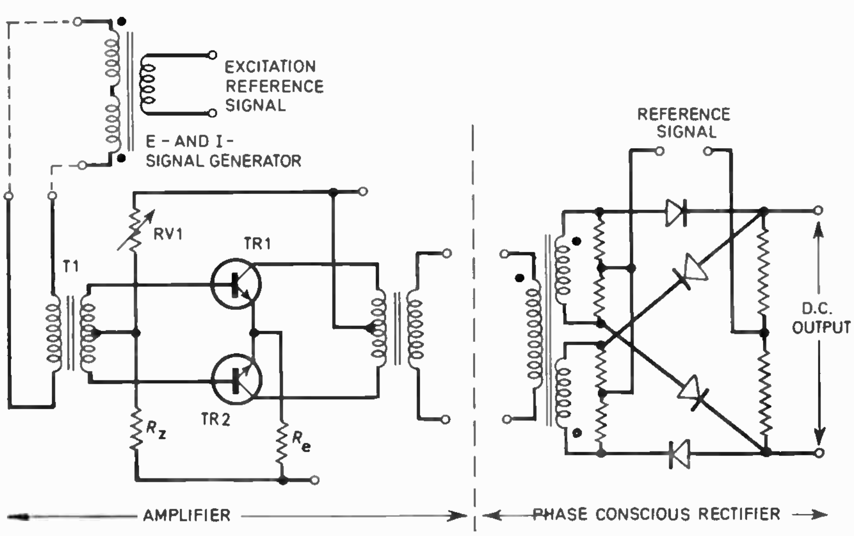


Fig. 5.
The amplifier and the phase conscious rectifier.

The amplifier output is coupled to the phase-conscious rectifier by transformer coupling of suitable turns ratio. The p.c.r. envisaged here could be a full-wave rectifier employing four diodes as indicated in Fig. 5.

3.4 *The Rate Gyro and Integrator*

A rate gyro is a gyro with a single gimbal ring restrained by spring tension, so that it has only two planes of freedom—freedom for the rotor to spin on its axis, and freedom for the axis to turn in the horizontal plane. The spring tension restrains the axis from moving in the vertical plane. The axis thus lies horizontally inside the single gimbal ring which may be held firm by magnetic or spring tension. Any force attempting to change the horizontal angle in which the gyro axis is pointing creates a precessional force acting on the gimbal ring causing it to move and distend the springs (This spring movement itself creates a force which when precessed through 90° causes the axis to move in the direction of the original force applied to it.) The rate of application of the original force will govern the total angular vertical movement of the gimbal ring.

A suitable transducer, frequently an E- and I-transformer, converts this angular movement into an a.c. signal with its phase governed by the direction of change, and amplitude governed by the rate of change. This signal can then be fed into an integrator along with the path error heading signal previously produced by the automatic plotter, so that the rate of change of heading combining in the integrator with the path error heading signal feeds directly into the heading demand amplifier of the autopilot. The effect of this would be to remove an offset value, or steady state error, created by a prevailing environmental force driving the crafts head off to one side. An example of this would be a wind velocity in the case of aircraft, or a tidal effect in ships, driving the head say 5° to starboard. Automatic rudder is applied until the head returns to its correct value, but once this is reached the correcting force is removed. As a result the head once more swings off to starboard with the result that a continuous oscillation takes place offset to one side. This can be removed by integral compensation.

3.5 *Automatic Reorientation*

The Decca marine automatic track plotter in ships has a manual pattern selector switch which enables a display with ship's heading upmost presented at all times. The operation of this switch changes the direction of rotation of the servo motors which drive the pen carriage and paper roller. Under normal operating conditions the pen moves from left to right for an increasing reading of the Decca lattice readings, and is referred to as a positive drive. The paper in the same condition is said to have a positive drive when readings increase from the bottom of the paper upwards.

The method of switching involves the use of a wafer switch which in itself would not be easy to reproduce directly for automatic operation, and since automatic re-orientation would be necessary to facilitate large changes of heading an alternative solution was looked for which would be less involved.

The diagram shown in Fig. 6 illustrates the general principle of operation of the track plotter.† In a particular case we may assume that the output from the receiver is from Master/Red to drive the pen, and from Master/Green to drive the roller. The principle of each is the same so in considering the operation of the plotter, only one input will be considered, i.e. the pen drive obtained from the Master/Red.

The signal arriving from the receiver is in the form of a sine and cosine value. This is fed to a d.c. reversal circuit in order to provide four output signals of sine, –sine, cosine and –cosine value. These are then fed to a ring resistor at four 90° displaced tapping points producing a rotating potential diagram having two null points. As the ship moves over the lattice the null points will rotate with the changing phase of the incoming signal. By using an appropriately graded track on the ring resistor a linear relationship between change of phase angle and displacement of the null positions will be obtained. Mounted on the ring resistor is a movable wiper arm which feeds the servo amplifier and motor

† 'Handbook for Flight Log', Decca Navigator Co., Ltd,, New Malden, Surrey.

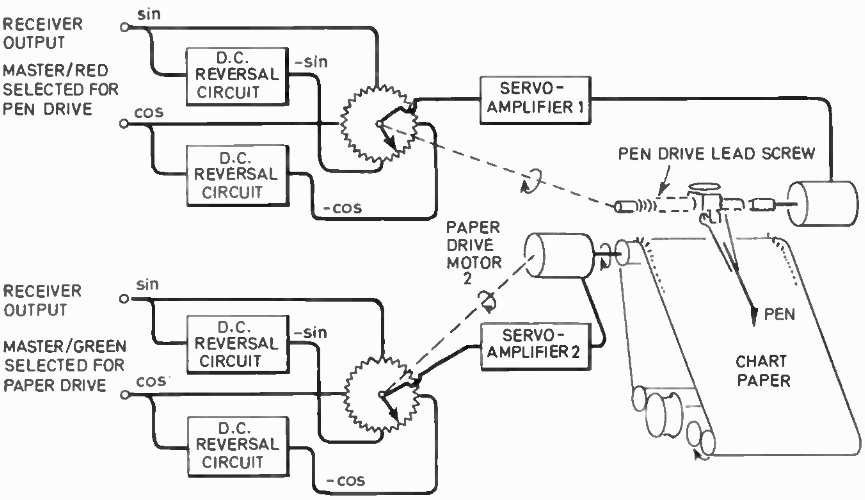


Fig. 6. Principle of the automatic plotter.

driving the pen carriage. The pen drive is also geared to the wiper arm so that as the motor is driven the wiper arm rotates until it is in the null position. Subsequent movement of the ship is then going to result in a movement of the null position and automatic follow-up results.

Before considering any suggestions for automatic re-orientation, let us first examine the movement of the servo motors for a craft moving on an imaginary track which follows a rectangular path as shown in Fig. 7. It can be seen from the Figure that the drive for each of the headings is considered positive simply because readings increase from left to right and from bottom to top. The servo motors will change direction naturally and not through any change brought about by switching at each corner.

In order, however, for the route to be presented as head up throughout, switching action would need to take place at each of the corners in the sequence A, B, C, and D. The turns are rounded to facilitate craft's manoeuvrability.

On the Northerly run from A to B, the pen is driven by Master/Red feeding servo motor 1, whilst the paper is driven by Master/Green feeding servo motor 2. Positive drive is given to both servo motors.

At position B, the system should be switched so that the pen is driven by Master/Green and the roller by Master/Red. This means that the drive to the pen has to be interswitched with the drive to the roller. Secondly, the pen drive now has to be made negative since readings decrease from left to right. This means the servo motor of the pen drive has to be reversed in direction.

At position C, switching action is again required. Once more, interswitching needs to take place between pen and roller. Further, the pen drive has to be negative and so has to be the roller drive.

At position D, interswitching is again necessary, but in this case the pen drive is positive whilst the roller drive is negative.

At position A, the final interswitching is required with both drives back to positive. A tabled analysis of the switching necessary is shown in Table 1. It can be seen that for any situation only two actions are required. The first is interswitching, whilst the second is to reverse the phase of the requisite servo motor drive.

The process of achieving this automatically could be carried out by the use of bistable multivibrators. They would need to be triggered by pulses derived at the right instant from the edge of the chart and picked up by a stationary contact (or set of contacts) bearing against the surface of the paper.

3.6 *Interswitching the Drive Signals*

The feed to the servomotors is normally provided by the power coil of a magnetic amplifier which forms the basis of the servo amplifier. It would be possible to interswitch these connexions by using two bistable multivibrators in the manner shown in Fig. 8. The transistors have only two conditions under which they are operating in this type of circuit—either they are conducting fully,

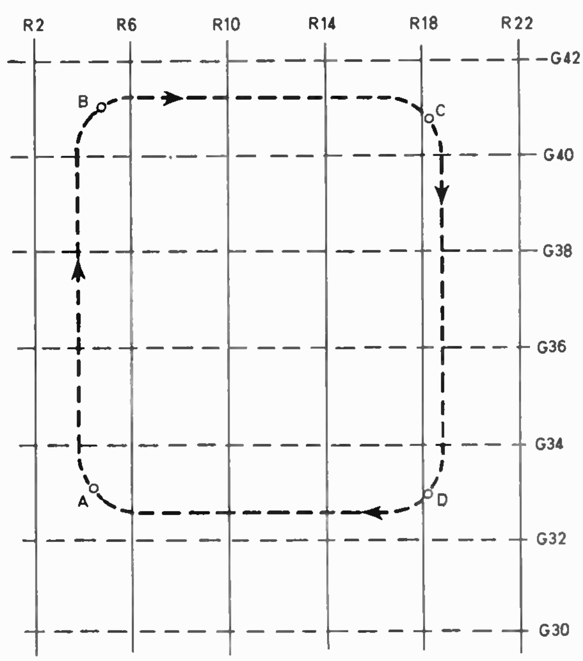


Fig. 7. Rectangular track.

POSITION	PEN DRIVE	PAPER DRIVE	REQUIREMENT	SWITCHING ACTION
A	MASTER/ RED +	MASTER/ GREEN +	NORMAL	R–A–1 + PEN R–A–2 + PAPER
B	MASTER/ GREEN −	MASTER/ RED +	INTERSWITCH DRIVES CHANGE DIRECTION OF SERVO 1	R–A–1 − PEN R–A–2 + PAPER
C	MASTER/ RED −	MASTER/ GREEN −	INTERSWITCH DRIVES CHANGE DIRECTION OF SERVO 2	R–A–1 − PEN R–A–2 − PAPER
D	MASTER/ GREEN +	MASTER/ RED −	INTERSWITCH DRIVES CHANGE DIRECTION OF SERVO 1	R–A–1 + PEN R–A–2 − PAPER
A	MASTER/ RED +	MASTER/ GREEN +	INTERSWITCH DRIVES CHANGE DIRECTION OF SERVO 2	R–A–1 + PEN R–A–2 + PAPER

Table 1.
Analysis of switching required.

or they are not conducting at all. Using a switch analogy it means that they are either 'open' or 'closed'. In the circuit diagram the inter-transistor coupling and steering diodes required, which would follow standard practice, have been omitted for simplicity.

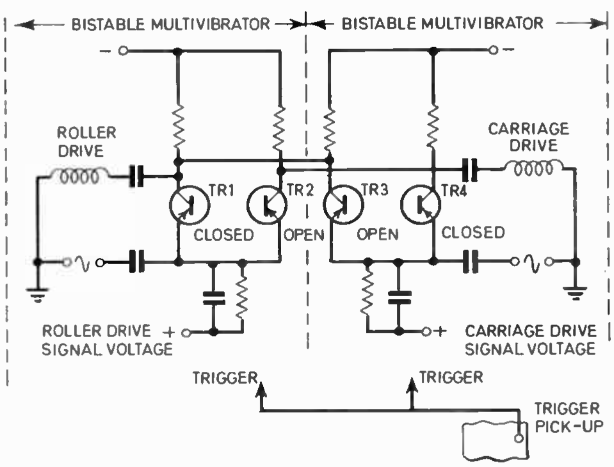


Fig. 8. Interswitching the drive signals.

Four transistors are involved. Assume that TR1 is closed, TR2 open, TR3 open, and TR4 closed with bistables in the quiescent condition. In this state, the 'paper drive voltage' taken from the magnetic amplifier has a closed path to the 'paper servo motor' via TR1. Similarly, the 'pen drive signal' voltage has a closed path to the 'pen servo motor' via TR4.

If a trigger, derived from the 'paper' at the requisite turning point is then fed via steering diodes to all four transistors then both bistables will flop into the opposite state resulting in TR1 opening, TR2 closing, TR3 closing, and TR4 opening. In this condition the 'paper (roller) drive signal' voltage is fed to the 'pen (carriage) servo motor' via TR2, and the 'pen (carriage) drive voltage' is fed to the 'paper servo motor' via TR3.

The transistors would need to be capable of carrying sufficient current, and should be biased sufficiently to prevent the a.c. input from changing the state of conduction.

3.7 *Reversing the Motor Direction*

A similar arrangement also using two bistable multivibrators could be used to reverse the phase of either the driving voltage to the servo, or the reference voltage required by the servo motor.

In Fig. 9, TR1 and TR2 belong to one bistable multivibrator, and TR3, TR4 belong to a second bistable multivibrator. The associated circuitry and trigger input have again been omitted for simplicity.

In the condition shown, TR1 is open, TR2 is closed, TR3 open, and TR4 closed. The supply voltage is fed to the base of the transistors, and can therefore reach the motor winding via the base-collector path of TR2 and TR4. On receiving a trigger pulse however, the state changes so that TR1 closes, TR2 opens, TR3 closes and TR4 opens. In this case the supply reaches the motor winding via TR1 and TR3 and is therefore in anti-phase to the original.

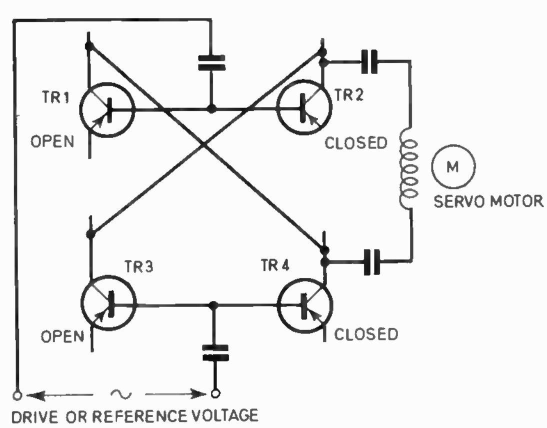


Fig. 9. Reversing the servo direction.

An alternative solution to the problem would be to interswitch the sine and cosine connexions at the input end of the plotter. This would have the advantage of low current operation, and the circuit design would be similar to that of the switching device described for the cross-over of the servo motors. In any event the method chosen could be decided upon by experimentation.

4 Applications

The method outlined of coupling the two aids together appears to be theoretically feasible.

It would seem on balance to be more readily suited to straight line operation, unless the operator is prepared to meet the cost of additional equipment required for the interswitching of the Decca drive signals to the automatic plotter. This, in itself, would probably not amount to a great deal extra, but costs may rise with the design of suitable Decca charts for individual 'one-off' routes. However, straight-line operation is found in hovercraft navigation, airways flying, airfield approach paths, and on many standard sea routes for shipping. The system may also have an application in the navigation of land vehicles over desert country.

The traditional presentation of the flight log and the marine automatic plotter need not be interfered with at all, if it is felt that the presence of an E-transformer would mar the presentation. A separate plotter could be mounted in some convenient position in the vehicle and used for this purpose only. An interesting extension of this idea, and which would also help to reduce the 'one-off' problem, would be to use a marker in place of the pen to lay down a route using a conducting fluid in place of ink. If the route was first traversed accurately using conventional navigation procedures, the I-strip would be laid out by the marker and thus provide a memory run which could be used again. The route would not necessarily have to be a long one, and could perhaps be an approach to land for any airfield in the country. It would then provide an approach aid which would be completely independent of facilities at the airfield. A similar application exists for the harbour control of shipping, and the exact run to a marker buoy or lock entrance with overall monitoring by shore-based radar control could prove useful. Military applications in providing precise runs to a target or in the patrol of fixed borders also might be considered.

Short of installing an elaborate simulation system in the laboratory, an analysis of errors can only be satisfactorily carried out by conducting extensive operational trials. A thorough knowledge of craft steering and control dynamics would be needed before any sensible attempts could be made. It is appreciated that errors will arise from the pictorial presentation itself apart from inherent errors in the transducer design employed. Such trials will reveal limitations, some of which have been indicated (Sect. 3.1).

5 Acknowledgments

The author would like to thank Professor D. R. Towill for helpful suggestions, and Professor A. D. Couper for his encouragement. Also, he wishes to acknowledge the facilities provided by the University of Wales Institute of Science and Technology.

Manuscript first received by the Institution in March 1970 as a Technical Paper submitted to gain exemption from the Graduateship Examination. Received in revised form on 13th January 1972 and in final form on 3rd August 1972. (Paper No. 1508/AMMS55.)

The Author

Mr. W. H. P. Canner (Member 1970) received technical training as an Aircraft Apprentice Electrician at No. 1 School of Technical Training, RAF Halton, and subsequently volunteered for flying training as a navigator. He was commisioned in 1950 serving both in Bomber Command and Flying Training Command, and he lectured at the RAF College, Cranwell, in navigation and electronic aids to navigation prior to leaving the service in 1956 as a Flight Lieutenant. Mr. Canner then taught at the College of Air Training, Hamble, and studied for a civil Flight Navigator's Licence. He joined Smiths (England) Ltd., as instructor on automatic pilots and flight systems in 1960, but moved to the Department of Navigation at the then College of Advanced Technology, Cardiff, in 1961. He currently lectures on ship control systems and electronics in the Department of Maritime Studies in the University of Wales Institute of Science and Technology.

UDC 621.396.677.73

Corrugated conical horns with arbitrary corrugation depth

M. S. NARASIMHAN, Ph.D.*

Based on a paper presented at the 1971 European Microwave Conference in Stockholm.

SUMMARY

The form of fields in a conical horn with uniform circumferential corrugations, when the corrugation depth assumes arbitrary values in the interval $0{\cdot}25 \leq (h/\lambda_0) \leq 0{\cdot}5$ is investigated in this paper. Assuming the corrugations to be infinitely thin and sufficiently close-packed, an impedance boundary condition is imposed on the fields in the axial region. Subject to a far-field approximation, accurate expressions are derived for the aperture field through a hybrid-mode formalism, which are subsequently used to calculate numerically the diffracted far-field of the horn supporting the HE_{11} mode, using Silver's formula. Satisfactory agreement between calculated and measured values of the far-field pattern for a wide-flare horn having arbitrary values of h/λ_0 supports the validity of the theory presented. When the half-flare angle (α_0) is less than 30°, a closed form expression is derived for the on-axis gain of the horn.

* *Department of Electrical Engineering, Indian Institute of Technology, Madras 36. India.*

List of Principal Symbols

α_0, α_1	half-flare angle of the horn measured at the boundary of the corrugations and at the base of the corrugations respectively.
r_0	flare-length of the horn
L	axial length of the horn
a	aperture radius of the horn
t	thickness of the fins forming the corrugated boundary
w	spacing between any two adjacent fins
h	depth of the corrugations
r, θ, ϕ	spherical polar coordinates of a point at which electromagnetic fields are considered
$\mathbf{a}_r, \mathbf{a}_\theta, \mathbf{a}_\phi$	unit vectors associated with (r, θ, ϕ)
$\mathbf{E}_t, \mathbf{H}_t$	aperture fields tangential to the spherical cap defined by $r = r_0$ and $\theta = \alpha_0$
E^i, H^i	fields within a corrugation close to the aperture
$J_n(x)$	Bessel function of first kind and nth order
$U_n(w, z)$	$\sum_{m=0}^{\infty} (-1)^m \left(\frac{w}{z}\right)^{n+2m} J_{n+2m}(z)$
$b_n(x)$	spherical Hankel function of second kind
$B_n(x)$	$xb_n(x)$
$P_n^1(\cos\theta)$, $Q_n^1(\cos\theta)$	associated Legendre function of first and second kinds, respectively, of order n
$L_n^1(\cos\theta)$	$P_n^1(\cos\theta) - B_{11} Q_n^1(\cos\theta)$
B_{11}	constant associated with function $L_n^1(\cos\theta)$
C_{11}	normalization constant associated with the far-field
b_{11}, a_{11}	amplitudes of the potentials associated with the H_{11} and E_{11} modes, respectively, in the axial region of the horn
λ_0	free-space wavelength
k	$2\pi/\lambda_0$
Z_0	impedance of free space
Y_0	$1/Z_0$
μ_0, ε_0	permeability and permittivity of free-space
P_t	total power radiated by the horn
$P(\theta, \phi)$	power radiated per unit solid angle in the direction (θ, ϕ)
$G(0, 0)$	on-axis gain

1 Introduction

The corrugated conical horn has been the subject of study for many workers in recent years, because of its attractive features as a primary feed in large paraboloidal reflectors for communication-satellite earth stations and for radio astronomical research. Most of the workers who have significantly contributed to the analytical study of corrugated conical horn[1-5] have focused their attention on the balanced HE modes of

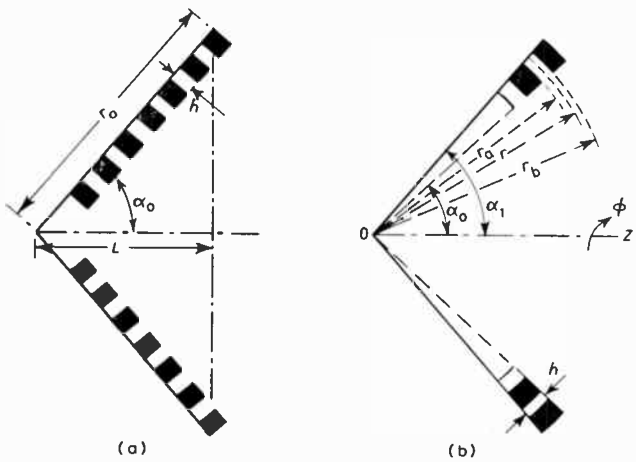


Fig. 1. Geometry of the corrugated conical horn.

excitation of the corrugated feed which correspond to a corrugation depth of one-fourth of the operating wavelength at the aperture edge. Thomas[6] has presented a limited theoretical analysis of the effect of varying the corrugation depth, by varying the operating frequency over an octave bandwidth. In this paper a more general analysis of spherical hybrid modes in corrugated conical horns than that by Jenken[1] is considered when the corrugation depth assumes arbitrary values in the interval ($\lambda_0/4 \leqq h \leqq \lambda_0/2$). Furthermore, an analytical study of the radiation properties (supported by experimental observations) of conical horns with circumferential corrugations of arbitrary depth is also presented.

2 Solution for Spherical Hybrid Modes

Analysis of fields in the axial region of a corrugated conical horn shown in Fig. 1(a), when the corrugation depth assumes arbitrary values in the interval $\lambda_0/4 \leqq h \leqq \lambda_0/2$ is based on a number of approximations. It is assumed first that the horn is excited by a circular waveguide proportioned to carry the dominant TE_{11} mode, and secondly that the thickness of the fins forming the corrugated boundary is very small when compared with the spacing between any two adjacent fins ($t \ll w$). Further, the number of corrugations within one wavelength of the operating frequency is assumed to be large (i.e. $t+w \ll \lambda_0$), so that the space harmonics may be ignored and fields in the axial region may be represented by a single spherical hybrid mode. This can be realized in practice by choosing $\lambda_0/(t+w) \geqq 10$ and $t \leqq w$.

The corrugated boundary shown in Fig. 1(a) does not coincide with any one of the three independent orthogonal surfaces of a coordinate system in which Maxwell's equation can be easily solved and the labour involved in deriving an exact solution of Maxwell's equation for this geometry is not commensurate with the results that can be expected. Therefore the boundary of the corrugations is slightly modified, as shown in Fig. 1(b), in order to facilitate solution of fields within the corrugations and in the axial region as well, without significant loss of accuracy.

The electromagnetic fields (E^i, H^i) within a corrugation (which is close to the aperture edge) can be derived from the vector potential $\mathbf{G} = \mathbf{a}_r u_r^i$, where

$$u_r^i = A_{11} B_n(kr) L_n^1(\cos\theta)\, e^{j\phi}. \qquad (1)$$

The time convention $e^{j\omega t}$ is implicit throughout.

When there is a large number of corrugations within one wavelength, the TE modes cannot exist within the corrugations and only TM modes are possible. The reason for this is that for TE modes of excitation ($E_r = 0$) the closely spaced corrugations would force both E_θ and E_ϕ to vanish within the slot, when

$$\left(\frac{t+w}{\lambda_0}\right) \to 0.$$

The TM fields ($\mathbf{E}^i$, $\mathbf{H}^i$) within the corrugations may be obtained from equation (1) and when α_0 is small ($\alpha_0 < 30°$) a quasi-cylindrical approximation of the spherical wave function indicated by the author in previous papers[7,8] may be used in order to facilitate the analysis. This implies that for field representation within the corrugations, cylindrical wave functions corresponding to the spherical wave function $L_n'(\cos\theta)$ may be used. Such a procedure, after considerable algebraic manipulations, leads to the following expression for the admittance of a corrugation (sufficiently away from the horn-apex and close to the aperture) at its open end:

$$Y_r \simeq Y_z = \frac{H_\phi^i}{E_z^i} = \frac{jY_0[J_1'(k_0a)N_1(k_0p) - J_1(k_0p)N_1'(k_0a)]}{J_1(k_0p)N_1(k_0a) - J_1(k_0a)N_1(k_0p)}$$
$$\simeq jY_0 \cot\left(\frac{2\pi}{\lambda_0} h\right) \qquad (2a)$$

where $ka \gg 1$ and $p = a+h$.

For large values of $\alpha_0(\sim 90°)$, an analytically simple and sufficiently accurate asymptotic solution[9] for the spherical wave function $L_n^1(\cos\theta)$, may be used to obtain expressions for H_ϕ and E_r^i. Subsequently these expressions also lead to the result

$$Y_r = H_\phi^i/E_r^i = jY_0 \cot\left(\frac{2\pi}{\lambda_0} h\right). \qquad (2b)$$

Identity of equations (2a) and (2b) indicates the validity of using a single expression for Y_r for both small and wide-flare horns.

In the axial region when the corrugation depth is arbitrary, unbalanced spherical hybrid modes of the form HE_{11} is present, which can be decomposed into TE_{11} and TM_{11} modes. The potentials associated with the TE_{11} and TM_{11} modes are given by

$$\begin{bmatrix} u^m \\ u^e \end{bmatrix} = \begin{bmatrix} b_{11} \\ a_{11} \end{bmatrix} B_s(kr) P_s^1(\cos\theta)\, e^{j\phi} \qquad (3)$$

respectively.

Subject to a far-field approximation, the components of the aperture field are given by

$$E_r = -ja_{11}\left(\frac{\mu_0}{\varepsilon_0}\right)^{\frac{1}{2}} \frac{B_s(kr)}{r}\frac{s(s+1)}{kr} P_s^1(\cos\theta)\, e^{j\phi} \qquad (4)$$

$$E_\theta = -jb_{11}\frac{B_s(kr)}{r}\left[\frac{P_s^1(\cos\theta)}{\sin\theta} + \gamma\frac{dP_s^1(\cos\theta)}{d\theta}\right] e^{j\phi} \qquad (5)$$

$$E_\phi = b_{11}\frac{B_s(kr)}{r}\left[\frac{dP_s^1(\cos\theta)}{d\theta} + \gamma\frac{P_s^1(\cos\theta)}{\sin\theta}\right]e^{j\phi} \tag{6}$$

where

$$\gamma = -j(a_{11}/b_{11})\sqrt{\mu_0/\epsilon_0}. \tag{7}$$

It follows from equation (7) that γ is the ratio of the E-field to that of the H-field present in the axial region. On applying the boundary condition, $E_\phi = 0$ at $\theta = \alpha_0$, one obtains

$$\left[\frac{dP_s^1(\cos\theta)}{d\theta} + \gamma\frac{P_s^1(\cos\theta)}{\sin\theta}\right]_{\theta=\alpha_0} = 0. \tag{8}$$

The following equation results on matching the fields within the corrugations with those in the axial region

$$\frac{s(s+1)\alpha_0}{kr_0(\gamma-1/\gamma)} = \tan(kh). \tag{9}$$

For prescribed values of α_0 and γ (where $0 \leqq \gamma \leqq 1$), the values of s appearing in equation (8) were computed numerically using a digital-computer based iterative algorithm. Details relating to this have been treated elsewhere.[8] Values of s computed as function of α_0 with γ as a parameter are shown in Fig. 2. From known values of γ and kr_0, the normalized corrugation depth (h/λ_0) can be calculated from equation (9). Alternatively, one can read the value of γ for any arbitrary value of $h(\lambda_0/4 \leqq h \leqq \lambda_0/2)$ from a graph plotted between γ and h/λ_0. Once γ is known for a prescribed value of h/λ_0, the aperture field is also known.†

3 Calculation of the Far-Field Radiation Patterns

Different techniques are available for calculating accurately the radiation patterns of flared circular horns from a knowledge of the aperture field distribution. The classical multipole expansion technique, first used by Potter[10] and subsequently by others,[2] appears to be one of the most accurate methods of calculating the radiation patterns of circular aperture antennas. Another method of computation of far-field uses Silver's formula.[11] James and Longdon[12] have established that the modal function (or multipole) expansion method and Silver's formula are mathematically equivalent, when the observation point lies in the far-field. However, Silver's formula involves fewer special functions and integrations than the other method for calculating the radiated far-field. Therefore Silver's formula is used for the analysis of far-field patterns.

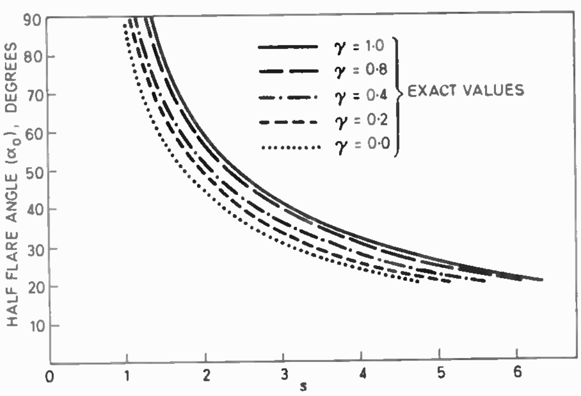


Fig. 2. Variation of the eigenvalue s with α_0 for different values of γ.

† It may be noted that when $\gamma = 0$, the horn supports the TE_{11} mode and for $\gamma = 1$, it supports a balanced HE_{11} mode.

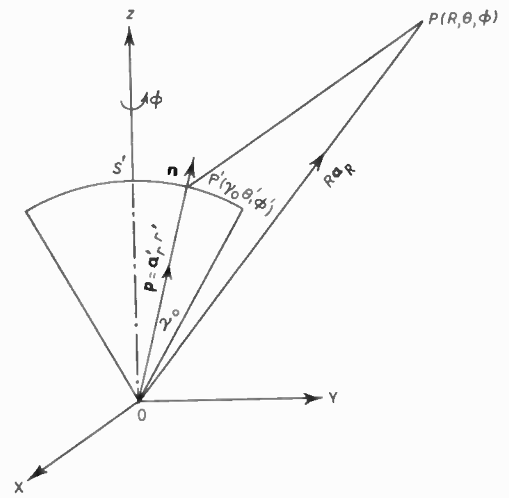


Fig. 3. Coordinate system for radiation formula.

Figure 3 shows the coordinate system for the radiation formula given by[10]

$$\mathbf{E}_P = \frac{-jk}{4\pi R}e^{-jkR}\mathbf{a}_R\times\int_{S'}[\mathbf{n}\times\mathbf{E}'_t - Z_0\mathbf{a}_R\times(\mathbf{n}\times\mathbf{H}'_t)]\times \times e^{jk\mathbf{P}\cdot\mathbf{a}_R}\,dS' \tag{10}$$

In equation (10), variables relating to the aperture are primed and those of the far-field are unprimed. Assuming that $\mathbf{E}'_t$ and $\mathbf{H}'_t$ obey a far-field approximation $kr_0 \gg s \gg 1$ so that[13]

$$\frac{E'_\theta}{H'_\phi} = -\frac{E'_\phi}{H'_\theta} = Z_0 \text{ over } S'. \tag{11}$$

The following expression is obtained for the far-field radiation patterns from equation (10) after extensive manipulations[9]:

$$\mathbf{E}_P = C_{11}\left(\frac{e^{-jkR}}{R}\right)(\mathbf{a}_\theta N_\theta^F \sin\phi + \mathbf{a}_\phi N_\phi^F \cos\phi) \tag{12}$$

where

$$N_\theta^F = N_\theta^{F_r} + jN_\theta^{F_i}, \quad N_\phi^F = N_\phi^{F_r} + jN_\phi^{F_i}$$

$$N_\theta^{F_r,F_i} = \int_0^{\alpha_0}\left[(p_1+p_2)\frac{\cos}{\sin}(u) \mp p_3\frac{\sin}{\cos}(u)\right]d\theta'$$

$$N_\phi^{F_r,F_i} = \int_0^{\alpha_0}\left[(p_1+p_2)\frac{\cos}{\sin}(u) \mp q_3\frac{\sin}{\cos}(u)\right]d\theta'$$

$$p_1 = F_1(\theta')(\cos\theta\cos\theta'+1)[J_0(u_1) - J_2\,u_1)]\sin\theta'$$

$$p_2 = F_2(\theta')(\cos\theta+\cos\theta')[J_0(u_1)+J_2(u_1)]\sin\theta'$$

$$p_3 = 2F_1(\theta')\sin\theta\sin^2\theta'\,J_1(u_1)$$

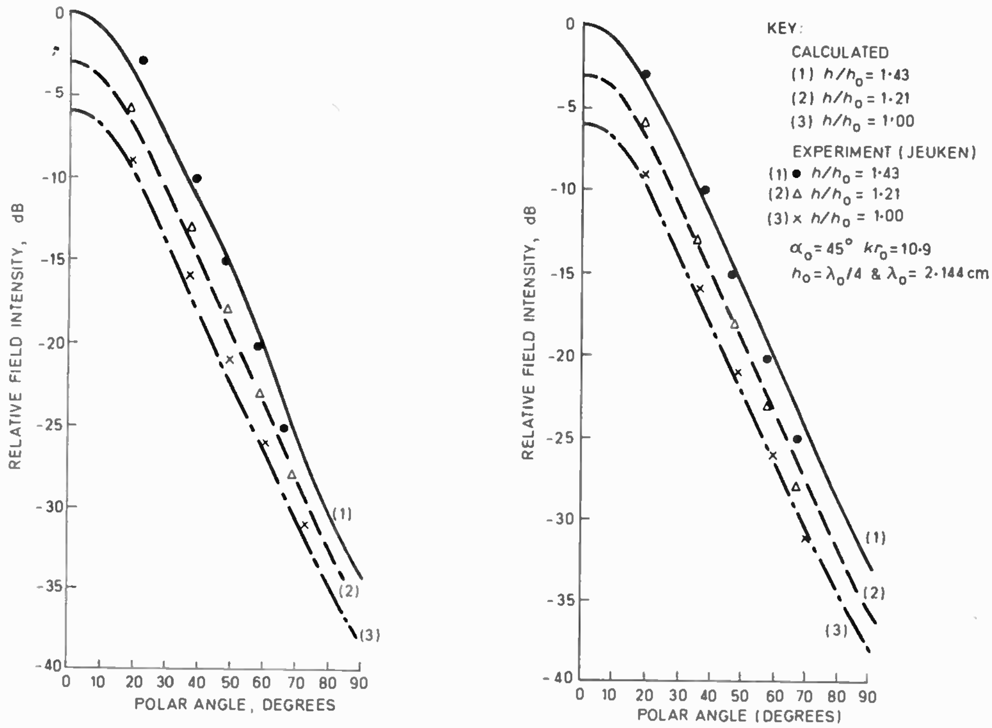
$$q_3 = 2F_2(\theta')\sin\theta\sin^2\theta'\,J_1(u_1)$$

$$u = kr_0\cos\theta\cos\theta' \text{ and } u_1 = kr_0\sin\theta\sin\theta'$$

$$F_1(\theta) = \frac{P_s^1(\cos\theta)}{\sin\theta} + \gamma\frac{dP_s^1(\cos\theta)}{d\theta}$$

$$F_2(\theta) = \frac{dP_s^1(\cos\theta)}{d\theta} + \gamma\frac{P_s^1(\cos\theta)}{\sin\theta}$$

(a) E-plane,



(b) H-plane.

Fig. 4. Far-field radiation patterns of a corrugated conical horn for arbitrary value of h/λ_0. To facilitate comparison, patterns (2) and (3) have been displaced from pattern (1) by 3 dB and 6 dB respectively.

Far-field radiation patterns of a wide-flare-corrugated conical horn with $\alpha_0 = 45°$ and $kr_0 = 10{\cdot}9$ obtained by calculation using equation (12), as well as by experiment, for three different values of 'h' are compared in Figs. 4(a) and 4(b) and a satisfactory agreement is noticed between the calculated and experimental results. Similar comparisons were made for several other horns with different values of α_0, r_0 and h and good agreement was noticed between calculated and experimental results. Figure 5 indicates how the far-field pattern is affected when h assumes several values† in the interval ($0{\cdot}25 \leqq h/\lambda_0 \leqq 0{\cdot}5$) for a typical conical corrugated feed with $\alpha_0 = 45°$ and $kr_0 = 15$.

It is of interest to obtain an expression for the on-axis gain. Here the attention is focused on horns with $\alpha_0 < 30°$, since in this case a closed form expression can be obtained for on-axis gain in terms of the significant dimensions of the horn. Furthermore, given the axial length, the value of a_0 in order to realize an optimum value for G_a is generally found to be less than 30°, unless L is too short. In order to arrive at a closed form expression for the on-axis gain, it is necessary to express the transverse E-field components over the aperture in a form simpler than the one given by equations (5) and (6). For doing so, an analytically simple and sufficiently accurate solution used by the author in previous papers,[4,5,7] for studying modes in conical and conical scalar horns with small and wide flare angles is employed. With some algebra, one obtains the following expressions for $\mathbf{E}_t = \mathbf{a}_\theta E_\theta + \mathbf{a}_\phi E_\phi$, where

$$E_\theta = jb_{11}v_{11}B_s(kr)(1/r)\left[\frac{J_1(x)}{x} + \gamma J_1'(x)\right]e^{j\phi} \tag{13}$$

$$E_\phi = b_{11}v_{11}B_s(kr)(1/r)\left[J_1'(x) + \gamma\frac{J_1(x)}{x}\right]e^{j\phi} \tag{14}$$

and where $x = v_{11}\theta/\alpha_0$ and v_{11} is the first non-vanishing root of

$$(1-\gamma)\frac{J_1(p)}{p} = J_0(p). \tag{15}$$

Roots of equation (15) have been tabulated[14] for a few discrete values of γ in the interval $0 \leqq \gamma \leqq 1$. It has

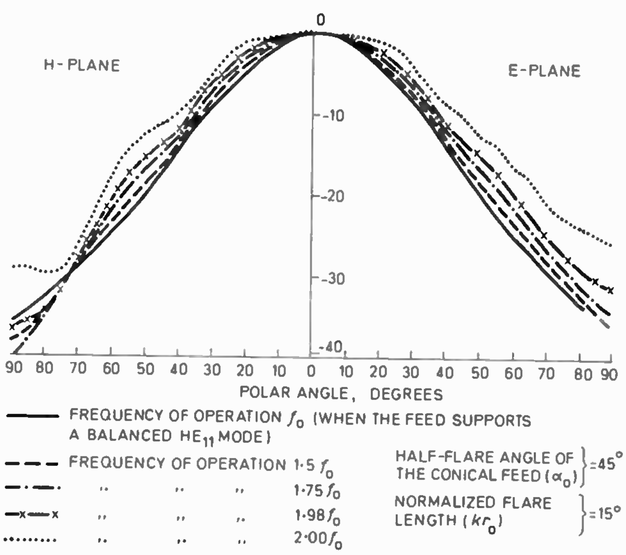


Fig. 5. Variation of E- and H-plane radiation patterns of the horn for changes in the operating frequency.

† In this case it is assumed that the variation in h is effected by varying the excitation frequency in the interval $1{\cdot}0 \leqq (f/f_0) \leqq 2$ where $\lambda_0 f_0 = c$.

been verified that the transverse electric field components over the aperture given by equations (14) and (15) are close to the exact solution for E_θ and E_ϕ (given by equations (5) and (6)) even for large values of α_0.

The expression for on-axis gain is given by

$$G_a = G(0, 0) = \frac{4\pi P(0, 0)}{P_t}. \tag{16}$$

One obtains the following expressions for $P(0, 0)$ and P_t (appearing in equation (16)) from equations (14) and (15) after extensive algebraic manipulations and by assuming that $\sin\alpha_0 \simeq \alpha_0$ and $\alpha_0 \simeq a/L$,

$$P(0, 0) = \frac{b_{11}^2 L^2(1+\gamma)^2[U_1^2(2q, v_{11}) + U_2^2(2q, v_{11})]}{8Z_0} \tag{17}$$

$$P_t = \frac{\pi b_{11}^2 a^2}{8Z_0}\left[(1+\gamma)^2\{J_0'^2(v_{11}) + J_0^2(v_{11})\} + \right.$$
$$\left. +(1-\gamma)^2\left\{J_2'^2(v_{11}) + \left(1 - \frac{4}{v_{11}^2}\right)J_2^2(v_{11})\right\}\right] \tag{18}$$

$$G_a = \left(\frac{4L^2}{a^2}\right)\left[\frac{U_1^2(2q, v_{11}) + U_2^2(2q, v_{11})}{J_0'^2(v_{11}) + J_0^2(v_{11}) + \left(\frac{1-\gamma}{1+\gamma}\right)^2\left\{J_2'^2(v_{11}) + J_2^2(v_{11})\left(1 - \frac{4}{v_{11}^2}\right)\right\}}\right] \tag{19}$$

It is of interest to note that when $\gamma = 1$, the expression for on-axis gain (equation (19)) reduces to that of a scalar horn supporting the balanced HE_{11} mode[6] and when $\gamma = 1$, with some algebraic manipulation the expression for G_a may be shown to reduce to the on-axis gain of a conical horn supporting a TE_{11} mode.[7] Clarricoats and Saha[15] have reported that the expression for G_a for the balanced HE_{11} mode (given by equation (19) with $\gamma = 1$) deviates from the more accurate form of G_a, based on the exact solution for E_θ and E_ϕ, only to the extent of 2·2% even when $\alpha_0 = 30°$. Observations stated above bear testimony to the validity of the expression for G_a given by equation (19).

4 Conclusions

A solution has been obtained for spherical hybrid modes in corrugated conical horns with arbitrary corrugation depth and the corrugated surface treated as an impedance boundary. When the circumferential corrugations have a depth equal to one-half of the operating wavelength, the hybrid-mode solution degenerates to the solution for the TE_{11} mode in a conical waveguide, thereby implying that for this particular choice of the corrugation depth, the corrugated surface appears to be a perfectly conducting metallic boundary for fields in the axial region. Satisfactory agreement between computed and measured far-field patterns of corrugated conical horns for a set of values of h and α_0 indicate the validity of the hybrid-mode formalism used as well as several assumptions made in the study of fields in the axial region of the horn. From a study of on-axis gain of horns with $\alpha_0 < 30°$, one observes that given h, α_0 and L, the upper and lower limits for the on-axis gain are given by the axial gains of conical and conical scalar horns of identical dimensions respectively.

5 Acknowledgments

The work reported here has been supported by the Centre for Systems and Devices, Department of Electrical Engineering, Indian Institute of Technology, Madras.

The author is thankful to Mr. R. V. Sitaram of the Microwave Engineering Group, Tata Institute of Fundamental Research, Bombay, for helpful discussions in the initial stages of the work. Thanks are also due to Dr. M. Jeuken of the Department of Electrical Engineering, Technological University, Eindhoven, for providing the experimental data on the far-field patterns of corrugated conical horns.

Computing facilities at the Computer Centre of the Tata Institute of Fundamental Research, Bombay, were made use of for the numerical computation of far-field patterns.

6 References

1. Clarricoats, P. J. B., 'Analysis of spherical hybrid modes in corrugated conical horn', *Electronics Letters*, **5**, pp. 189–90, 1st May 1969.
2. Clarricoats, P. J. B. and Saha, P. K., 'Radiation from wide-flare-angle scalar horns', *ibid*, **5**, pp. 376–8, 7th August 1969.
3. Jeuken, M. E. J. and Lambrechtse, C. W., 'Small corrugated conical-horn antenna with wide flare angle', *ibid*, **5**, pp. 489–90, 2nd October 1969.
4. Narasimhan, M. S. and Rao, B. V., 'Hybrid modes in corrugated conical horns', *ibid*, **6**, pp. 32–4, 22nd January 1970.
5. Narasimhan, M. S. and Rao, B. V., 'Diffraction by wide-flare-angle corrugated conical horns', *ibid*, **6**, pp. 469–71, 23rd July 1970.
6. Thomas, B. MacA., 'Bandwidth properties of corrugated conical horns', *ibid*, **5**, pp. 561–3, 30th October 1969.
7. Narasimhan, M. S. and Rao, B. V., 'Modes in a conical horn: new approach', *Proc. IEE* **118**, pp. 287–92, February 1971.
8. Narasimhan, M. S. 'Eigenvalues of a class of spherical wave functions', I.E.E.E. International Symposium Digest on Antennas and Propagation, Los Angeles, Cal. September 1971.
9. Narasimhan, M. S., 'Radiation properties of a class of microwave horn antennas', Ph.D. Thesis, Indian Institute of Technology, Bombay, September 1970.
10. Potter, P. D., 'Application of spherical wave theory to Cassegrainian-fed paraboloids', *IEEE Trans. on Antennas and Propagation*, **AP-15**, pp. 727–35, November 1967.
11. Silver, S., 'Microwave Antenna Theory and Design', Chapter 5. (McGraw-Hill, New York, 1949).
12. James, J. R. and Longdon, L. W., 'Calculation of radiation patterns', *Electronics Letters*, **5**, pp. 567–9, 30th October 1969.
13. Narasimhan, M. S. and Rao, B. V., 'Transmission properties of electromagnetic waves in conical waveguides', *Intl J. Electronics*, **27**, pp. 119–39, August 1969.
14. Abramovitz, M. and Stegun, I. A., 'Handbook of Mathematical Functions', pp. 355–433. (Dover, New York, 1965).
15. Clarricoats, P. J. B. and Saha, P. K., 'Propagation and radiation behaviour of corrugated feeds: Part 2', *Proc. IEE*, **118**, pp. 1177–86, September 1971.

Manuscript first received by the Institution on 23rd May 1972 and in final form on 3rd August 1972. (Paper No. 1509/CC 161).

UDC 621.315.61:621.372.81.09

An overview of point matching

Discussion contributions on the paper: 'Point-matched solutions for propagating modes on arbitrarily shaped dielectric rods'

Part I
R. H. T. BATES, D.Sc.(Eng.), C.Eng., F.I.E.E.*

Parts II and IV
J. R. JAMES, B.Sc., Ph.D., A.F.I.M.A., C.Eng., M.I.E.R.E.‡ and
I. N. L. GALLETT, B.Sc.‡

Part III
R. F. MILLAR, D.Sc.†

* Department of Electrical Engineering, University of Canterbury, Christchurch, New Zealand.

† Radio and Electrical Engineering Division, National Research Council of Canada, Ottawa, Canada, K1A 0R8.

‡ Department of Electrical and Electronic Engineering, Royal Military College of Science, Shrivenham, Swindon, Wiltshire SN6 8LA.

Part I: R. H. T. BATES

James and Gallett[1]§ have recently given a penetrating critique of point matching techniques and have concluded that the characteristics of all open dielectric waveguides can be accurately determined by straightforward applications of point matching. Their arguments and their computational and experimental results are compelling, and would be completely convincing were there not other detailed results[2,3] which are in partial disagreement with theirs.

The purpose of this note is to attempt to gain a comprehensive overview of point matching by making a number of comments on James and Gallett's paper,[1] indicating where there is agreement and disagreement with other work.

1. James and Gallett underestimate the power of Millar's analysis of field singularities and they do not reference the main body of Millar's work.[4–8] In essence, Millar's thesis can be paraphrased as follows: the two series in equation (5) of James and Gallett[1] necessarily possess analytic continuations into the complex ρ-plane, so that straightforward point matching is valid only if the two series converge for all values of ρ in the annulus $\rho_1 \leqslant |\rho| \leqslant \rho_2$ (refer to Fig. 1 of James and Gallett,[1] reproduced herein in modified form as Fig. A, for the definitions of ρ_1 and ρ_2).

2. Recent results for hollow waveguides[3] confirm that Millar's analysis places only weak constraints on field representations in waveguides with cross-sections which are circular, elliptical or of the shape shown in Figs. 9 and 10 of James and Gallett.[1] Bates and Ng[3] show that accurate results can sometimes be obtained even when there is a sharp corner at which the field must be singular (in its derivatives). Computational experience for hollow waveguides[2,3,9] suggests that straightforward point matching gives as accurate results as any method provided the cross-section is convex when viewed from outside or is only weakly concave (compare the cross-section shown in Figs. 9 and 10 of James and Gallett[1] with the cross-section treated by Fuller and Audeh[10]).

3. Straightforward point matching becomes virtually useless (at least for field computations; eigenvalues are inaccurate but still of the correct order) when Millar's analysis places strong constraints on field representations, as is true for hollow waveguides with strongly concave cross-sections, such as ridge waveguides. (Refer to Tables II and III and Fig. 3 of Ng and Bates.[2] Other appropriate examples are given by Bates and Ng.[3]) Note that the breakdown of straight-forward point matching is not necessarily dependent on the presence of sharp corners at which the field is necessarily singular. Bates and Ng[3] present examples of waveguides for which the field is necessarily analytic close to all points on their boundaries, and yet straightforward point matching breaks down completely.

§ References other than those given in the original paper are listed on page 200.

4. Computational experience[2,3,9] suggests that the equilateral triangle cross-section of Figs. 11–14 of James and Gallett[1] should cause little trouble for straightforward point matching of the internal field. However, the results[2,3] quoted in item 3 above would suggest that there might be difficulty for the external field, which is confirmed by James and Gallett's[1] remark concerning the computational difficulties associated with cross-sections having sharp corners. They observe that these difficulties are greater for dielectric waveguides than for closed metallic waveguides. However, the results of James and Gallett exhibit no sign of the large errors found in similar cases (but for closed metallic waveguides) by Ng and Bates[2] and Bates and Ng.[3]

5. The intermediate expansions of James and Gallett[1] are related both to Bolle and Fye's method[13] of 'overlapping domains' and to a systematic procedure, called extended point matching, described in detail by Bates and Ng.[3]

6. The penultimate paragraph of Section 5 of James and Gallett[1] overlooks an essential simplicity of all point matching techniques. In cases where they are theoretically sound, the computational accuracy is a function only of the number of terms used in the field representations. The accuracy of the uniformly valid integral equation techniques usually depends critically on details of numerical integration algorithms, which often can only be identified by computationally protracted procedures.[2]

7. James and Gallett's[1] recognition of the stability of point matching solutions of field problems governed by Laplace's equation highlights the inappropriateness of attempting to construct general computational approaches suitable for handling both conservative and non-conservative fields.[9]

8. Reference 20 of James and Gallett[1] is surely a sufficient answer to Burrows (references 17, 19 and 21 of James and Gallett[1]). However, the supposed universal equivalence of straightforward point matching and Waterman's extended boundary condition formulation (reference 49 of James and Gallett) is contradicted spectacularly (in particular cases for which straightforward point matching is invalid) by Fig. 3 of Ng and Bates[2] and by other appropriate examples in Bates and Ng.[3]

9. Section 8.5 of James and Gallett[1] needs some amplification. The extended boundary condition formulation and other uniformly valid integral equation formulations are based on Green's theorem and/or other exact deductions from Maxwell's equations. On the other hand, straightforward point matching is based merely on the *postulated* completeness of certain simplified field expansions. Erma observes that the same limitation applies to his iterative solutions (references 50 and 52 of James and Gallett[1]). The same is true of the simplified solutions obtained by Waterman in his later work.[11,12] Millar's analysis[4–8] can be looked on as a step towards establishing procedures for determining under what conditions simplified field expansions are valid.

There are two final comments which need to be made:

(*a*) In item 4 above it was pointed out that the errors in James and Gallett's computations for equilateral triangular dielectric waveguides appeared small compared with what might be expected from computational experience with hollow metallic waveguides. It would be interesting to know if the errors really do decrease with increasing N. So, it is hoped that it will be feasible for James and Gallett to make the required computations.

(*b*) If in the case considered in item 4 above the errors do tend to increase with N then the computational experience of Bates and Ng[2,3,9] will be confirmed. However, if the errors decrease it will indicate *either* that there may be significant differences between the numerical convergence of internal and external point matching solutions *or* that, for any type of point matching solution, the only characteristic of the waveguide cross-section which significantly degrades the computational accuracy is the presence of deep re-entrant parts.

Since writing the above, I have seen James and Gallett's reply. Although I remain certain that Waterman and I have demonstrated the global theoretical validity of the e.b.c. formulations (even when the kernels of the null-field integrals are expanded in cylindrical eigenfunctions), yet I do not disagree from a practical computational point of view with the spirit of James and Gallett's comments on my items 8 and 9. The point is that for the null-field method to be significantly superior to straightforward point matching, when C is a curve with deep re-entrant parts, the null-field integrals have to be evaluated extremely accurately and the eigenfunctions in which $F(C)$ is expanded must closely satisfy the physics of the situation (as James and Gallett note in their comment on my item 8). Ng and I discuss this at length elsewhere.[2,3,9]

Manuscript first received by the Institution on 21st August 1972 and in final form on 8th January 1973.

Part II: J. R. JAMES and I. N. L. GALLETT

Dr. Bates kindly made available references 2 and 3 in advance of publication and therefore we have been able to comment on these in addition to some of the other issues that he raises. We will take the items in the order presented above.

1. We singled out this particular reference of Millar because it was directly relevant to our work and our criticism of his sufficiency condition was based on the following mathematical details. We take without loss of generality, an arbitrarily shaped infinitely long

dielectric rod whose cross section is bounded by contour C, Fig. A. Γ_1 and Γ_2 are the inscribed and exscribed circles respectively with respect to the origin O. Let the Rayleigh hypothesis be assumed to hold then the fields external to C can be derived from electric and magnetic vector potentials $\mathbf{A}^E$ and $\mathbf{A}^H$ respectively where

$$\begin{Bmatrix} \mathbf{A}^E \\ \mathbf{A}^H \end{Bmatrix} = \sum_{n=-\infty}^{\infty} \begin{Bmatrix} s_n^E \\ s_n^H \end{Bmatrix} H_n^{(1)}(k_2\rho)\, e^{jn\phi}\, \hat{\mathbf{z}} \qquad \text{(i)}$$

and k_2 is positive imaginary for non-leaky surface wave solutions and n is an integer. $\hat{\mathbf{z}}$ is a unit vector in the axial z direction. The corresponding external E and H fields derived from eqn. (i) satisfy the appropriate boundary conditions on C and if subscript T denotes the tangential components we may express these boundary conditions as

$$(E_z, E_T, H_z, H_T)|_C = \{f_z^E(\rho,\phi), f_T^E(\rho,\phi), f_z^H(\rho,\phi), f_T^H(\rho,\phi)\};\ \rho \text{ on } C \qquad \text{(ii)}$$

where the functions $f(\rho, \phi)$ are defined by the internal E and H fields. Let $\mathbf{A}_0^E$ and $\mathbf{A}_0^H$ denote the analytic continuation of $\mathbf{A}^E$ and $\mathbf{A}^H$ respectively, eqn. (i) into the domain D bounded by C and Γ_1. Following Millar's development we next assume that $\mathbf{A}_0^E$ and $\mathbf{A}_0^H$, and hence their corresponding E and H fields have no singularities in D, thus finiteness of all expansions is ensured in D.

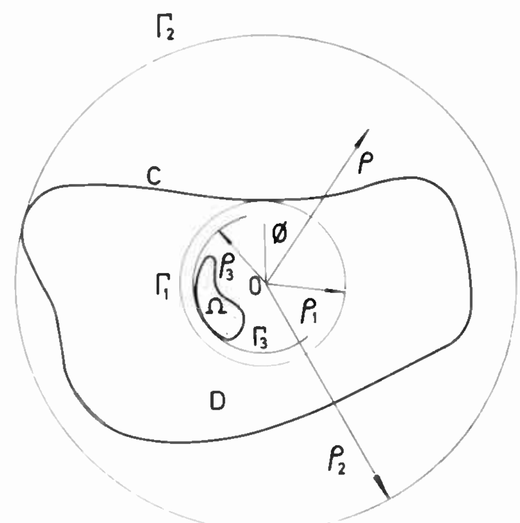


Fig. A. Configuration of inscribed and exscribed circles Γ_1 and Γ_2 respectively relative to contour C and origin O.

Suppose now that $\mathbf{A}_1^E$ and $\mathbf{A}_1^H$ are expansions of cylindrical functions that satisfy the boundary conditions imposed by C and Γ_1 on D. Then by the assumed hypothesis a valid expansion in D which is also valid in the annulus bounded by Γ_1 and Γ_2 is

$$\begin{Bmatrix} \mathbf{A}_1^E \\ \mathbf{A}_1^H \end{Bmatrix} = \sum_{n=-\infty}^{\infty} \left[\begin{Bmatrix} a_{n1}^E \\ a_{n1}^H \end{Bmatrix} J_n(k_2\rho) + \begin{Bmatrix} b_{n1}^E \\ b_{n1}^H \end{Bmatrix} Y_n(k_2\rho) \right] e^{jn\phi}\, \hat{\mathbf{z}}. \qquad \text{(iii)}$$

We now note that boundary conditions are unspecified on sets of points of Γ_1 of non-zero measure, i.e. as sketched in Fig. A the boundary conditions on Γ_1 are only specified at the point where it coincides with C; thus by the uniqueness theorem we are free to choose a sequence of vector potentials $\mathbf{A}_i^E$ and $\mathbf{A}_i^H$ ($i = 2, 3, 4 \ldots$) with expansion coefficients in eqn. (iii) a_{ni}^E, b_{ni}^E, a_{ni}^H and b_{ni}^H. Clearly the boundary conditions as specified are not sufficient to uniquely-define the interior potentials. On C however the E and H fields corresponding to each choice of internal vector potential ($\mathbf{A}_i^E$, $\mathbf{A}_i^H$) all satisfy eqn. (ii), thus the difference potentials ($\mathbf{A}_i^E - \mathbf{A}_j^E$) and ($\mathbf{A}_i^H - \mathbf{A}_j^H$), $i \neq j$, ($j = 1, 2, 3 \ldots$) are identically zero in the region exterior to C by the uniqueness theorem. Consequently $\mathbf{A}_i^E$, $i = 1, 2, 3 \ldots$ are identical in this exterior region bounded by C and Γ_2 and by the assumed Rayleigh hypothesis they are identical to $\mathbf{A}_i^E$ eqn. (A) in this latter region; likewise for the $\mathbf{A}_i^H$ potentials. $\mathbf{A}_i^E$ and $\mathbf{A}_i^H$ are therefore various forms of the analytic continuations $\mathbf{A}_0^E$ and $\mathbf{A}_0^E$ respectively in D and are non-unique irrespective of their finiteness as postulated.

We have proved that finiteness, brought about by postulating that the singularities of eqn. (i) are contained within the circle Γ_1, is not a sufficient conditions to ensure uniqueness of the analytic continuation $\mathbf{A}_0^E$ and $\mathbf{A}_0^H$ of eqn. (i) into the region D. We cannot therefore accept Millar's sufficiency proof in its present form which relies on the uniqueness of the fields continued into D. The analytic continuation of a function is of course a unique process but the unusual situation here is that the continuation is being attempted over a domain where the function is not uniquely defined.

The non-uniqueness of the fields in D can be more readily established by an alternate proof which is direct and physically appealing as follows. Suppose that the singularities of the potentials $\mathbf{A}^E$ and $\mathbf{A}^H$ are located in the region Ω within C (Fig. A). The singularities corresponding to $\mathbf{A}^E$ and $\mathbf{A}^H$ may validly be regarded as filaments of current of cross-sectional area da', current density $J_{\hat{z}}^E$ and $J_{\hat{z}}^H$ respectively per unit length and a behaviour on the circle at infinity of the form $H_0^{(1)}(k_2\rho)$. These singularities or equivalent sources generate a field in the region external to C that satisfies eqn. (ii); Ω can be a multiplicity of disconnected regions. The potentials corresponding to these currents are

$$\begin{Bmatrix} \mathbf{A}^E(\rho,\phi) \\ \mathbf{A}^H(\rho,\phi) \end{Bmatrix} = -\frac{j}{4}\int_\Omega \begin{Bmatrix} J^E(\rho',\phi') \\ J^H(\rho',\phi') \end{Bmatrix} H_0^{(1)}(k_2|\boldsymbol{\rho}-\boldsymbol{\rho}'|)\, da'\, \hat{\mathbf{z}} \qquad \text{(iv)}$$

where it understood that the integration is replaced by summation if the singularities are discrete line sources and primes on ρ, ϕ and da indicate integration variables. By the addition theorem we may expand $H_0^{(1)}(k_2|\boldsymbol{p}-\boldsymbol{p}'|$ in cylindrical eigenfunctions and equate the expansion coefficients of eqns. (iv) and (i) thus

$$\begin{Bmatrix} s_n^E \\ s_n^H \end{Bmatrix} = -\frac{j}{4}\int_\Omega \begin{Bmatrix} J^E(\rho',\phi') \\ J^H(\rho',\phi') \end{Bmatrix} J_n(k_2\rho')\, e^{jn\phi'}\, da' \qquad \text{(v)}$$

The necessary and sufficient condition for eqn. (v) and hence eqn. (i) the Rayleigh hypothesis to be valid is that ρ in eqn. (iv) is greater than ρ_3, the radius of the smallest circle Γ_3 centre O (Fig. A), containing the singularities J^E and J^H.

If $\rho_1 \geqslant \rho_3$ then eqn. (i) with coefficients as defined by eqn. (v) is a valid representation of the field in the region exterior to C and the Rayleigh hypothesis holds. In fact if and only if the exterior field uniquely defines the location and distributions of current in Ω then Millar's necessary and sufficient conditions are clearly a natural consequence of this alternate proof but to assume uniqueness violates established facts about equivalent sources. For instance, we can always choose a set of equivalent sources on C itself; as another example let it be supposed that the distribution Ω lies within Γ_1 and that Ω is uniquely determined by eqn. (ii); then by what has been proved in eqn. (v), the Rayleigh hypothesis of eqn. (i) is valid but the latter is known to represent a system of superimposed multipole equivalent line sources at the origin O which is contrary to the assumption that the sources Ω are uniquely defined within C. We have therefore deduced that the equivalent sources within C and hence their fields within C are not uniquely defined by the boundary conditions eqn. (ii) on C, thus agreeing precisely with the result above. When looked upon from the above equivalent source standpoint this is of course a well-known result and is amply discussed by Rumsey.[18]

Cases are easy to construct whereby the equivalent sources can never be contained wholly within Γ_1 since the external field can contain both incoming and outgoing waves in the proximity of C. (See, for instance, Fig. 7 of our paper.[1]) For such cases there will always be some extremity of Ω that lies a limiting distance in excess of ρ_1 from the origin and the Rayleigh hypothesis does not hold. We therefore arrive at the following theorem: A necessary and sufficient condition for the Rayleigh hypothesis (eqn. (i)) to be valid is that a set of equivalent sources exists which is bounded by Γ_1 and satisfies the boundary condition (eqn. (ii)) on C. The conversion of the theorem for the internal fields involves sources exterior to Γ_2 and is evident.

The outcome of this is that no particular singularity within C is relevant to the Rayleigh hypothesis and our practical computations[1] bear this out. For instance, we took C to be a circle and also an ellipse in which case the singularities identified by Millar are simply the origin and foci respectively; we found that 4th place eigenvalue convergence and good field plots were obtainable even when these singularities extended appreciably outside the circle Γ_1. Round-off error effects must inevitably obscure our observations but it is undeniable that the behaviour is compatible with our new necessary and sufficient conditions. That is the singularity at the origin O is just one of an infinite number of sources that will generate the field exterior to the circular contour C and from the high accuracy of the calculated eigenvalues it would appear that a set of equivalent sources exist within Γ_1, in particular the superimposed line sources representing the Rayleigh hypothesis at the origin O′ (Fig. 2 of our paper[1]). A similar interpretation applies to the elliptical case.

As we see it the situation is now as follows: if a given external field uniquely defines the internal sources and hence singularities within C then Millar's conditions are established. However, such a situation would be contrary to established fact about equivalent sources; it seems Millar has not considered the question of uniqueness of singularities in his papers[4–8] and apart from one remark[5] has not physically identified the latter as the source of the external fields. Bates says that we underestimate the power of Millar's analysis but we are not aware that a direct numerical test of the theory, other than our circle and elliptic examples,[1] has ever been carried out. Many recent discussions have, however, freely used the theory[16,17] to endorse various arguments.

It now remains to discuss whether the new necessary and sufficient conditions can be easily used. Millar[5–7] has shown how to probe the position of internal singularities by examining the radius of convergence of an expansion; it is not clear to us at present how these techniques stand if the singularities are not uniquely defined in position. One possible test of the Rayleigh hypothesis would be to attempt to find a set of equivalent sources situated on a circle of radius slightly less than Γ_1 (Fig. A), that satisfy eqn. (ii) but as far as we can see at present this would lead to an extensive numerical process on C which is reminiscent of Waterman's formulation[12] and is not attractive as a quick non-numerical test. In Section 3.3 of our paper[1] we gave two necessary conditions for the Rayleigh hypothesis to be valid which were based on the geometry of C and from these it is clear that for the type of contours that are of engineering interest, multiple field expansions are required and seldom will the Rayleigh hypothesis be valid. The Rayleigh hypothesis defined as a single expansion of outgoing waves, while of theoretical interest, is a rather special restrictive case when considered in the light of practical field requirements which generally demand multiple expansions for their representations by point-matching.

4. We have subsequently carried out extensive computations on these examples and find that both the internal and external fields very close to the apexes of the triangular contours show marked deviations from the anticipated field. It would appear from detailed tests so far that expansions of cylinder functions containing a computationally realistic number of terms cannot be made to fit these apexes. In fact not only the fields, but the eigenvalues can show an undesirable divergent behaviour for a large number of expansion terms; values of N in our paper[1] up to 24 have subsequently been examined and the best results have so far been obtained for $N = 18$ when an independent check showed that the eigenvalues were correct to four places of decimals.

6. It is evident[1] that for some geometries multiple expansions are necessary for a valid point-matching technique. Under these circumstances the good field fit is obtained at the expense of an increase in unknown coefficients and hence computation effort. Under these circumstances the simplicity of basic point-matching with only one expansion is lost. Since, as Bates asserts, the computational accuracy is a function of the number of terms used in the expansions it is evident that multiple expansions can incur a great loss in accuracy due to round-off error alone. We therefore endorse our conclusions in Section 5 of the paper.[1]

7. We would like to supplement our discussion in Section 3.2 of our paper[1] by drawing attention to an interesting text on 'Maximum principles in differential equations' by Protter and Weinberger.[19]

8. These results quoted by Bates are readily explained. We proved in condensed form[1] (Sect. 8.3) that if the extended boundary condition (e.b.c.) method (also called null-field method) derived by Bates is valid then it is mathematically equivalent to basic point-matching over a circular domain of the field space. Figures 7 and 10 show[3] that basic point-matching gives a bad fit for the case quoted while Fig. 10 includes some good results for an alternative expansion called extended point-matching. Figure 3 again shows[2] that basic point-matching gives poor results by comparison with a new adaptation of the above mentioned null-field method. Nowhere in these quoted results is a strict comparison made between the two theoretical cases that we originally analysed. We therefore cannot regard these results as contradictory to our analysis. It is interesting to note, however, that this new null-field method is derived from the e.b.c. formulation of eqn. (2) in reference 2 by forcing the current distribution $F(C)$ to have the appropriate periodicity for the sector waveguide under examination. This is done by a Fourier expansion of $F(C)$ over the arc length L (eqn. (3) of reference 2) and as such we would expect this to lead to a better field fit and hence more accurate results than basic point-matching. This is seen to be so from Table 1 of reference 2.

9. Bates states that the e.b.c. (null-field) methods that he has derived are free from the difficulties surrounding the basic point-matching technique in that no assumptions are required about completeness of expansions. We can find no basis for this suggestion not only because of our theoretical comparisons made in Section 8.3 of our paper,[1] but also in view of evidence presented by Bates himself: in ref. 22 of our paper.[1] Bates proved that basic point-matching and his original null-field formulation give identical eigenvalues for expansions truncated at the same term. Basic point-matching can fail to converge to the correct eigenvalues as acknowledged by Bates in item 3 above; thus the e.b.c. (null-field) method can also be inaccurate in this respect. We believe that the e.b.c. principle, originated by Waterman, is sound but it is our opinion that the completeness issue is introduced when the kernels of the subsequent integrals are taken as expansions of eigenfunctions. The range of validity of these expansions of eigenfunctions is confined to inscribed and exscribed circles for cylinder functions and the e.b.c. condition cannot therefore be enforced over all space either interior or exterior to the contour C with this particular choice of expansions.

In conclusion we have shown that much can be established about the Rayleigh hypothesis by giving the field singularities their valid physical interpretation of equivalent source distributions; in particular a new necessary and sufficient condition has been derived. Arguments have been put forward that even if these new conditions could be quickly applied in practice, the Rayleigh hypothesis defined as a simple expansion of outgoing waves is too restrictive from a user standpoint; most fields of engineering interest require multiple expansions as described in our paper[1] (Sect. 3.3). As such, the point-matching technique will continue to require the user to experiment with a variety of expansions when confronted by a new configuration. Dr Bates and his colleagues have made considerable progress in this respect in recent years and have demonstrated how accurate eigenvalues can be obtained if progressively better field-fitting expansions are used.

We would like to thank our colleague Dr. L. W. Longdon for several helpful discussions.

Manuscript first received by the Institution on 11th October 1972 and in final form on 12th December 1972.

Part III: R. F. MILLAR

James and Gallett[1] have critically examined the application of point-matching methods to electromagnetic problems involving guided waves on dielectric rods and in their Section 8, they have discussed some recent theories dealing with point matching. I should like to make a few comments on Section 8.1, which concerns some of my own work.[14] (The numbering of subsequent equations refers to reference 1.)

The necessary and sufficient condition mentioned by James and Gallett does not, in fact, refer to the validity of point-matched solutions. It concerns the validity of expansions such as (5) in representing the interior and exterior solution everywhere on C. To a given solution, there correspond unique coefficients in the series (5). These series may or may not converge on C. We will assume that the correct values of the coefficients have been found. (Whether or not the result of the point-matching process is convergent to these correct values, when (5) are valid on C, is a different matter. It has been pointed out recently by Holford[15] that there is no theory to guarantee convergence as the number of matching points tends to infinity.)

James and Gallett agree that in order for the infinite series for both the exterior and the interior solutions to converge on C, it is necessary that all singularities of the exterior field and its analytic continuation across C lie inside Γ_1, and that all singularities of the interior field and its analytic continuation across C lie outside Γ_2. On the other hand, they state that if the singularities satisfy this criterion, then the consequent convergence of the expansions on C does not ensure that the latter are valid field representations satisfying the boundary conditions.

That this statement is incorrect may be argued as follows. We suppose that the singularities of the exterior solution lie inside Γ_1. As demonstrated in reference 14, knowledge of this fact determines a bound on the coefficients of the second series of (5) for large values of n. There will be an infinity of different sets of coefficients that satisfy this bound, and for which the series converges on and outside C. To each such set there correspond different singularities distributed inside Γ_1 and different boundary values of the solution and its normal derivative on C. The same is true for the expansion of the interior field. But for only one set of coefficients in the exterior expansion, and one set in the interior expansion, will these series represent the (unique) solution to the given boundary-value problem. Thus, if by some means we are able to show that the singularities satisfy this criterion, it will follow that the appropriate series representation (5), with coefficients that correspond to the boundary-value problem under consideration, converges on and outside C to the exterior solution. Similarly, the interior representation converges on and inside C to the interior solution. Between Γ_1 and C, the exterior representation converges to the analytic continuation of the exterior solution. Between Γ_2 and C, the interior representation converges to the analytic continuation of the interior solution. Both continuations are unique for a given boundary-value problem.

Therefore, satisfaction of the condition on the location of the singularities is sufficient to ensure that the series expansions (5) are valid field representations satisfying the boundary conditions. This condition provides a criterion to determine whether or not the infinite series (5), with coefficients corresponding to the given problem, will converge on C. If the series do not both converge on C, then one cannot expect that the coefficients determined by truncating the series and matching at N points (some of which lie in the region of divergence) will tend to the correct limits as N tends to infinity. This is borne out by the work of Nielsen[16] (which the writer believes to be correct) on a different problem. Nielsen shows rather convincingly how use of a divergent representation can lead to error. On the other hand, the work of James and Gallett indicates that point matching does not yield poor numerical results in all cases where there is divergence of the infinite series somewhere on C, and Lewin[17] has observed that there may be an optimal number of terms that will give best results. Thus the above condition that guarantees convergence of the infinite series evidently can be violated sometimes without occasioning unacceptable error in numerical results.

Since writing the above comments, I have had the opportunity to see the observations of the authors on the contribution by R. H. T. Bates. In their reply, they expand on some of the points in the original paper.[1] The remaining remarks refer to their reply.

I have been unable to follow some of the authors' arguments, especially those presented in the paragraph that includes equations (iii). Since the Rayleigh hypothesis is satisfied, the series in equations (i) converge everywhere outside Γ_1 and may be used to continue $\mathbf{A}^{\mathrm{E}}$ and $\mathbf{A}^{\mathrm{H}}$ across C and into D. Thus $\mathbf{A}_0^{\mathrm{E}}$ and $\mathbf{A}_0^{\mathrm{H}}$ may be represented by the right-hand sides of equations (i). This representation is unique because analytic continuation is unique. Furthermore, the fact that boundary values are not specified on Γ_1 is essential, in order that the problem not be overdetermined. The values taken on Γ_1 by the analytic continuations of $\mathbf{A}^{\mathrm{E}}$ and $\mathbf{A}^{\mathrm{H}}$ are determined automatically through equations (i), and there is no freedom to specify them. The coefficients in equations (i) will be determined by matching the exterior fields (with wave number k_2) to the interior fields (with wave number k_1) on the boundary C. Finally, the reason for introducing the potentials $\mathbf{A}_i^{\mathrm{E}}$, $\mathbf{A}_i^{\mathrm{H}}$ ($i = 1, 2, 3, \ldots$) is not apparent. The authors argue that $(\mathbf{A}_i^{\mathrm{E}} - \mathbf{A}_j^{\mathrm{E}})$ ($i \neq j$) vanish identically in the region exterior to C. From this we may conclude that $\mathbf{A}_i^{\mathrm{E}}$ and $\mathbf{A}_j^{\mathrm{E}}$ have identical coefficients and that $\mathbf{A}_i^{\mathrm{E}} \equiv \mathbf{A}_j^{\mathrm{E}}$ wherever the series converge. Hence, if $\mathbf{A}_i^{\mathrm{E}}$ coincides with $\mathbf{A}^{\mathrm{E}}$ outside C, it provides the analytic continuation of $\mathbf{A}^{\mathrm{E}}$ across C and in D. Thus the series for $\mathbf{A}^{\mathrm{E}}$, $\mathbf{A}_0^{\mathrm{E}}$ and $\mathbf{A}_i^{\mathrm{E}}$ ($i = 1, 2, 3, \ldots$) coincide, as do the series for $\mathbf{A}^{\mathrm{H}}$, $\mathbf{A}_{\mathrm{H}}$ and $\mathbf{A}_i^{\mathrm{H}}$.

In a subsequent paragraph, the authors state that the singularities may validly be regarded as a distribution of electric current: but it is not true that such a distribution necessarily produces field singularities, and the validity of their statement has not been established. Certainly an isolated current filament will correspond to a field singularity. In the case of a perfectly conducting circular cylinder excited by a plane wave, the analytic continuation of the exterior scattered field is singular at the centre of the cylinder. But the surface distribution of current that actually gives rise to the scattered field does not produce field singularities on the surface: the exterior field can be continued across this surface as far as the centre of the cylinder. If the circular cylinder is composed of an imperfectly conducting material, the actual exterior scattered field is produced by currents flowing throughout the material. This field, too, can be continued into the interior of the cylinder as far as its singularity, which, once again, is at the centre of the cylinder. Consequently, in the authors' work, the domain Ω is not necessarily coincident with the field singularities.

As the authors point out, there are many equivalent sources that will produce the same exterior scattered field. But in general these equivalent sources are not singularities of the field. It is the confusing of equivalent sources with field singularities that may have led the authors to their criticism of the earlier work.[14]

As a further illustrative example, consider the scattering of a plane wave by a perfectly conducting elliptic cylinder. The exterior scattered field is actually produced by currents on the surface, but the singularities lie at the foci of the ellipse. (The currents are members of an infinite set of equivalent sources.) Outside any circle containing the ellipse—in fact, it suffices that it contain the foci—the field can be represented as a superposition of multipoles (which we may consider to be equivalent sources) at the centre of the circle. But it does not follow that this centre is a field singularity; at points outside the circle the field only seems to arise from multipoles at its centre, and when the solution is continued into this circle the true singularities are encountered at the foci.

In a given problem, the field singularities are unique. For the solution ($u(x, y)$, say) is an analytic function of

x and y, and its singularities are, by definition, just those points at which it fails to be analytic. In fact, the field behaviour in the neighbourhood of the singularities determines the solution in the following manner. Suppose that some component, u, of the exterior scattered field is represented in a transverse plane by

$$u(\rho, \phi) = \sum_{-\infty}^{\infty} A_n H_n^{(1)}(k_2 \rho)\, e^{jn\phi}.$$

(This representation is certainly valid outside Γ_2.) Then it may be shown than

$$A_n = \tfrac{1}{4} j \int_{\Lambda} \left[u \frac{\partial}{\partial v} - \frac{\partial u}{\partial v} \right] J_n(k_2 \rho)\, e^{-jn\phi}\, ds,$$

where Λ is a simple closed contour containing Γ_2. Here (ρ, ϕ) is a point on Λ and differentiation is along the outward normal. By Green's theorem, the value of the integral is unchanged if Λ is deformed through any domain in which u satisfies the Helmholtz equation. (This formula for A_n may be verified by choosing Λ to be a circle, and employing the above representation for $u(\rho, \phi)$ on the circle.) The exterior scattered field can be regarded as arising from equivalent sources on Λ. The above expression for A_n ($n = 0, \pm 1, \pm 2, \ldots$) shows how the densities (u and $\partial u/\partial v$) should be chosen to produce the required field. In particular, if Λ is contracted onto the singularities of the solution, then it is clear that the coefficients (and hence the solution) are determined by the field behaviour near the singularities.

To sum up, I believe that the criticisms of some results in reference [14] are based on misunderstanding. The authors apparently accept their validity if the singularities of the solution are unique. Since this is the case, I trust that they will now agree that the earlier work is sound.

Manuscript first received by the Institution on 23rd October 1972 and in final form on 29th January 1973.

Part IV: J. R. JAMES and I. N. L. GALLETT

We have shown by proof that Millar's convergence condition is not sufficient and have identified his form of continuation with Huygen's Principle; this leads to a modified version of the necessary and sufficient conditions embracing equivalent sources. Millar has now given reasons for rejecting our criticism but we are unable to accept his arguments because they do not resolve the underlying fundamental issues that our objections are based on. Our present opinion is that more mathematical evidence may be required, in addition to that presented by ourselves, to pin-point these fundamental issues and it is instructive to isolate the latter as follows; we will refer specifically to a two-dimensional elliptical wave problem of an open dielectric guide where C (Fig. A) is an ellipse; the generalization of these issues to other geometries and field behaviour is evident.

A Mathieu function expansion of the external field is known to be valid and when continued within C has singularities at the foci. If the external field can also be validly expanded in cylinder functions about some origin O then this particular representation when continued within C will yield a singularity at O. If the location of O can be chosen with some freedom then a variety if not infinite number, of singularities can be generated. The issue to settle is whether the singularities corresponding to the Mathieu functions have additional mathematical significance as claimed by Millar to those generic to other eigen-functions expansions. Clearly if no representation of the external field other than one involving Mathieu functions is valid, then only the foci singularities will exist; thus uniqueness of the foci singularities would imply that the Rayleigh hypothesis is not valid for an elliptical contour.

Another related issue is the two different methods of analytic continuation that have been used recently and their precise mathematical and physical meaning. The method used by Wilton and Mittra[20] is such that the continued fields are solutions to the boundary value problem (b.v.p.) everywhere in the domain of continuation which is entirely restricted to one side of C; consequently the continued fields are unique and have a unique physical interpretation. Millar's use of continuation leads to fields continued across C but these are not simultaneously solutions to both the interior and exterior b.v.p. with respect to C in the domain of continuation. The singularities corresponding to those fields continued across C may be identified as equivalent sources whose distribution within C is not uniquely defined by the boundary values on C. For example the foci singularities are elliptical multi-pole sources whereas a cylinder function expansion, if it exists, of the external field involves a superposition of cylinder type multipoles at the chosen origin. The fact that it is possible to generate the same external field from different source distributions and fields within C is a consequence of Huygen's Principle as explained by Rumsey[18] and in this sense the Rayleigh hypothesis is valid if the source distribution can be taken as a superposition of cylinder multi-poles at some nominated origin. The issue to recognize here is that Millar's use of continuation is in the Huygen's Principle sense in which case he cannot validly claim above that . . . 'in general these equivalent sources are not singularities of the field', and also . . . 'the true singularities are encountered at the foci'. In the example under discussion the Mathieu functions arise 'naturally' from the solution of the differential equation by the separation of variables method; one could therefore refer to the singularities corresponding to these functions as 'natural' rather than 'true' singularities.

We believe that when these fundamental issues are firmly mathematically established then the necessary and sufficient conditions for the validity of the Rayleigh hypothesis will become clearer. We should like to thank Dr. Millar for his above comments and also for helpful private correspondence on this topic.

Manuscript received by the Institution on 8th February 1973.

References

1. James, J. R. and Gallett, I. N. L., 'Point-matched solutions for propagating modes on arbitrarily-shaped dielectric rods', *The Radio and Electronic Engineer*, **42**, No. 3, pp. 103–13, March 1972.
2. Ng, F. L. and Bates, R. H. T., 'Null field method for waveguides of arbitrary cross section', *IEEE Trans. on Microwave Theory and Techniques*, **MTT-20**, pp. 658–62, 1972.
3. Bates, R. H. T. and Ng. F. L., 'Point matching computation of transverse resonances', *Int. J. Num. Meth. Engng*, **5**, No. 4, April 1973.
4. Millar, R. F., 'On the Rayleigh assumption in scattering by a periodic surface', *Proc. Camb. Phil. Soc.*, **65**, pp. 773–91, 1969.
5. Millar, R. F., 'The location of singularities of two-dimensional harmonic functions. 1: Theory', *SIAM J. Math. Anal.*, **1**, pp. 333–44, 1970.
6. Millar, R. F., 'The location of singularities of two-dimensional harmonic functions. II: Applications', *ibid.*, pp. 345–53, 1970.
7. Millar, R. F., 'Singularities of two-dimensional exterior solutions of the Helmholtz equation', *Proc. Camb. Phil. Soc.*, **69**, pp. 175–88, 1971.
8. Millar, R. F., 'On the Rayleigh assumption in scattering by a periodic surface. II', *ibid.*, pp. 217–25, 1971.
9. Bates, R. H. T. and Ng, F. L., 'Polarization-source formulation of electromagnetism and dielectric-loaded waveguides', *Proc. Instn Elect. Engrs*, **119**, pp. 1568–74, 1972.
10. Fuller, J. A. and Audeh, N. F., 'The point matching solution of uniform nonsymmetric waveguides', *IEEE Trans.*, **MTT-17**, pp. 114–5, 1969.
11. Waterman, P. C., 'Scattering by dielectric obstacles', *Alta Frequenza*, **38**, numero speciale, pp. 348–52, 1969.
12. Waterman, P. C., 'Symmetry, unitarity and geometry in electromagnetic scattering', *Phys. Rev.*, D, **3**, pp. 825–39, 1971.
13. Bolle, D. M. and Fye, D. M., 'Application of point-matching method to scattering from quadrilateral cylinders', *Electronics Letters*, **7**, pp. 577–9, 1971.
14. Millar, R. F., 'Rayleigh hypothesis in scattering problems', *Electronics Letters*, **5**, pp. 416–18, 1969.
15. Holford, R. L., 'Scattering of sound waves at a periodic, pressure-release surface: an exact solution', Technical report prepared by Bell Telephone Laboratories, Whippany Road, Whippany, N.J. 07981. Contract N00014-69-C-0074, Office of Naval Research.
16. Nielsen, E. D., 'On the inadequacy of discrete mode-matching techniques in some waveguide discontinuity problems', *IEEE Trans.*, **MTT-18**, pp. 369–71, 1970.
17. Lewin, L., 'On the restricted validity of point-matching techniques', *IEEE Trans.*, **MTT-18**, pp. 1041–1047, 1970.
18. Rumsey, V. H., 'Some new forms of Huygen's principle', *IRE Trans. on Antennas and Propagation*, **AP-7**, pp. S103–4, 1959.
19. Protter, M. H. and Weinberger, H. F., 'Maximum Principles in Differential Equations', (Prentice Hall, Englewood Cliffs, N.J., 1967).
20. Wilton, D. R. and Mittra, R. A. J. 'A numerical approach to the calculation of electromagnetic scattering properties of two-dimensional bodies of arbitrary cross-section', *IEEE Trans on Antennas and Propagation*, **AP-20**, pp. 310–7, 1972.

Corrections

The following corrections should be made to the paper 'Designing a television line flywheel generator using a phase-locked loop integrated circuit', published in the November 1972 issue of *The Radio and Electronic Engineer*:

Page 484, equation (17) *should read*

$$\theta_p(t) = \Delta\theta\left(\cos(\omega_n t\sqrt{1-\zeta^2}) - \frac{\zeta}{\sqrt{1-\zeta^2}} \times \times \sin(\omega_n t\sqrt{1-\zeta^2})\right)\exp(-\zeta\omega_n t) \quad \ldots(17)$$

In addition the symbol θ_i *should read* θ_t in Fig. 5, Fig. 12 and on page 487, line 10.

U.D.C. 621.376.56:621.395:621.397

An experimental differential p.c.m. encoder-decoder for Viewphone signals

G. A. GERRARD, *(Graduate)* *

and

J. E. THOMPSON, Ph.D., B.Sc.(Eng.), D.I.C. *

Based on a paper presented at the IERE Conference on Digital Processing of Signals in Communications held in Loughborough from 11th to 13th April 1972.

SUMMARY

A fully digital differential pulse code modulation (d.p.c.m.) system has been constructed to provide a flexible and stable basis for subjective optimization of a technique for efficient coding of video telephone picture signals. The experimental encoder accepts a precoded 8-bit p.c.m. input and delivers a 4-bit d.p.c.m. output which may be corrupted by simulated channel errors before decoding and display. The implementation of the system is described and photographic results of the coding technique are presented.

* *Post Office Research Department, Dollis Hill, London NW2 7DT.*

1 Introduction

Proposals for a video telephone service have stimulated an accelerating interest in the application of redundancy removal techniques to television coding. The nature of the picture signal and the probable need for tandem switching provide strong arguments in favour of digital transmission for distances beyond about ten miles if acceptable overall performance is to be achieved economically. Extensive penetration of such a service will therefore depend upon the development of wideband systems which are particularly suitable for digital transmission. Direct p.c.m. encoding of the 1 MHz bandwidth television signal requires at least 7 bits/sample for adequate quality but this is a highly redundant code for the envisaged applications and subjective requirements of the video telephone. More efficient coding methods are being studied to reduce the gross digit rate to a fraction of the figure required by p.c.m.

One such encoding technique is that of 'differential p.c.m.', patented by Cutler[1] in 1952 and currently being applied by A.T.&T. to the digital transmission of Picturephone.[2] The technique belongs to a general class of predictive encoders for which numerous theoretical and computer-simulated optimizations have been attempted during the previous two decades. However, these assumed necessarily simple models of the human visual system to describe the relative subjective importance of various coding degradations. Very recently, the advancement of digital circuit technology has allowed the implementation of a 'real time', completely flexible experimental system which will facilitate an empirical optimization of the various coding parameters in accordance with the complex and ill-understood fidelity criteria of the human viewer.

This paper describes an all-digital differential p.c.m. system designed to process video telephone signals. The current experimental parameters for the U.K. Post Office system, 'Viewphone', are given in Table 1.

Table 1

Tentative picture parameters for Viewphone system

Line scanning frequency (horizontal)	8 kHz
Number of lines	319
Field scanning frequency	50·15 Hz
Interlace	2 : 1
Aspect ratio	11 : 10
Transmission bandwidth (nominal)	1 MHz

2 Differential P.C.M.

The three essential features of differential p.c.m. (d.p.c.m.) are the operations of differentiation, tapered quantization and integration. The integration process at the receiver is complementary to the differentiation process at the transmitter. Because the system transmits quantized differences, it is possible for the receiver's integrator to accumulate quantizing error unless the quantizer is placed within a feedback loop at the transmitter. Differentiation is therefore performed by feedback around an integrator and subtractor and the quantizer is included in the loop as shown in Fig. 1. Thus the quantized difference which is transmitted is not the difference between two input picture samples but is the difference between a new picture sample and the accumulation of all the quantized differences previously sent to the receiver.

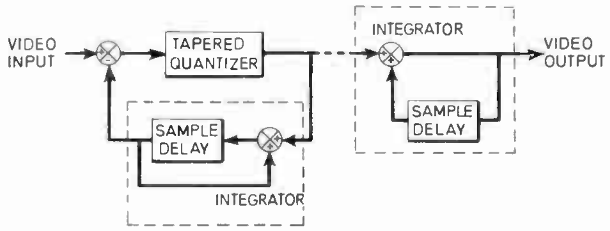


Fig. 1. Conceptual diagram of a differential encoder/decoder.

The subjective justification for d.p.c.m. is that the eye is particularly critical of noise and quantizing contours in the low detail (i.e. gradually changing) regions of a picture while considerable noise and amplitude distortion can be tolerated on samples in detailed regions and at edges and boundaries. This phenomenon is probably connected with the cognitive organization of the visual system which attaches particular significance to the edges and boundaries of an image causing their subjective brightness distributions to appear peaked relative to the objective luminance distributions (known as the Mach phenomenon—see Ref. 3, for example). With d.p.c.m. the combined operations of differentiation followed by tapered quantization have the effect of separating the areas of low and high detail and of quantizing those areas accordingly. In low detail regions where the sample differences are small, the system operates at the centre of the tapered quantizer characteristic and makes suitably small quantizing errors. As picture detail and the sample difference amplitude increase, quantizing errors are increased proportionately. Optimum use can therefore be made of a restricted number of quantizing levels by adjusting the inner thresholds to minimize granularity (i.e. noise) and contouring in low detail areas, while compromising in making the outer levels as large as possible to reduce the effect known as 'slope overload'.

The d.p.c.m. system transmits samples describing the instantaneous slope of the picture signal so that coarse quantization has the effect of restricting the accuracy with which the system output can follow a rapidly changing input signal. Typically, for a 3 bits/sample (i.e. 8 quantization levels) d.p.c.m. system, the output levels of the tapered characteristic might be $\pm 2\%$, $\pm 8\%$, $\pm 14\%$, $\pm 30\%$ of the peak input video amplitude. Thus this system would need about three sample periods to construct a sudden black-to-white transition of maximum amplitude and would perceptibly blur such an edge in the picture. Of course, for input transitions of just less than 30% the rise-time would be preserved and for some edges the system may even overshoot.

In accordance with the statistical concept of signal coding, the d.p.c.m. system predicts that each sample of the television signal will be equal to the previous one and merely transmits to the receiver, sample by sample, the amount by which its prediction is in error. Prediction using other than the previous sample has been proposed but it has been shown (see Ref. 4 for example) that within a television scan line there is negligible advantage in using more than the previous sample.

With a restricted data rate and with the available levels adjusted to minimize granularity and contouring, the d.p.c.m. system blurs vertical and near-vertical edges in the picture. If the 'sample delay' in Fig. 1 is replaced by one television line-scan period, then previous line prediction may be realized which behaves similarly except that blur now occurs on horizontal and near-horizontal edges in the picture. The blur is slightly worse than for previous-sample prediction since the adjacent sample in the previous line is spatially further removed (being twice the line pitch away because of interlace) than the previous sample in the same line; however, for a particular picture having a predominance of vertical edges the picture quality is far superior using previous-line prediction to that using the previous sample. To cope with all types of pictures, an efficient system can be constructed[5] using a combination of previous-line and previous-sample prediction as shown in Fig. 2. Slope overload now occurs predominantly on diagonal edges but picture quality is in general superior to that using either prediction singly.

'Two dimensional' prediction has an advantage in respect of channel error performance—an inherent difficulty of all differential coders which have integrating receivers and which therefore perpetuate channel errors. With a one-dimensional prediction it is essential to provide the d.p.c.m. integrator with a 'leak' or a periodic reset to limit the period for which an error is propagated, but the streak caused by an impulsive error is much more visible than the single sample error which would occur in conventional p.c.m. Obviously, the streak can be made shorter by increasing the integrator leak but the advantages of differential coding are then increasingly lost and objectionable 'leak contouring'[2] and noise are generated

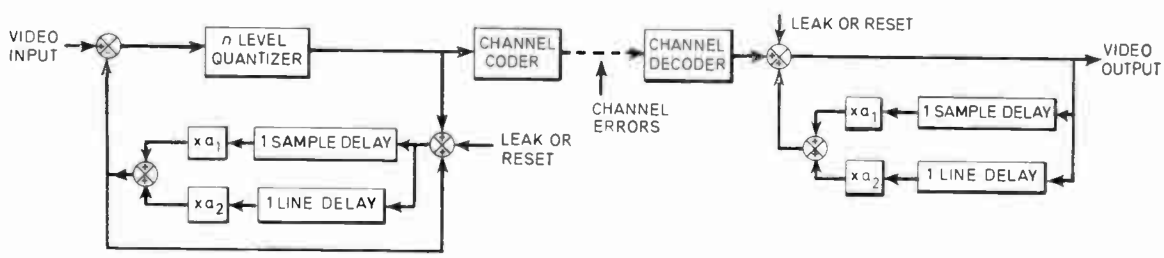


Fig. 2. Block diagram of a differential p.c.m. system using two-dimensional prediction.

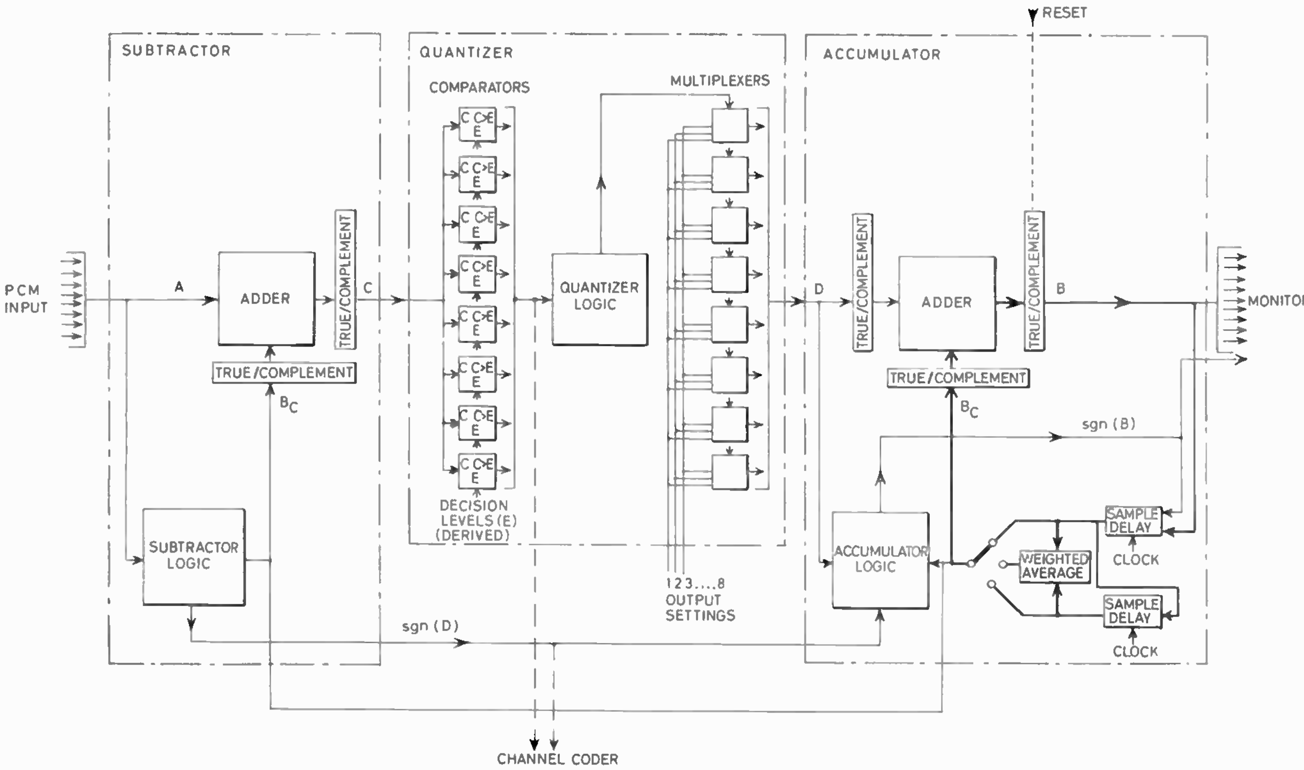


Fig. 3. Block diagram of the digital d.p.c.m. encoder.

in the picture. To minimize such quantizing effects, a long integrator time-constant is required but then errors are allowed to propagate to the end of a line or field before they can be cleared from the accumulator during the blanking intervals. With two-dimensional prediction however, as in Fig. 2, an impulsive error is attenuated by a_1 or a_2 each time it recirculates within the accumulation loop. Thus in a simple system where $a_1 = a_2 = \frac{1}{2}$, the error decays exponentially both horizontally and vertically and, except for those errors which occur near the end of a line or field, no additional benefit is achieved by periodically resetting the accumulators. The visibility of an error is thus considerably reduced over the streak which characterizes one-dimensional prediction.[5]

3 The Experimental Encoder

The experimental d.p.c.m. system described in this paper was constructed to facilitate subjective optimization of the parameters indicated in Fig. 2. The number of quantization levels (n) and their values were designed to be fully variable from the front panel to allow investigation of the compromise between contouring or granularity and slope overload, and any dependence of these values upon picture content. With two-dimensional prediction, the respective weights a_1 and a_2 assigned to the samples in the same and previous lines should perhaps be chosen according to the sampling pitch and the line pitch respectively. The effects of varying these parameters on picture quality and on channel error performance are of interest.

One of the most important parameters is the integration leak time-constant. This determines not only the rate at which the effect of a channel error decays in the received image, but also the ease with which mistracking between the integration processes at the receiver and transmitter can be minimized. Mistracking can arise either from differing responses of the two integrators or from inaccuracy in the assignment of values to the quantizer output levels (particularly the smallest ones). Clearly, the longer the integration time-constant, the more serious mistracking becomes. For these reasons it appears advantageous[6] to employ a digital accumulator rather than a continuous integrator since this effects an infinite time-constant and facilitates precise classification and integration of the quantizer output levels. Channel error effects can be restricted if necessary by resetting the accumulator or by a progressive digital decrement.

Limb[6] described a system in which only the integration process was performed digitally, a digital-analogue converter being inserted in the feedback loop before the subtractor. However, to take advantage of recent advances in m.o.s. and t.t.l. technology and in view of the amount of flexibility coupled with precision which is required, it was decided to realize the system of Fig. 2 entirely digitally. Figure 3 indicates the digital functions performed in the encoder in accordance with the system diagram of Fig. 2. The transmitted signal is derived from the comparator outputs, as indicated by the broken lines in Fig. 3, and encoded into a reduced data rate by the channel coder. At the receiver, after a complementary channel decoder, the d.p.c.m. decoder comprises a duplicate of the accumulator unit shown in Fig. 3. In the absence of channel errors, a replica of the receiver output signal can be monitored at the output of the encoder accumulator.

The experimental encoder was constructed using medium scale integration (m.s.i.) arithmetic and storage units to accept a precoded 8-bit p.c.m. input. This gives a margin of accuracy over the minimum coding resolution of 7 bits/sample which is required to avoid visible contouring on particularly critical test pictures.

3.1 *Quantization*

Previous experiments with a 3 bits/sample d.p.c.m. system had suggested that adequate picture quality might be achieved using 4 bits/sample. Allowance was therefore made for 16 variable output levels with the facility for introducing an additional 17th level at zero if required. For most pictures of interest, the distributions of positive and negative sample differences are similar and it is reasonable to assume a symmetrical quantizing characteristic—thus halving the required number of controls. The number can be further halved by deriving the quantizer decision levels from the output representative levels automatically. Max[7] presented rules for designing a quantizer with minimum error and showed that the assumption of a law of quantizing error visibility which increases monotonically with error amplitude implies that the error is minimized when the decision levels lie midway between neighbouring output levels. The decision levels can therefore be derived by averaging adjacent pairs of output level settings.

In the experimental encoder (Fig. 3) therefore, eight knobs directly control the quantizer output levels, each level appearing as an 8-bit word wired on the parallel contacts of a corresponding eight-pole wafer switch. Adders derive 9-bit decision levels by adding adjacent pairs of output words and shifting by one bit, and these are applied to the respective 'E' inputs of eight 9-bit comparators. The 8-bit output of the subtractor unit is applied in parallel to all 'C' inputs of the comparators and their '$C > E$' outputs are logically combined to operate a multiplexer which selects the appropriate 8-bit output setting. An additional output level can be switched in at zero and any number of quantal levels below 17 can be achieved by progressively disabling comparators. The sign of each sample difference (C) is transmitted directly from the subtractor unit to the accumulator. The decimal value assigned to each switch position can be read from three numerical indicators.

3.2 *The Accumulator*

The accumulator performs the function of algebraically adding the incoming difference word D and the word B_c stored in the digital delay connected between the output and one of the inputs of the adder. The words D, B_c and B represent magnitude information only. The sign associated with D and B_c is conveyed by a sign bit fed to the accumulator logic to determine the sign of the output word B and control the true/complement units connected to the adder. If both D and B_c are positive then addition must be carried out and the accumulator logic instructs the true/complement units to pass the true or unaltered word. If however D and B_c are of opposite sign then subtraction must be performed and this done by complementing the largest word and the output of the adder. The complement produced is the ones complement, obtained by changing each digit of the word.

The algorithm used in the accumulation process produces true magnitude information at the output B of the accumulator, sgn(B) only being used to accommodate negative-going video overshoots. Also, although the accumulator input word D has only 8 magnitude bits, a 9th bit is introduced in the accumulator to accommodate any positive going video overshoots.

For d.p.c.m. with previous sample prediction, the delay around the accumulator adder consists of D-type flip-flops clocked at the sampling rate to delay B by one clock pulse period, giving B_c. For previous line prediction, 256-bit m.o.s. shift registers are connected in series with the D-type flip-flops and the clock frequency is made 257 times the Viewphone line-scanning frequency to produce one line period delay. The predictions may also be combined as indicated in Fig. 3, and a digital switch allows rapid comparison between coding with the three alternative predictions. The true/complement unit at the output of the accumulator adder has the facility for setting its output to zero so that both the transmitter and receiver accumulators may be cleared periodically to restrict the effects of channel errors.

3.3 *The Subtractor*

The subtractor unit determines the difference between each 8-bit video input word 'A' (always positive) and the 10-bit (magnitude and sign) prediction B_c. The subtractor logic determines the sign of '$A-B$' which will be associated with the respective quantizer output (D), and also controls the subtraction algorithm. If B_c is negative, addition must be performed and both true/complement units pass the true words. If however, B_c is positive, then subtraction must be performed and two cases have to be considered: (i) if $A > B_c$, then a logical '1' is fed to the carry input of the adder and B_c is complemented; (ii) if $B_c \geqslant A$ then both B_c and the output of the adder are complemented. The output word C has 8 amplitude bits and covers the peak–peak video range. This is adequate during normal operation but additional logic is incorporated to avoid latch-up conditions at switch-on.

3.4 *Analogue/Digital Converter*

The 8-bit a.d.c., which generates the digital input to the d.p.c.m. encoder, is a hybrid encoder[8] comprising two cascaded 16-level quantizers. The quantizing error between the input and output of the first stage is amplified 16 times and presented to the second set of comparators. The four most significant digits are clocked from the first stage simultaneously with the four least significant digits from the second stage and the entire conversion is carried out within one sample interval. The negative tips of the picture synchronizing pulses are clamped to the lower end of the coding range between the sample and hold circuit and the converter input.

In order to use readily available 256-bit shift registers as the delay elements for previous line prediction, it was necessary to use a sampling frequency with the appropriate harmonic relationship to the 8 kHz line scan frequency. This was achieved by conventional phase lock

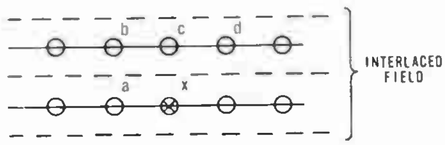


Fig. 4. Diagram of picture elements near the present sample (x) which may be used for prediction.

principles using a programmable divider. Variation of the dividing ratio can be used to alter the position on the previous line of the predicted sample (points b, c or d in Fig. 4).

4 The Decoder

The decoder comprises multiplexers and an accumulator identical to those of the encoder. In the present system, the quantizer logic of the encoder and the multiplexers of the decoder correspond to the channel coder and decoder respectively shown in Fig. 2. The quantizer output settings are carried across to the decoder by wires so that both may be adjusted simultaneously by the front panel controls of the encoder.

The four bits transmitted to the decoder comprise the 3-bit word from the output of the quantizer logic and the 'sgn(D)' bit. Channel errors are simulated by passing the bits through four 2-input 'exclusive-OR' gates (see Fig. 5), the other input to each gate being connected to an error generator. A logical '1' from the error generator causes the bit to be inverted.

4.1 *Pseudo-random Error Generator*

Errors which might be incurred during serial transmission of the 4-bit d.p.c.m. signal (about 8 Mb/s) can be generated pseudo-randomly according to the output of a multiple-input AND gate which monitors the states of a 33-stage feedback shift register generator clocked at the sampling frequency of 2 MHz (see Fig. 5). For example with an 8-input AND gate programmed to recognize a particular pattern of eight 1's and 0's, the generator produces a 1 rather than 0 with probability 2^{-8} or

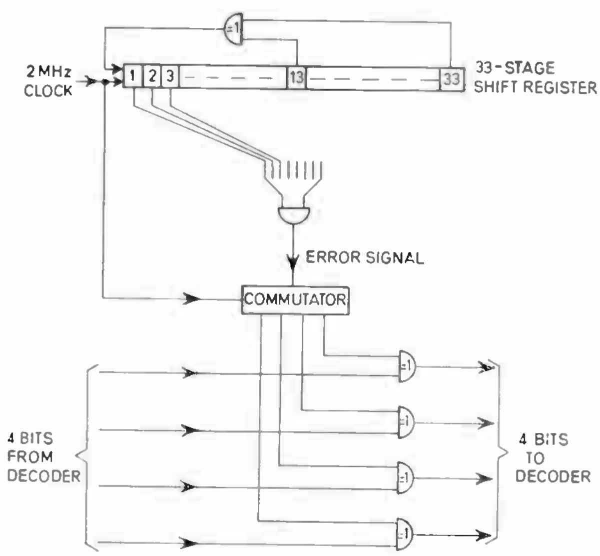


Fig. 5. Diagram of error generation and insertion.

$3{\cdot}9\times10^{-3}$. This error signal is then commutated between the four bits of the d.p.c.m. signal thus simulating a channel error probability of $0{\cdot}98\times10^{-3}$, or approx 1 in 10^3, with the restriction that no more than one error can occur in each sample. A 'reasonable' error distribution can be achieved by intelligent choice of the 8-bit pattern sought by the AND gate and error rates sufficiently close to powers of 10 in the range 10^{-1} to 10^{-8} are obtained by appropriate control of the number of inputs to the AND gate in the range 1 to 25. The complete error sequence repeats every $2^{33}-1$ clock periods, or approximately 72 minutes.

4.2 *Digital/Analogue Converter*

The effects of sampling and differential quantization can be observed by decoding the signal 'B' at the encoder's accumulator output ('monitor' point indicated in Fig. 3) since in the absence of channel errors this would be identical to the receiver output. Additionally, the effect of channel errors can be observed by d./a. conversion of the 'B' signal output of the receiver's accumulator. B is not immediately in a suitable form for d./a. conversion since it is in symmetrical binary number code (i.e. magnitude and sign). However code conversion to binary is easily achieved by complementing the magnitude digits and adding one to the resulting binary number whenever the sign digit of B is '0'. The ten parallel output digits of the code translator are reclocked before connexion to a voltage-switched thin-film ladder network and output amplifier. Due to code translation, the video emerges offset by the d.c. magnitude of the 10th (most significant) digit.

5 Results

A selection of photographic results is presented to illustrate the dependence of picture quality on the coding parameters. It should be noted that the photographs (Figs. 6 and 7) do not show the true aspect ratio of the Viewphone image since they only show part of the 11 : 10 raster.

For these results, the video input to the a./d. converter was band-limited by a phase equalized low-pass filter which is substantially flat to 1 MHz and provides a minimum of 20 dB stop-band attenuation. Its 3 dB and 20 dB points are 1·06 and 1·13 MHz respectively. The output of the d./a. converter was passed through the experimental Post Office Viewphone shaping filter which provides a minimum stop-band loss of 30 dB and has 3 dB and 20 dB points at 700 kHz and 1000 kHz respectively.

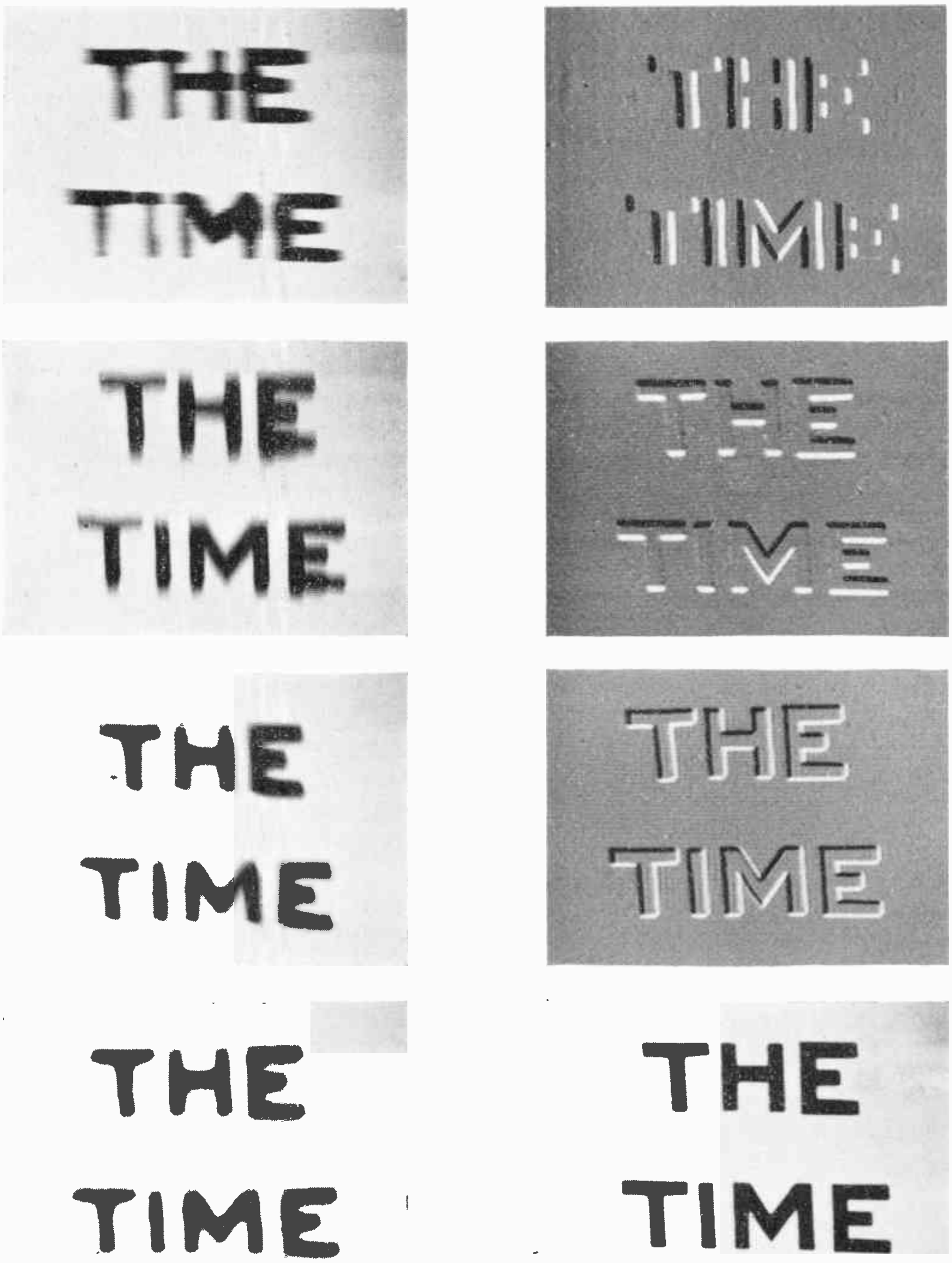
For ease of illustration, slope overload can be exaggerated by disabling the outer four levels of the quantizing characteristic, i.e. by reverting to 3-bit d.p.c.m. with non-optimal level spacing. Using the notation of Fig. 4 to denote the samples used for previous-sample (a) or previous-line (b, c or d) predictions, Figs. 6(i) and (iii) show severe slope overload on vertical and horizontal edge components using predictions 'a' and 'c' respectively. For previous-line prediction, it was found that while 'b' gives improved rendition of edges with negative slope and 'd' gives best performance on positive slopes, 'c' gives

(i) (ii)

(iii) (iv)

(v) (vi)

(vii) (viii)



3-bit d.p.c.m.
(i) previous element prediction 'a'
(iii) previous line prediction 'c'
(v) average prediction $\frac{1}{2}(a-c)$
(vii) average prediction $\frac{1}{2}(a+d)$

4-bit d.p.c.m.
(ii) 'a' difference signal
(iv) 'c' difference signal
(vi) $\frac{1}{2}(a+c)$ difference signal
(viii) average prediction $\frac{1}{2}(a+d)$

Fig. 6. Illustrations of slope overload.

the best overall performance on 'typical' pictures. Combining the predictions 'a' and 'c', Fig. 6(v) shows the improvement in picture quality obtained by sharing the slope overload between two dimensions. This improvement might be argued intuitively to be about 3 dB (or half a bit) as the orthogonal half-components of slope overload contribute approximately equally to the overall impairment (the eye being relatively isotropic) and combine according to a root-sum-of-squares law.

With two-dimensional prediction, it is found that 'c' no longer gives the best average contribution from the previous line. Figure 6(vii) shows that $\frac{1}{2}(a+d)$ gives the

(i)

(ii)

(iii)

4-bit d.p.c.m. with error rate 1 in 10^4 using:
(i) previous element prediction 'a'
(ii) previous line prediction 'c'
(iii) average prediction $\frac{1}{2}(a+c)$

Fig. 7. Channel errors.

appearance of a sharper picture overall with more symmetrical blurring than $\frac{1}{2}(a+c)$—note in particular the symmetry of the letter 'M'—though it does exhibit a weakness for edges of intermediate negative slope. As stated, the previous pictures were obtained with non-optimal 3-bit d.p.c.m. in order to illustrate the orientation and relative severity of slope overload. Figure 6(viii) shows the high quality achievable using 4-bit d.p.c.m. with $\frac{1}{2}(a+d)$ prediction while Figs. 6(ii), (iv) and (vi) further illustrate slope overload with the 4-bit prediction error signals (D in Fig. 3) corresponding to the various predictions.

Figure 7 shows single-frame photographs (taken by means of a digital frame store) which typify the performance of the various types of prediction with 4-bit d.p.c.m. under conditions of channel error probability 10^{-4}. For predictions 'a' or 'c', streak duration is restricted by clearing the accumulators during the line synchronizing pulse and for a whole line during the field blanking period, respectively. Note that the vertical streaks for prediction 'c' are broken since they exist only in the interlaced lines of one field. The 'comet-tail' error patterns of $\frac{1}{2}(a+c)$ decay without the need for reset in the approximate direction of $-45°$ whereas the alternative prediction $\frac{1}{2}(a+d)$ skews the comet tails around to be nearly vertical. While $\frac{1}{2}(a+d)$ may be better than $\frac{1}{2}(a+c)$ in respect of slope overload performance, there is probably no significant difference in the visibility of the error patterns and only $\frac{1}{2}(a+c)$ is shown in Fig. 7(iii). This shows that the effects of errors are considerably reduced in visibility by two-dimensional prediction although single-frame photographs exaggerate the improvement since the comet tails are more easily confused with picture detail than the corresponding one-dimensional streaks. Even with two-dimensional prediction, preliminary tests indicate that an error rate better than 10^{-6} appears necessary for satisfactory Viewphone picture quality unless some additional form of error concealment is used.

Clearly, still photographs cannot give a true impression of the visibility of temporally-varying impairments such as granular noise, edge busyness and error patterns, though they are useful in qualitatively illustrating slope overload. The quantizing characteristic used in generating these pictures was found after considerable practice at empirically compromising between edge effects and noise, and was found to give good results on most pictures. The output levels were set at ± 1, 3, 6, 16, 27, 40, 64 and 92 steps out of the 256 corresponding to the full 8-bit coder range. In practice, the maximum signal change which can occur between samples on a black-white transition after the input low-pass filter is 156 steps, and if a 2 dB margin is allowed on the signal input to the a./d. converter against the possibility of overload then the maximum change is only 124 steps.

Larger scale subjective experiments are currently being devised to determine a more confident estimate of the optimum quantizing characteristic and also to allow deduction, from data on how the optimum characteristic changes with different pictures, of suitable adaptive strategies which might further reduce the data rate required for acceptable Viewphone picture quality.

6. Conclusion

Excluding the a./d. converter, the described encoder comprises more than 150 integrated circuits and the decoder 75, there being a total of 102 m.s.i. arithmetic units. The total propagation delay of the various circuits in the d.p.c.m. feedback loop imposes a maximum signal sampling rate of 2·5 MHz. The codec fulfills all the requirements of an experimental system and is currently being used to quantify the performance of differential coding. Its fully digital implementation offers the following advantages:

(i) Digital accumulation allows precise classification of the quantizer output amplitudes and ideal tracking of the receiver and transmitter accumulators. Even though it is more difficult to realize a digital leak, the preferred reset method of limiting channel errors is potentially simpler than with analogue integration. Furthermore, parity check and error concealment techniques are more easily implemented in digital form.

(ii) A digital input to the 'quantizer' simplifies the realization of fully variable output levels with automatic derivation of thresholds. The use of analogue threshold devices requires either precise setting up for each chosen set of output levels[6] or, to obtain automatic derivation, a d./a. converter would have to be attached to each output level.

(iii) Thresholding of an analogue signal is the only stage at which care is needed to ensure sufficient speed and accuracy. By taking the a./d. conversion out of the d.p.c.m. loop, the conversion time can be considerably increased, and more time is available in the loop for complex prediction algorithms.

7 Acknowledgments

The authors gratefully acknowledge the skills and persistence of M. Burl and E. Weremuik who constructed most of the encoder and the decoder respectively, and of A. M. Hill in the construction of the error generation and insertion logic. Valued criticism of the paper was given by I. F. Macdiarmid.

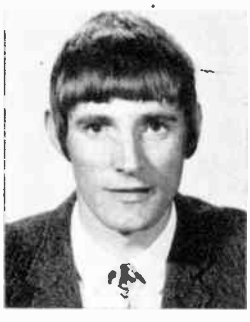
Acknowledgment is made to the Director of Research of the Post Office for permission to publish this paper.

8 References

1. Cutler, C. C., 'Differential Quantization of Communication Signals', U.S. Patent 2,605,361, 29th July 1952.
2. Millard, J. B. and Maunsell, H. I., 'Digital encoding of the video signal', *Bell Syst. Tech. J.*, **50**, No. 2, pp. 459–79, February 1971. (Special issue on the Picturephone system).
3. Cornsweet, T. N., 'Visual Perception', (Academic Press, London 1970).
4. O'Neal, J. B. Jnr., 'Predictive quantizing systems (differential p.c.m.) for the transmission of television signals', *Bell Syst. Tech. J.* **45**, No. 5, pp. 689–721, May–June 1966.
5. Connor, D. J., Pease, R. F. W. and Scholes, W. G., 'Television coding using two dimensional spatial prediction', *Bell Syst. Tech. J.*, **50**, No. 3, pp. 1049–61, March 1971.
6. Limb, J. O. and Mounts, F. W., 'Digital differential quantizer for television', *Bell Syst. Tech. J.*, **48**, No. 7, pp. 2583–99, September 1969.
7. Max, J., 'Quantizing for minimum distortion', *IRE Transactions on Information Theory*, IT–6, pp. 7–12, March 1960.
8. Devereux, V. G., 'Digital television: a.d.c.'s, d.a.c.'s and preferred parameters for coding', International Broadcasting Convention, London, 7–11 September 1970 (IEE Conference Publication No. 69, pp. 69–70).

Manuscript first received by the Institution on 6th January 1972 and in final form on 20th December 1972. (Paper No. 1510/Com. 63.)

The Authors



Dr. J. E. Thompson received the B.Sc.(Eng.) degree in electrical engineering from Imperial College, London, in 1964 and the Ph.D. Degree in 1968 for a thesis on television data compression at the same College. In 1968 he joined the Visual Telecommunications Branch of the Post Office Research Department where he currently has responsibility for investigation of digital television systems and coding for Viewphone.

Mr. G. A. Gerrard (Graduate 1960) was a Research Apprentice at EMI Ltd from 1954 to 1959 and during this period received his technical education at Southall Technical College and Borough Polytechnic, London. Initially he worked in television and later in radar at EMI and he subsequently moved to the European Organisation for Nuclear Research (CERN) in Geneva, Switzerland, where he was concerned with instrumentation for experiments in nuclear physics. In 1967 Mr. Gerrard took up his present appointment of Experimental Officer at the Post Office Research Department where he works in the Visual Telecommunications Branch.

UDC 621.376.24: 621.396.621.52

Direct conversion s.s.b. receivers:

a comparison of possible circuit configurations for speech communication

S. R. AL-ARAJI, M.Sc., Ph.D.*

and

Professor W. GOSLING,
B.Sc., C.Eng., F.I.E.R.E.†

SUMMARY

Direct conversion single sideband receivers are of interest because they are less subject to spurious responses than the conventional superheterodyne, do not entail the use of high-gain narrowband amplifiers with high centre frequency, and lend themselves better to integration in monolithic form.

Two approaches to the design of receivers of this kind have been described: (1) the phasing technique, in which the local oscillator is tuned to the nominal signal carrier frequency and the unwanted sideband response eliminated by phase cancellation; (2) Weaver's 'third method', in which two oscillators are used, the first tuned to the centre of the incoming s.s.b. spectrum, and the second to the centre of the reconstituted a.f. spectrum.

The characteristics of the two types of receiver are compared, and the superiority of the relatively neglected Weaver approach, particularly in a new a.c. coupled variant, is demonstrated, so far as the reception of speech modulated signals is concerned.

* *Formerly at the University College of Swansea; now with the University of Baghdad.*

† *Department of Electrical and Electronic Engineering, University College of Swansea, University of Wales, Singleton Park, Swansea SA2 8PP.*

1 Introduction

Conventional superheterodyne receivers for s.s.b. reception are of the filter type, in which the incoming signal is mixed with the output of a local oscillator at a frequency equal to the sum or difference of the i.f. and the signal carrier frequency. The receiver is made to respond only to one sideband by means of a filter in the i.f. amplifier, the unwanted sideband falling in the stop band of the filter. It is possible to choose either upper or lower sideband response by switching alternative filters into circuit. The wanted sideband is converted, after amplification, to an audio signal by heterodyning with a sinusoid at the i.f. The sideband filter specification is severe, and even with the crystal filters now in universal use a high i.f. is precluded. Thus double or triple superheterodyne configurations are commonplace, particularly at higher signal frequencies.

Multiple conversion also helps to overcome the problems inherent in the use of high-gain narrrow-band i.f. amplifiers, with centre frequencies in the megahertz range. With only a single i.f. it is difficult to prevent circuit instability, particularly with the physically smaller amplifiers now possible using monolithic 'gain blocks'. However, in solving some problems multiple conversion creates others. Even if the increased complexity of the receiver is accepted, the problem of additional spurious responses due to the extra conversion processes has to be reckoned with. In congested radio bands spurious responses have been shown to be the limiting factor in receiver performance.[1]

Thus although the superheterodyne has been brought to a high state of development it appears to be approaching a limit to further advance which prompts a review of alternative approaches to s.s.b. reception. In particular, techniques compatible with monolithic circuit integration would be attractive. One such is the use of direct conversion in which the incoming signal is heterodyned with a sinusoid at or near the carrier frequency, so that the output from the conversion process, after filtering to remove high frequency components, falls within the audio range. A receiver of this type for a.m. has recently been described[2] and simple direct conversion receivers are already in widespread use by radio amateurs.

One of the principal difficulties in applying direct conversion is that, in addition to the response to the wanted sideband, the receiver has a 'second channel' response. For example, if the first oscillator is tuned to the signal carrier frequency, this corresponds to a reponse to the unwanted sideband on the other side of the carrier. In the case of i.s.b. or close-spaced channels such responses will contribute to interchannel cross talk. To eliminate this effect one approach is to build the receiver with two initial mixers, in which the signal is mixed with sinusoids of identical frequency but in phase quadrature. As before, the products are subjected to low-pass filtering, to produce two quadrature outputs within the a.f. spectrum. These outputs will for convenience be referred to in what follows as being at intermediate frequency, since their spectral distribution by no means necessarily corresponds to that of the finally demodulated audio. There are two quite different ways in which

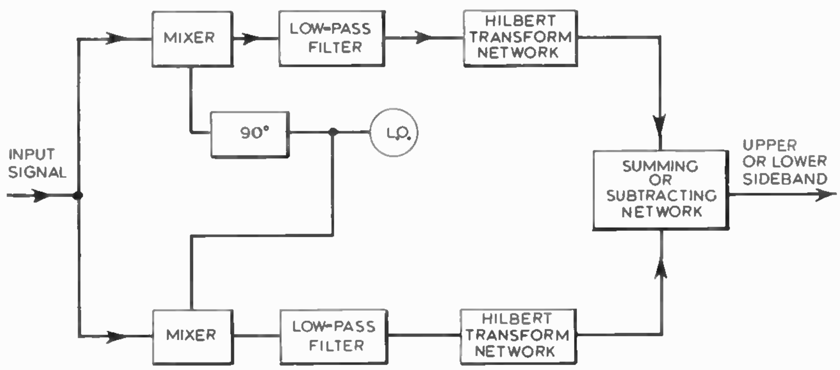
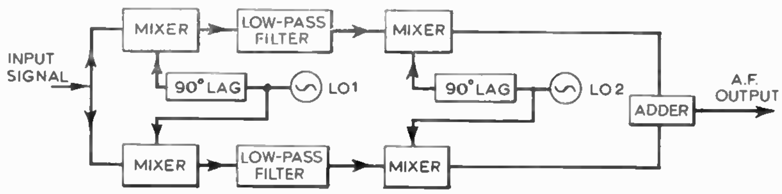


Fig. 1.

(a) An s.s.b. receiver using the phasing method for unwanted sideband suppression.



(b) The Weaver receiver.

these i.f. signals can be combined, to give the required signal response and suppress the unwanted spurious response. These are:

(1) One of the i.f. outputs may be passed through a network which phase-shifts all its components by 90°. (A network of this type is sometimes referred to as a Hilbert transform network.) On adding the output from the phase-shifted channel to that from the other, one of the receiver responses is reinforced and the other approximately cancelled (Fig. 1(a)). This is known as the phasing, or outphasing, method of s.s.b. reception.[3]

(2) The two i.f. outputs may be passed to a second pair of mixers, in which they combine with quadrature outputs from a second local oscillator. The outputs from the second mixing process are then simply added (Fig. 1(b)). This receiver is essentially equivalent to a class of band pass filter proposed by Barber,[4] and the circuit was first suggested for s.s.b. generation by Hall.[5] Weaver[6] drew attention to its possible use as a receiver, and for this reason it is known as Weaver's 'third method' of s.s.b. reception.

A mathematical analysis of the two types of receiver has been given elsewhere[7] and is briefly reviewed in the Appendix.

To restate this result, the Weaver receiver used two identical 'i.f.' channels (actually operating at audio frequency). In each the first conversion uses a local oscillator tuned to the centre of the received s.s.b. spectrum. The centre frequency of the s.s.b. is thus translated to zero, and the lower half of the spectrum is folded over to fall on the upper, in a total i.f. bandwidth extending from zero to half the total bandwidth of the s.s.b. as transmitted.

A second conversion process using an oscillator at the centre frequency of the original modulating audio shifts the zero frequency i.f. components to their correct position, and produces a correct audio spectrum, with, however, a coincident inverted spectrum (that is, one in which high and low audio frequencies have changed places). By the use of two i.f. channels, with the oscillator feed to both first and second conversions in quadrature as between the two channels, two audio outputs are produced. These have the erect spectrum in phase but the inverted spectrum in antiphase. Simple addition of the two outputs will thus result in cancellation of the inverted spectrum, provided that the two i.f. channels are identical. If cancellation is imperfect the result is heard as distortion, but not as adjacent channel interference.

Virtually all direct conversion s.s.b. receivers described hitherto which attempt to suppress the spurious response appear to use the phasing method. The purpose of this communication is to review the properties of the two configurations and to report experimental results obtained with both. Provided that certain simple modifications are made to the basic principle, receivers based on Weaver's third method can show marked superiority over receivers based on phasing.

2 Common Aspects of Receiver Design

Although the phasing and Weaver receivers are different, they have certain features in common. In particular both employ a two-phase local oscillator for the first pair of converters. The available methods of generating quadrature local carriers are different at different frequencies. In the l.f. and m.f. ranges a passive phase-shift network provides a particularly simple solution,[8] but at higher frequencies the component values required become too small if the networks are to operate at realistic impedance levels. A digital technique, in which quadrature square waves are obtained by counting from a clock running at twice the frequency,[9] can be operated successfully up to 30 MHz with presently available Schottky t.t.l., but its disadvantage is the high

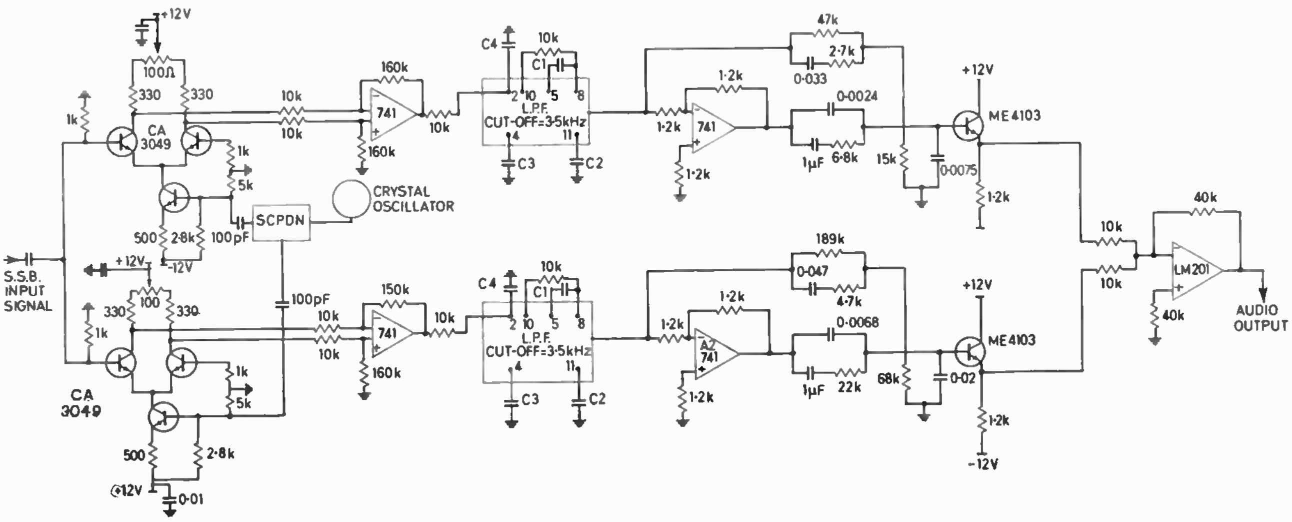


Fig. 2. Circuit diagram of the phasing receiver (SCPDN = servo-controlled phase delay network, LPF = active low-pass filter).

harmonic content of the square waves produced, which enhances mixer harmonic responses. An alternative technique, recently employed at v.h.f.[2] and capable of extension to u.h.f., is to use a phase-shifting network incorporating varactors, which can be electrically servo-controlled to give a phase shift of 90° over a range of frequencies. When, as is often the case with v.h.f. and u.h.f. receivers, only relatively few stations at known frequencies are to be received, and free-tuning facilities are not required, a simple varactor phase shifter is adequate with facilities to switch the biass applied to the varactors.

Design of the mixers is conventional, and the first pair need not be of the balanced type, since components in their outputs which are at radio frequency will be attenuated by the subsequent filters. However, in the Weaver circuit the second pair of mixers must have a high degree of carrier suppression. So far as the noise performance of the mixers is concerned, it should be noted that in a direct conversion receiver the $1/f$ noise components in the mixer output will be within the pass band of the subsequent i.f. filters and amplifiers. To achieve a low mixer noise figure, the low-frequency response of the i.f. amplifiers should be curtailed as much as is consistent with receiver a.f. response, and the mixer design should be based on devices which have relatively little $1/f$ noise, such as junction f.e.t.s.

Both receivers obtain their adjacent channel selectivity by means of low-pass filters, which may be either passive LC or active RC types. The latter are noisier, but compact and cheaper, and are therefore preferred in receivers in which the signal level is above the equivalent filter input noise level. Active filters with from three to nine poles are used to obtain the required adjacent channel rejection, dependent on the channel spacing required.

The i.f. amplifiers operate on a signal in the audio range, thus monolithic amplifiers can be employed, and stability problems are less severe than in amplifiers operating at higher frequencies.

3 Receiver Design and Measurements

Although direct conversion receivers have been constructed successfully for m.f. and h.f. also, it is with a series of v.h.f. designs that the remainder of this paper will be concerned, covering signal frequencies in the range 76–100 MHz. All use the servo-controlled phase-shifter to generate quadrature first oscillator voltages. Only those measurements on the receivers will be reported which facilitate comparison of the properties of the phasing and Weaver receiver types, or which mark points of difference in performance between direct conversion receivers and conventional superheterodynes.

3.1 *The Phasing Receiver*

A receiver of the phasing type was constructed in accordance with the block diagram of Fig. 1(a). The simplified circuit diagram, omitting the pre-mixer r.f. circuits, audio stages and a.g.c. circuits, is shown in

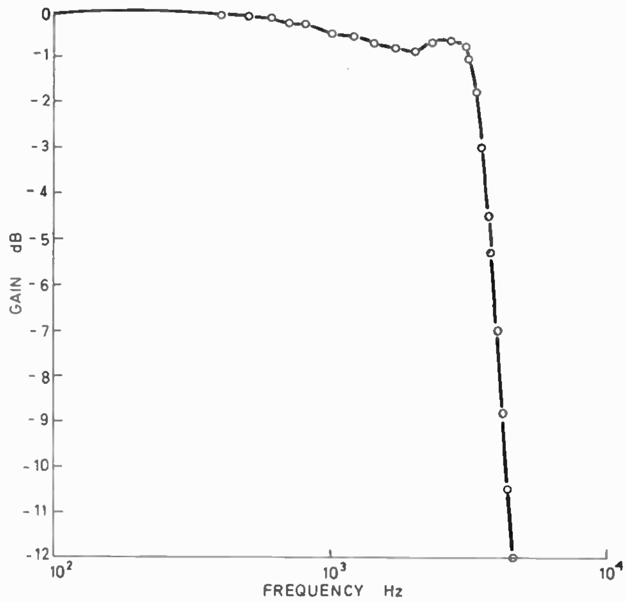


Fig. 3. Measured frequency response of the active low-pass filters used in Fig. 2.

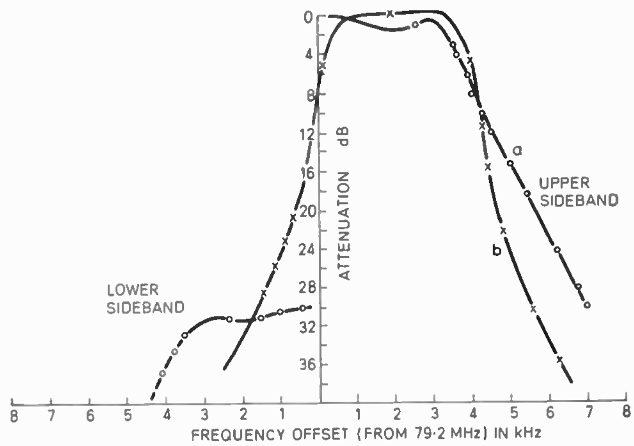


Fig. 4. (a) Frequency response of the phasing receiver.
(b) Frequency response of the Weaver receiver.

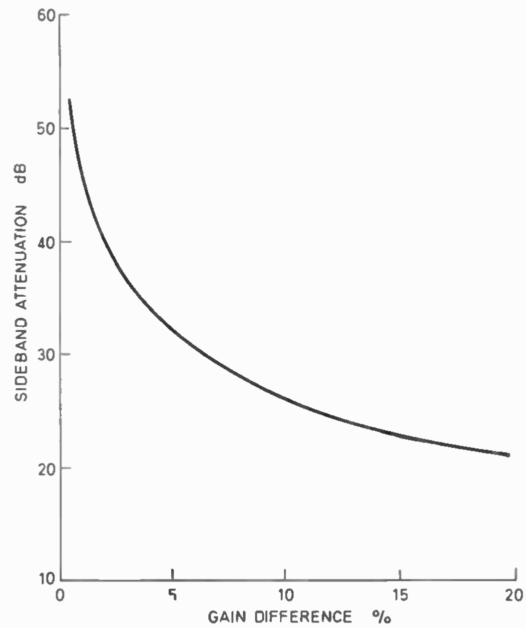


Fig. 5. Unwanted channel suppression as a function of gain difference between the two i.f. channels. (Phasing errors negligible).

Fig. 2. The low-pass filters are four-pole active types based on a modified Sallen and Key circuit, and have a measured frequency response as in Fig. 3. The 90° relative phase shift of one i.f. channel is given by a passive audio wideband phase shifting network of a type long used in s.s.b. exciter design,[10] constructed from resistors and capacitors of 5% tolerance.

The measured frequency response of the receiver is shown in Fig. 4(a), from which it will be seen that the unwanted sideband is suppressed by only some 30 dB. Imperfect suppression is due to three factors:

(a) The two local oscillator voltages at the mixers may not be perfectly in quadrature.

(b) The total gain, including conversion gain, between the common mixer signal input terminal and the output of each i.f. channel may not be identical.

(c) The wide-band 90° phase shift (Hilbert transform) network may not give precisely the right phase shift.

Practically, 30 dB of unwanted sideband suppression is typical of what is readily achieved, and it is only with great difficulty and by individual adjustment of the circuits that this can be raised to 40 dB. Figure 5 illustrates this point by showing the relationship between unwanted channel suppression and gain difference between the two i.f. chains for the case in which the phase angles have been adjusted to be so nearly correct that their effect on channel suppression can be neglected.

This spurious response arises from the channel adjacent to the wanted one, and its effect in the receiver output is thus to contribute interchannel crosstalk. Since the signal in the adjacent channel may be many tens of decibels above the level in the wanted channel, the degree of rejection obtained is not sufficient.

3.2 *The Direct-coupled Weaver Receiver*

A v.h.f. Weaver receiver was constructed in accordance with the block diagram of Fig. 1(b), the simplified

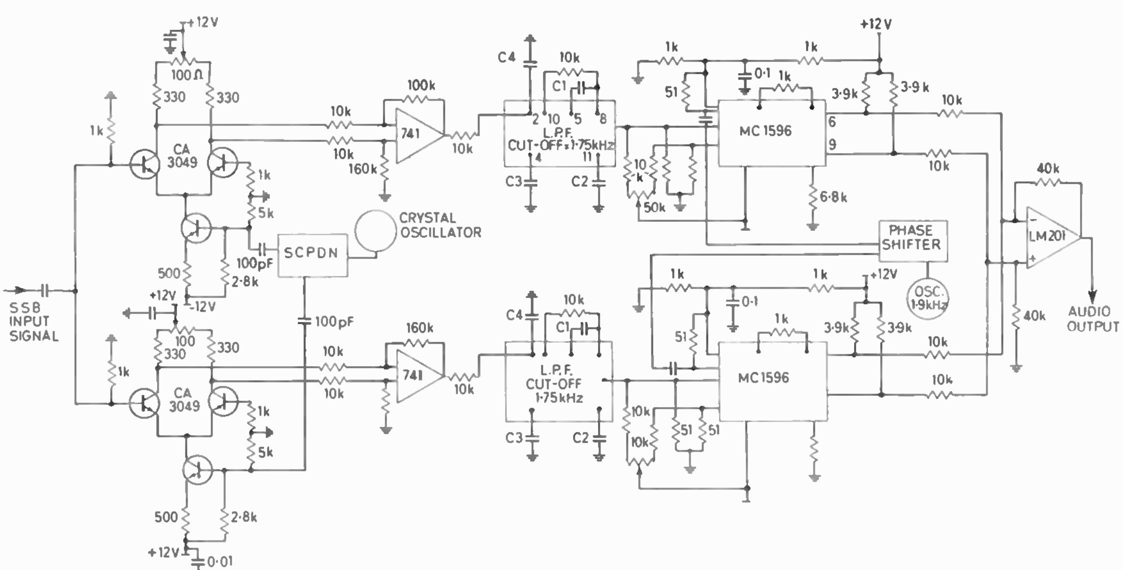


Fig. 6. Direct-coupled Weaver receiver.

circuit of which is given in Fig. 6. The first oscillator, which was crystal controlled at 79·2 MHz, produces two quadrature outputs of equal amplitude, using the servo-controlled phase-shift network. The second local oscillator employs passive phase-shifting networks to give a two phase output at 1·9 kHz. The cut-off frequency of the low-pass filters is reduced to 1·75 kHz, resulting in an audio pass-band of 1·9 ± 1·75 kHz, or from 150 Hz to 3·65 kHz.

The selectivity curve is shown in Fig. 4(b). The response is symmetrical about the frequency to which the first oscillator is tuned. Rejection of adjacent channels depends only on the low-pass i.f. filters, and can be further improved by using filters of higher order.

In the present case, both the phasing and Weaver receivers used a four-pole filter of similar design (but different cut-off frequency) to afford a fair basis of comparison. Had, for example, an eight-pole filter been used, the stop-band rejection would have doubled in the Weaver case, whilst in the phasing receiver no improvement in rejection of the unwanted sideband would have resulted. It should also be noted that the cut-off frequency of the filters in the Weaver receiver is about one octave lower than in the phasing receiver, thus well away from the pass band the response of the Weaver receiver will be up to $6n$ decibels less than that of its competitor, where n is the order of the filters used.

A further advantage of the Weaver configuration lies in the nature of its spurious response. When the receiver is correctly tuned, with the first oscillator frequency at the centre of the received s.s.b. spectrum, the spurious response observed in the audio output of the receiver is simply the wanted signal with its spectrum reversed. In the case of the receiver described, an audio component due to the wanted signal at a frequency of $(1{\cdot}9+f)$ kHz will give rise to a spurious output (of reduced amplitude) at a frequency of $(1{\cdot}9-f)$ kHz. The spurious response is not inter-channel crosstalk, as in the phasing case, but rather an unusual form of audio distortion. Subjectively, a high level of distortion of this kind can be tolerated without loss of intelligibility: and a ratio of 20 dB between the wanted and spurious responses would probably be acceptable. In practice, the receiver described achieves a ratio in excess of 30 dB. Suppression depends on correct phasing of the local oscillators, which is fairly easily achieved, and equality of gain between the two i.f. channels, which presents more difficulty. Figure 5 applies unchanged in this case, and indicates that good suppression depends on balance between the two channels which is not beyond what can readily be achieved.

A more difficult problem arises because the i.f. amplifier chain is d.c. coupled. This is necessary if all the transmitted frequencies are to be reproduced, since a component of the received signal which is at the centre of the s.s.b. spectrum, and coincides with the first oscillator frequency, will give rise to a d.c. component in the output of the first mixers. The signal level at this point may be only tens of microvolts, setting a very difficult d.c. drift specification for the mixer and following amplifiers. Spurious d.c. output from the i.f. amplifiers due to drift will, after the second frequency conversion, result in an audio output at the frequency of the second oscillator. In the case of the present receiver this is at 1·9 kHz, and is heard as an unpleasant whistle. Although the required drift specification can now be approached using the technique of hot-substrate monolithic amplifiers, drift is a serious disadvantage of the direct-coupled Weaver receiver, and may be the main reason for its relative neglect hitherto.

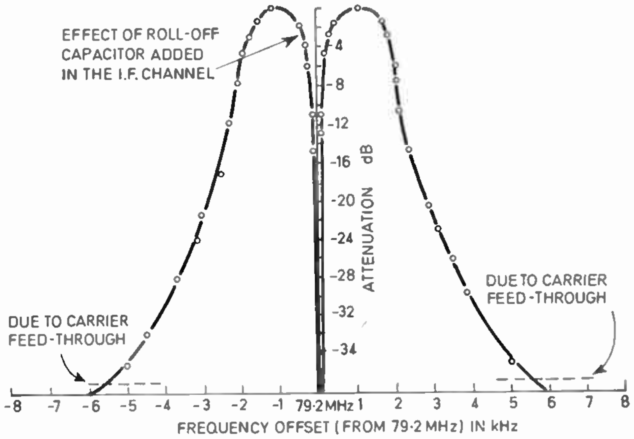


Fig. 7. Frequency response of the a.c.-coupled Weaver receiver.

3.3 *The A.c.-coupled Weaver Receiver*

Because of the difficulties caused by drift, attention was given to the use of a.c. couplings in the i.f. channels of the Weaver receiver. The circuit used is identical to that of Fig. 6, except that the second pair of mixers are each preceded by an a.c. coupling. The result is a receiver having a selectivity curve as in Fig. 7. The pronounced stop band or 'hole' in the middle of the receiver response curve, which has a half-power width of 500 Hz, results from the a.c. couplings.

It is well known that the introduction of a stop band in the middle of the speech spectrum has little effect on intelligibility, provided that it is not more than a few hundred hertz wide. A computed plot of stop-band width against random word and sentence intelligibility is shown in Fig. 8, based on data from French and Steinberg[11] and Fletcher.[12] A wide stop-band causes

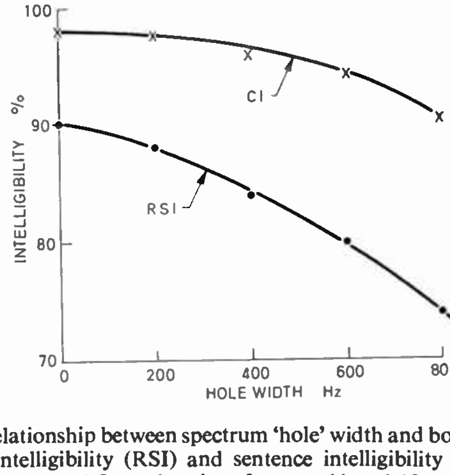


Fig. 8. Relationship between spectrum 'hole' width and both random syllable intelligibility (RSI) and sentence intelligibility (CI) computed from data in references 11 and 12.

greater reduction in intelligibility, but if it is made very narrow there are perceptible transient effects. Thus the choice of stop-band width depends on subjective factors, and values between 200 and 500 Hz appear acceptable to most hearers. A.c. coupling completely overcomes the problems caused by drift, which would otherwise set a lower limit to the input signal level which could be received. Instead, the ultimate sensitivity of the a.c. coupled Weaver receiver, like that of a more conventional counterpart, is set by the noise level of the early stages. In this connexion, the curtailment of the low frequency response of the i.f. amplifiers will reduce the effects of mixer $1/f$ noise.

The dynamic range of the receiver, that is to say the range of maximum to minimum signal level within which the receiver a.g.c. must keep the demodulated a.f. level, deserves some comment. The maximum audio output is determined by non-linearity effects, but whereas the minimum signal level would be set, in the case of a receiver using an envelope detector, by the relative inefficiency of a diode as a rectifier at low levels, in the case of the Weaver receiver it is determined by the level of carrier leakage in the second pair of demodulators. Although carrier leakage levels as low as −60 dB are claimed by the manufacturers for integrated semiconductor balanced mixers, such as the type 2401, this is considerably affected by circuit layout. Using printed circuit construction and normal care in layout, but no special precautions, the receiver described here achieved a ratio of 40 dB between carrier leak and maximum demodulated output. This dynamic range is more than adequate for speech reproduction, but could no doubt be further improved.

4 Discussion and Conclusions

The two types of receiver based on the direct conversion principle have been compared to the advantage of the a.c.-coupled Weaver form. However, both share a major advantage over the superheterodyne, in that they have many fewer spurious responses. Image or second-channel response (at a frequency removed from the signal frequency by twice the i.f.) vanishes, because in effect the i.f. is reduced to zero. Similarly the pairs of spurious responses observed in the superheterodyne, separated by the magnitude of the i.f. above and below each of the oscillator harmonics, are reduced to single responses at the harmonics. Higher-order spurious responses are reduced in number for similar reasons, and i.f. breakthrough does not occur. The interactions between first and subsequent converters which are often troublesome in multiple superheterodyne designs have not been observed in the direct conversion receivers described, or in any of the many others constructed in these laboratories for a variety of input frequency ranges. This is as would be expected, because the phasing receiver is of the single conversion type, and although two conversions occur in the Weaver configuration, the frequency of the second local oscillator is too low for it to interact with the first.

Direct conversion receivers should be cheaper to construct than superheterodynes of similar specification for a number of reasons. Because of the reduction in spurious responses, pre-mixer r.f. filter specifications can be less stringent. I.f. amplification uses low cost a.f. monolithic circuits and does not require tuneable components. Filtering is by fixed and relatively cheap, low-pass components. Thus, although the i.f. channels are duplicated, the overall cost is likely to be less than that of a single conventional i.f. strip. Multiple conversions are unnecessary. Circuits are non-critical if the d.c.-coupled form is avoided, and only a reasonable degree of gain equalization between i.f. channels is required, particularly in the Weaver circuit. For all these reasons, designers of receivers for s.s.b. applications ought to give serious consideration to the use of the direct conversion principle.

The problem of a.g.c. has not been considered here, but is dealt with conventionally. Application of a.g.c. to the mixers or i.f. amplifiers presents some difficulty if adequate i.f. gain balance between the channels is to be maintained. Otherwise, a.g.c. may be applied to the r.f. amplifier, or, as in the case of the v.h.f. receivers described here, to a p-i-n diode attenuator in the aerial circuit.[13]

To summarize, of the receiver configurations considered, the Weaver circuit is preferred to the phasing configuration because it can have better adjacent channel rejection, and its spurious response does not contribute interchannel cross-talk. In d.c.-coupled form, however, Weaver's receiver cannot be used for signals below, say 100 μV at the mixer input, because of drift, and even at this signal level will require all the resources of contemporary monolithic circuit technology to make drift performance acceptable. The problem is completely overcome in the a.c.-coupled receiver, which can have a sensitivity limited only by input circuit noise. The stop band introduced into the reproduced audio spectrum is narrow and need not significantly affect speech intelligibility.

5 References

1. Winn, R. F. E., 'The effect of receiver design in communication systems', Conference on Radio Receivers and Associated Systems, 1972. IERE Conference Proceedings No. 24.
2. Al-Araji, S. R. and Gosling, W., 'A direct conversion receiver for v.h.f.', *ibid.* Also *The Radio and Electronic Engineer*, **43**, 1973. To be published.
3. Pappenfus, E. W. *et al.*, 'Single Sideband Principles and Circuits, p. 305 (McGraw-Hill, New York, 1964).
4. Barber, N. F., 'A narrow band pass filter using modulation', *Wireless Engineer*, **24**, p. 132, 1947.
5. Hall, J. R., 'A phase rotation single sideband generating system', *RCA Review*, **16**, p. 43–51, March 1955.
6. Weaver, D. K., Jnr., 'A third method of generation and detection of single sideband signals' *Proc. Inst. Radio Engrs*, **44**, p. 1703–5, December 1956.
7. Al-Araji, S. R., 'Inductorless filters using integrated circuits', Ph.D. Thesis, University of Wales, 1971.
8. Chu, L. P., 'A phase lock a.m. radio receiver', *IEEE Trans on Broadcast and Television Receivers*, BTR-**15**, pp. 300–8, 1969.
9. Rustom, S., 'Realisation of i.f. filters using the Barber circuit', M.Sc. Thesis, University of Wales, 1971.
10. Pappenfus, E. W., *et al.*, *op. cit.*, pp. 39–41.

11. French, N. R. and Steinberg, J. C., 'Factors governing the intelligibility of speech sounds', *J. Acoust. Soc. Amer.*, **19**, p. 90, 1947.

12. Fletcher, H., 'Speech and Hearing in Communication", Chapter 17 (Van Nostrand, Princeton, 1953).

13. Walters, D.S. and Gosling, W., 'Helical filters with p-i-n diode Q control for pre-converter a.g.c.', Conference on Radio Receivers and Associated Systems, 1972, IERE Conference Proceedings No. 24.

6 Appendix: A brief analysis of the two main classes of direct conversion receiver

Let the incoming signal to the first mixers be $y(t)$, where

$$y(t) = \sum_r a_r.\cos(\omega_r t + \phi_r)$$

The quadrature first local oscillator outputs are x, x' where

$$x = 2\cos(\omega t), \qquad x' = 2\cos\left(\omega t + \frac{\pi}{2}\right)$$

After mixing (regarded as equivalent to multiplication) and low-pass filtering, two i.f. voltages result $y_1(t)$ and $y'_1(t)$, where

$$y_1 = \sum_r ka_r.\cos\left([\omega_r - \omega]t + \phi_r\right)$$

$$y'_1 = \sum_r k'a_r.\cos\left([\omega_r - \omega]t + \phi_r - \frac{\pi}{2}\right)$$

and k, k' are given factors.

The two receiver configurations will now be considered.

6.1 *The Phasing Receiver*

Assuming that the receiver is to be designed for upper sideband reception, the expression for y_1 may be re-written

$$y_1 = \sum_{\omega_r > \omega} ka_r.\cos\left([\omega_r - \omega]t + \phi_r\right) + \sum_{\omega > \omega_r} ka_r.\cos\left([\omega - \omega_r]t - \phi_r\right)$$

also

$$y'_1 = \sum_{\omega_r > \omega} k'a_r.\cos\left([\omega_r - \omega]t + \phi_r - \frac{\pi}{2}\right) + \sum_{\omega > \omega_r} k'a_r.\cos\left([\omega - \omega_r]t - \phi_r + \frac{\pi}{2}\right)$$

In each expression the first term on the right-hand side corresponds to the wanted and the second to the unwanted sideband. The voltage y'_1 then passes through the Hilbert transform network, which gives an output $H(y'_1)$ where

$$H(y'_1) = \sum_{\omega_r > \omega} k\,a_r.\cos([\omega_r - \omega]t + \phi_r) + \sum_{\omega > \omega_r} k'a_r.\cos([\omega - \omega_r]t - \phi_r + \pi)$$

Hence

$$y_1 + H(y'_1) = \sum_{\omega_r > \omega} (k + k')a_r.\cos([\omega_r - \omega]t + \phi_r) + \sum_{\omega > \omega_r} (k - k')a_r.\cos([\omega - \omega_r]t - \phi_r)$$

Evidently, if $k = k'$ the upper sideband only is reproduced, as required. If, alternatively, the receiver were to be sensitive to the lower sideband, the equivalent condition would be $k = -k'$.

6.2 *Weaver's 'Third Method'*

In this case the Hilbert transform network is not used, but y_1 and y'_1 are further mixed with a second pair of quadrature local oscillator outputs Z and Z', where

$$Z = 2\cos(\omega^* t), \qquad Z' = 2\cos\left(\omega^* t + \frac{\pi}{2}\right)$$

The result is two voltages in the a.f. range y_2 and y'_2 where

$$y_2 = \sum_r ka_r.\{\cos([\omega_r - \omega + \omega^*]t + \phi_r) + \cos([\omega_r - \omega - \omega^*]t + \phi_r)\}$$

$$y'_2 = \sum_r k'a_r.\{\cos([\omega_r - \omega + \omega^*]t + \phi_r) + \cos([\omega_r - \omega - \omega^*]t + \phi_r - \pi)\}$$

These expressions require some interpretation. The rth component of y is at a frequency ω_r, different by $(\omega_r - \omega)$ from ω. The rth components of y_2 are at two frequencies. One (at a frequency $(\omega_r - \omega + \omega^*)$) is different from ω^* by the same frequency difference as the original component of y from ω: it therefore corresponds to the original spectrum of y being translated along the frequency axis, so that if it was initially centred on ω it is now centred at ω^*. The other (at a frequency $(\omega_r - \omega - \omega^*)$) has a frequency difference from ω^* equal to the negative of the original $(\omega_r - \omega)$: it therefore corresponds to an inverted spectrum centred on ω^*, and constitutes the spurious response.

The receiver output is given by adding y_2 and y'_2, so that

$$y_2 + y'_2 = \sum_r (k + k')a_r.\cos([\omega_r - \omega + \omega^*]t + \phi_r) + \sum_r (k - k')a_r.\cos([\omega - \omega_r + \omega^*]t - \phi_r)$$

If $k = k'$ the spurious response is zero.

Manuscript received by the Institution on 25th July 1972. (Paper No. 1511/Comm. 64).

UDC 621.372.54: 621.374

Adjustment of digital filter characteristics after optimization

L. G. CUTHBERT, B.Sc.(Eng.)*

SUMMARY

One method of designing non-recursive digital filters to arbitrary frequency specifications is to use optimization. However, the program will only minimize one particular performance index which may not give the 'best' filter for the particular application. Furthermore, when performing the optimization the errors at specified points are usually weighted and the weights are generally chosen in a fairly arbitrary manner. This paper describes how two different optimization runs can be used to give initial conditions which enable the final filter characteristic to be simply adjusted, interactively with the computer, thus avoiding the problem of how best to choose these weights.

*Department of Electrical and Electronic Engineering, Queen Mary College, University of London, Mile End Road, London E1 4NS

1 Introduction

In a recent paper† it was shown that optimization techniques could be used to fit the characteristic of a non-recursive digital filter to an arbitrary frequency specification with good results. The procedure used was to let a computer program minimize an objective function (performance index) of the form

$$\sum_{i=1}^{N} W_i F(e_i, T_i) \tag{1}$$

where W_i is a weighting value, e_i the error at the ith specified point and T_i some form of tolerance. Various forms of the function F can be used but a useful one is a square term, making the overall function the sum of squares.

However, in many cases, specifications for filters are not particularly precise, only requiring that the response should be as near as possible to the specification. Thus, if the optimized solution is closer at some frequencies than at others, it would be convenient to have a technique to provide a 'fine adjustment' so that the designer can alter the characteristic a small amount, interactively with the computer, until the best compromise is obtained.

Optimization methods have the drawback that the computer is trying to minimize one particular performance index, and what is chosen for this will have a marked effect on the final characteristic. Also, the weight values are important because they distort the error surface and cause the heavily weighted parts of the specification to be met more closely at the expense of the other regions. One fundamental idea on this, discussed in the previous paper, was whether to optimize in the linear or logarithmic domains.

A possible interactive approach would be to do several optimization runs with different weights until the best characteristic is obtained. However, this would be very wasteful since optimization programs use a lot of computer time (typically 50–100 seconds of processor time on the Queen Mary College ICL 1904S). In this paper a very simple interpolation technique to adjust the characteristics is described which uses very little computer time (about 1 second).

2 Theory

Consider the M-tap coefficients to be the co-ordinates of a point in M-dimensional space. Thus the position vector of the point A is $(a_1, a_2, \ldots, a_M)$ and the frequency response is

$$G_A(\omega) = \sum_{k=1}^{M} a_k \exp(-jk\omega T) \tag{2}$$

where T is the sampling period. Similar expressions can be written for another vector **B**.

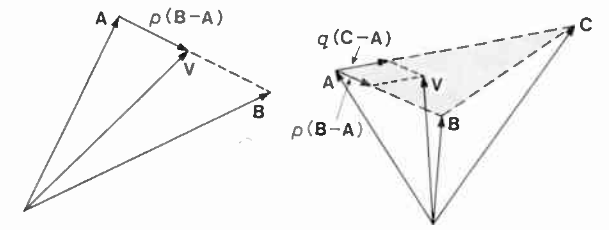
Now, in any vector space, a line can be drawn between two points and the position vector of a point along this

† Cuthbert, L. G. and Coward, P. R., 'The design of finite duration impulse response digital filters using optimization techniques', IERE Conference on 'Digital Processing of Signals in Communications', Loughborough, April 1972. IERE Conference Proceedings No. 23.

line is

$$\mathbf{V} = \mathbf{A}+p(\mathbf{B}-\mathbf{A}) \qquad (3)$$

p being a scalar parameter giving a measure of the distance along the line. This is illustrated in Fig. 1(a).



(a) Along a line. (b) In a plane.

Fig. 1. Interpolation between vectors along a line or in a plane.

Thus the frequency response of the coefficient set $\mathbf{V}$ is

$$G_{\mathbf{V}}(\omega) = \sum_{k=1}^{M} [a_k(1-p)+pb_k] \exp(-jk\omega T)$$

or

$$G_{\mathbf{V}} = (1-p)G_{\mathbf{A}}+pG_{\mathbf{B}} \qquad (4)$$

In general the gains, G, are complex so that equation (4) defines a complex interpolation. However, with linear phase filters the phase at any frequency is fixed, irrespective of the values of the tap coefficients, and, therefore, equation (4) can be re-written as

$$|G_{\mathbf{V}}| = (1-p)|G_{\mathbf{A}}|+p|G_{\mathbf{B}}| \qquad (5)$$

Thus having found the coefficient sets **A** and **B**, a third set can be found which has a response in between that of **A** and **B**. $p = 0$ corresponds to **A** and $p = 1$ to **B** and a value outside this range gives a vector to one side of the original vectors and use of such a value is discussed later.

This can be extended for three sets of parameters. The new point is chosen to be in the plane formed by the three position vectors **A**, **B** and **C** (Fig. 1(b)).

The position vector of a point in this plane is

$$\mathbf{V} = \mathbf{A}+p(\mathbf{B}-\mathbf{A})+q(\mathbf{C}-\mathbf{A}) \qquad (6)$$

or, for linear phase filters,

$$|G_{\mathbf{V}}| = (1-p-q)|G_{\mathbf{A}}|+p|G_{\mathbf{B}}|+q|G_{\mathbf{C}}| \qquad (7)$$

3 Practical Technique

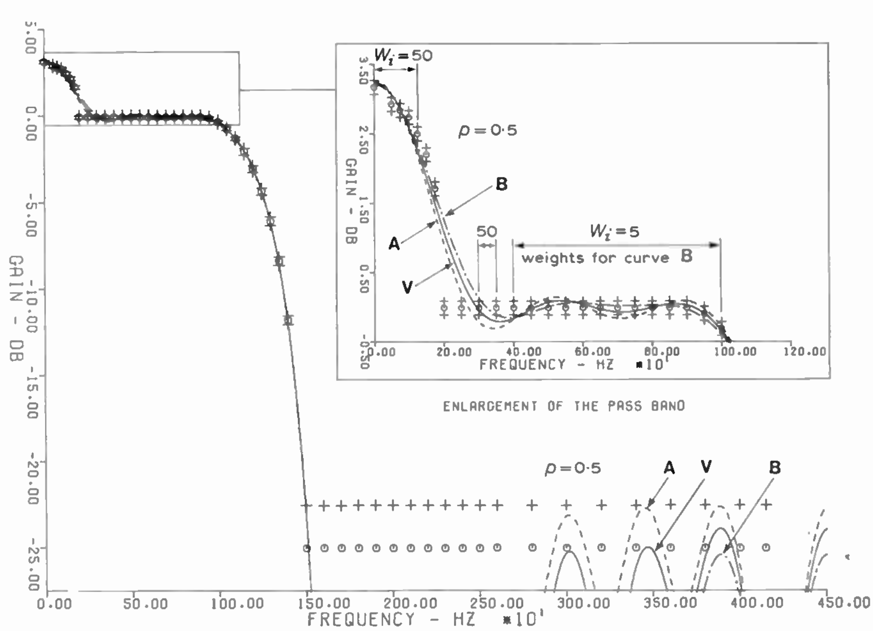
Firstly, a run of the optimization program with all weights unity is obtained and the resulting characteristic examined. The weights are then changed to force those parts of the characteristic most in error closer to the specification by heavily weighting the specified points in this area. The choice of which weights to alter and to what value is usually based on experience, but the advantage of the technique described here is that actual numerical values are not important provided that the required parts of the response are pushed nearer the specification. Suitable values used to date have been in the range 5 to 50, the parts most in error, or where it is most important that the specification is met, being given the heaviest weights.

After the optimization run it will usually be found that the response has deteriorated in the unity weighted regions.

A new characteristic can be found simply by interpolating between the two optimized responses by choosing a value of p so that the new characteristic will fit in between the other two and then using equation (3) to get the new characteristic. The value for p can either be chosen by selecting a particular point, perhaps where there is greatest difference between the two characteristics, and applying equation (4) formally to solve for p or by estimating a suitable value. Since the computer only uses a second to compute the new parameters and only a few more to plot the characteristic this procedure is particularly suited to interactive estimation.

It should be noted that this technique is most usefully applied to the pass-band and transition regions but it can, as Fig. 2 illustrates, provide a better stopband.

Fig. 2. Illustration of producing a response curve between two given frequency characteristics. The solid curve is derived from a point halfway between the other two.



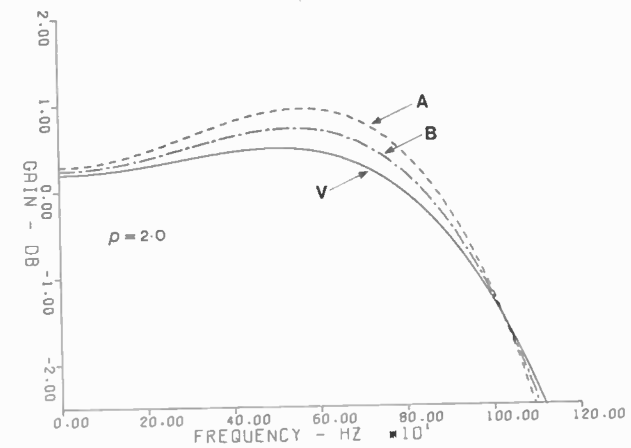


Fig. 3. Illustration of using p outside the range 0–1 in the pass-band.

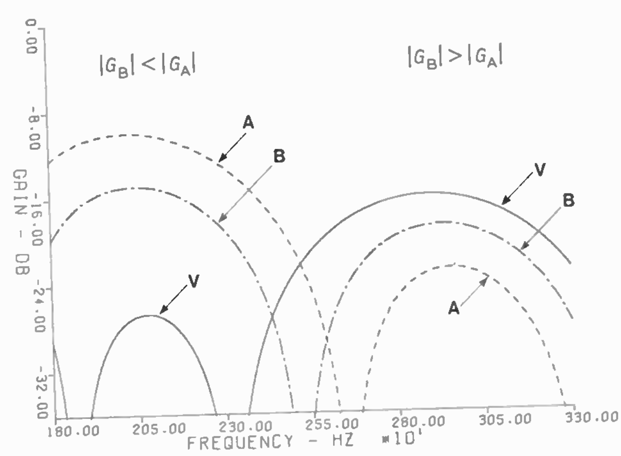


Fig. 4. Illustration of using p outside the range 0–1 in the stop-band. p is taken as 2·0 and the attenuation is reduced.

It is, of course, possible to use a value of p outside the range 0–1 and in this case a result is produced to one side of the two starting responses. This is illustrated in Fig. 3 where it is required to reduce the pass-band peak. However, when using such a value for p, it is important to take care in the stopband because the attenuation may well be less than that of *either* of the first two responses.

Consider $p = 2$. Then $|G_V| = 2|G_B| - |G_A|$ or $|G_V| = |G_B| + (|G_B| - |G_A|)$ and it can be seen that if $|G_B|$ is greater than $|G_A|$ the gain is increased, i.e. the attenuation reduced. This is shown in Fig. 4 and it is therefore recommended that a value for p should only be in the range 0 to 1 unless it is certain that this situation will not occur, or that where it will occur the result will be acceptable.

It is also possible to obtain 3 *independent* sets of parameters and interpolate in a plane to get an extra degree of freedom. Again p and q can be obtained by formally solving equation (7) at two frequencies or, perhaps better, by estimation with the computer. However, simply using two sets usually gives good enough results.

4 Design Example

The result of using this technique after two optimization runs for a linear phase filter of 40 taps is illustrated in Fig. 2. Circles on the plot represent the specification and crosses the limit of the acceptable response.

The coefficient set **A** (all weights 1) produced the dotted line, which fits the stop-band and transition region but has a higher ripple than desired in the pass band. The chain dotted line (**B**) was obtained using heavy weighting over the pass-band except around the low-frequency peak transition region. The weights used are marked on the illustration. Very heavy weighting (50) was used where it was important that the characteristic should meet the specification and a moderate value (5) used over the rest of the region. The ripple with the second curve was better but the slope from the peak was not steep enough. Therefore, the solid line, coefficient set **V**, was produced using the method described in this paper with an estimated value for p of 0·5 and it can be seen that this is a good compromise between ripple and steepness of slope.

5 Conclusion

A technique for adjusting the frequency characteristics of non-recursive digital filters after optimizing to arbitrary frequency specifications has been described and shown to give good results, even in its simple form. It has also been pointed out that this avoids the problem of how best to choose the weights for the optimization program.

It is not claimed that the final response is the best possible fit since optimization with an entirely different set of weights might produce a better curve. However, the method described is convenient since it only uses two optimization runs to set up initial conditions and the final characteristic can be determined interactively with the computer. It does give better results than just guessing at suitable weights.

The interpolation could also be usefully applied to coefficient sets produced by other means, for instance by optimization with two different methods of calculating the performance index.

6 Acknowledgments

The author would like to gratefully acknowledge the financial support of the Science Research Council and of Standard Telecommunication Laboratories.

Manuscript first received by the Institution on 15th August 1972 and in revised form on 9th November 1972. (Short Contribution No. 161/CC 162.)

UDC 621–521

An approach to the control of processes with pure time delay

P. ATKINSON,
B.Sc.(Eng.), A.C.G.I., C.Eng., M.I.E.E., M.I.E.R.E.*

Professor P. B. FELLGETT,
M.A., Ph.D., F.R.S.E., F.B.C.S., C.Eng., F.I.E.E., F.I.E.R.E.*

and

T. G. SWANN,
B.Sc., Ph.D.†

SUMMARY

The advent of on-line digital computers and relatively large-storage supervisory computers allows the possibility of improved control of processes containing pure time delay (distance/velocity lag). Such processes frequently occur in industrial plants and their control is also complicated by the fact that both the magnitude of pure time delay and the values of the parameters of the minimum-phase elements in the process may vary with time. Furthermore extraneous disturbances of time varying and unknown power spectra will cause further control difficulties. This paper contains an outline of a new general approach to this problem in which a digital computer is used as the main control element to implement a scheme designed to optimize the performance of such systems.

* *Department of Applied Physical Sciences, University of Reading, Whiteknights, Reading RG6 2AL*
† *Formerly at the University of Reading; now with Marconi-Elliot Computer Systems Limited, Borehamwood Hertfordshire.*

1 Introduction

Satisfactory control of a process requires that the output should both follow some predetermined input signal and remain unaffected by variations in the process parameters or by extraneous disturbances. Feedback techniques using high loop gain at frequencies of importance provide a perfectly satisfactory solution in the control of minimum phase processes. In practice many processes contain physically long transmission elements which transmit signals without distortion other than pure time delay T_d. For example in a steel rolling mill the thickness of steel will remain constant from the time it leaves a roller until it reaches the thickness gauge, which may be a metre or so away. The change of physical form of the variable from 'thickness' to 'measure of thickness' is of course irrelevant from the system engineer's point of view. Alternatively a pipeline, where no mixing occurs, will transmit the variable 'composition of flow' from one tank to another with no effect other than a delay in time. Since a disturbance at the input has no immediate effect on the output, the effect is also referred to as 'dead time'. The term 'distance/velocity lag' is also used to indicate the physical origin of the delay.

Where a process contains a large pure time delay the standard technique of control using high gain in a feedback loop is invariably impractical because of the enhanced possibility of loop instability. The transfer function of pure time delay is given by

$$D(s) = \exp(-sT_d)$$

where s is the Laplace operator.

There is no physically realizable inverse transfer function of $\exp(-sT_d)$. This is in fact self-evident since the inverse of time delay is pure prediction, the output appearing a time T_d before the input. Phase-lead terms of the form $(1+sT_c)$ which can normally be used successfully to counteract phase lags of the form $1/(1+sT_b)$, have virtually no effect when cascaded with a time delay. In fact they will more often lessen stability by increasing the gain at high frequencies while having a minimal effect on the phase. Where a process contains a time delay, T_d, comparable with the largest phase lag time-constant T_b, the time delay dominates the process response entirely.

Even in these circumstances it is still possible to maintain control by standard techniques, provided the performance requirements are not too stringent. This paper is concerned with the theory underlying a proposed new method of controlling systems with pure time delay in which performance requirements are stringent and in which the process itself is subject to slow random parametric changes together with extraneous disturbances. In particular the application of the digital computer to this problem has now made the implementation of the proposed scheme economically feasible.

In recent years a number of schemes have been proposed which enable the independent design of the response to commands and the regulation. Although details differ, the essence of all such schemes is the operation of a dynamic model of the process. This model is fed with the process input signal and therefore duplicates the process output except for the noise component, as shown in Fig. 1. The process transfer function here is $G(s)$ followed by pure time delay T_d, the model represents $G(s)$ by $G_1(s)$. The output components due to commands and noise can then be separated and controlled independently. This idea was first proposed by Lang and Ham,[1] and later developed by Smith,[2, 3, 4,] Lipfer and Oglesby,[5] and Petrovic.[6]

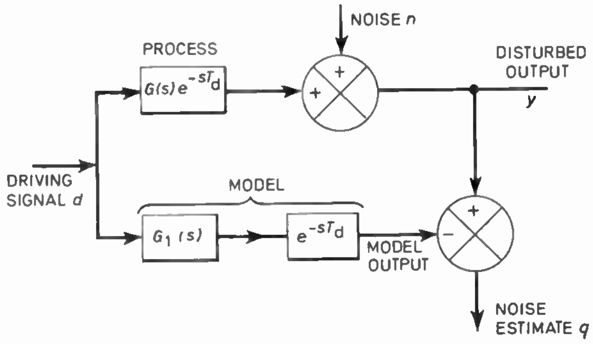


Fig. 1. Dynamic model of process.

The overall system block diagram can be built up from Fig. 1 in a number of ways, but the most straightforward is as follows. Since the response to commands is to be designed independently, there is no need to use a feedback loop. The best performance can be achieved by means of an open-loop compensator $C(s)$ placed in series with the process. The noise, derived in the manner of Fig. 1, can then be minimized by feeding back negatively through a suitable filter $P(s)$, (Fig. 2). It is important to note that if the model is accurate, any signal d at the process input has no effect on the estimated noise q. Thus although the noise is fed back to the system input, the element $P(s)$ is effectively not in a feedback loop. The output of $P(s)$ affects the process in the same way as open-loop commands. If the compensation provided by $C(s)$ is good, the overall response will approximate to a pure delay of T_d and the optimum form for $P(s)$ is a linear predictor. If the compensation is poor this approximation will be less accurate, but $P(s)$ will still be essentially a predictor.

2 Theory for the Implementation of the New System

Although the ideas behind Fig. 2 have been known for some years, their use has been offset by practical difficulties.

It is clear from Fig. 2 that any mismatch between process and model will not only affect the response to commands, but will effectively close a loop around the predictor. This will not only further impair the system performance but could lead to instability. Since any real process is subject to possible parameter change, it is essential that the model be made actively adaptive. It follows that the open-loop compensator must also be adaptive. Since the predictor can only be designed on the statistics of the noise, this too should vary as the process parameters alter. The system thus has the following functions to perform:

(i) Adaptable simulation of the process.

(ii) Adaptable representation of (a) compensator
(b) predictor.

(iii) Identification of the process characteristics.

(iv) Measurement of the noise characteristics.

(v) Adaptation of (a) model
(b) compensator
(c) predictor.

It is clear the the scheme is scarcely feasible without an on-line digital computer. However since these are now common, it is worthwhile re-examining the scheme on the assumption that a small computer (or equivalent power time-shared from a large computer) is available for the control of the single process. The additional problems due to sampling can be overcome quite easily. However there are still several difficulties, in particular (iii) above.

2.1 *Process Identification*

The process must be identified frequently to ensure that the process and model never differ significantly. This means that the identification must be carried out on-line without disturbing the normal process operation. Although there are nominally several means for doing this, most of them are in fact unusable as they rely on correlating the process output with some low-level high-bandwidth input such as pseudo-random binary sequence. This type of method yields the total transfer function between the signal injection channel and the pick-off point. It may be seen from Fig. 2, that as the process parameters wander, a loop is closed round the process incorporating the model and predictor.

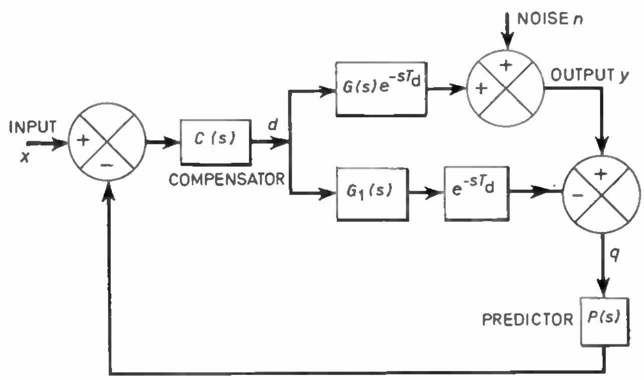


Fig. 2. Development of Fig. 1 to minimize noise.

As a result, a direct cross-correlation method would give some closed-loop transfer function from which it would be nearly impossible to calculate the wanted open-loop function. Initially, of course the result would be quite close to the open-loop transfer function, but there is no guarantee that the repeated process of identification and model adjustment would remain stable at the correct parameter values. It follows that identification is only possible by the comparison of the total input/output signals. This may be done by the generalized least-squares technique, developed by Clarke.[7] The process is represented by the z-transform model of eqn. (1):

$$y_t = \frac{z^{-k}P(z^{-1})}{1+Q(z^{-1})}d_t + n_t \tag{1}$$

in which

$$P(z^{-1}) = p_1 z^{-1} + p_2 z^{-2} \dots p_m z^{-m}$$

and

$$Q(z^{-1}) = q_1 z^{-1} + q_2 z^{-2} \dots q_m z^{-m}$$

and n_t is the noise, not necessarily white. The term z^{-k} incorporates the time delay, making it unnecessary to have many zero terms in P. However to monitor changes in the delay it is wise to continually adjust k so that the first term of P is always zero.

Rearranging:

$$y_t = -Q(z^{-1})y_t + z^{-k}P(z^{-1})d_t + [1+Q(z^{-1})]n_t \tag{2}$$

i.e. using matrix notation:

$$y_t = (z^{-1}y_t\ z^{-2}y_t \dots$$

$$\dots z^{-m}y_t \,\vdots\, z^{-k-1}d_t \dots z^{-k-m}d_t) \begin{bmatrix} -q_1 & p_1 \\ \cdot & \cdot \\ \cdot & \cdot \\ \cdot & \cdot \\ -q_m & p_m \end{bmatrix} + e_t \tag{3}$$

If long sequences of input/output samples are taken, eqn. (3) can be written in the vector form:

$$\mathbf{y} = \mathbf{X}\mathbf{u} + \mathbf{e}$$

where $\mathbf{X}$ is a matrix of y, d samples, $\mathbf{u}$ is the vector $(-Q, P)^T$, and $\mathbf{e}$ is the generalized noise vector.

Solving by the least-squares method:

$$\mathbf{u} = (\mathbf{X}^T\mathbf{X})^{-1}\mathbf{X}^T\mathbf{y} - (\mathbf{X}^T\mathbf{X})^{-1}\mathbf{X}^T\mathbf{e} \tag{5}$$

where it may be noted that $\mathbf{X}^T\mathbf{X}$, $\mathbf{X}^T\mathbf{y}$ and $\mathbf{X}^T\mathbf{e}$ represent correlation of the y, d samples with themselves, $\mathbf{y}$ and $\mathbf{e}$. If $\mathbf{e}$ is uncorrelated with $\mathbf{d}$ or $\mathbf{y}$, then $\mathbf{X}^T\mathbf{e}$ is zero and the identification is complete. The best estimate of $\mathbf{u}$, $\hat{\mathbf{u}}$ is given by:

$$\hat{\mathbf{u}} = (\mathbf{X}^T\mathbf{X})^{-1}\mathbf{X}^T\mathbf{y}. \tag{6}$$

Normally though $\mathbf{e}$ will be correlated, we can however estimate $\mathbf{e}$ by first forming an initial estimate of $\mathbf{u}$ from eqn. (6); then:

$$\hat{\mathbf{u}} = \mathbf{y} - \mathbf{X}\hat{\mathbf{e}}. \tag{7}$$

Suppose now that $\mathbf{e}$ is equivalent to white noise passed through an autoregressive process:

$$[1+A(z^{-1})]e_t = \xi_t \tag{8}$$

where

$$A(z^{-1}) = a_1 z^{-1} + a_2 z^{-2} + \dots a_n z^{-n}$$

where ξ_t is an uncorrelated sequence.

We then have:

$$e_t = -(z^{-1}e_t\ z^{-2}e_t \dots z^{-n}e_t) \begin{pmatrix} a_1 \\ \cdot \\ \cdot \\ \cdot \\ a_n \end{pmatrix} + \xi_t \tag{9}$$

or:

$$\mathbf{e} = \mathbf{E}\mathbf{a} + \xi \tag{10}$$

which may be solved by the method of least-squares to give the polynomial A. We can now filter the original records of $\mathbf{y}$ and $\mathbf{d}$ in such a way as to make the noise white.

Thus from eqn. (2):

$$[1+A(z^{-1})]y_t = -Q(z^{-1})[1+A(z^{-1})]y_t + z^{-k}P(z^{-1})[1+A(z^{-1})]d_t + \xi_t \tag{11}$$

If we write $\mathbf{y}'$, $\mathbf{d}'$ for the filtered records we have:

$$y_t' = -Q(z^{-1})y_t' + z^{-k}P(z^{-1})d_t' + \xi_t \tag{12}$$

Since the initial estimate of $\hat{\mathbf{u}}$ was only approximate, the values of ξ_t deduced from eqn. (12) will still be slightly correlated. However the same technique can be applied iteratively until the correlation is removed. This will be shown by the correlation matrix $\mathbf{E}^T\mathbf{E}$ from eqn. (10) becoming vanishingly small.

It may be noted here that the success of the method is independent of any feedback round the process provided only that a stochastic signal enters the loop at some point. From eqns. (4) and (5) it is seen that the crucial feature is the inversion of the matrix of correlation coefficients $\mathbf{X}^T\mathbf{X}$ to give a unique solution for $\mathbf{u}$. It is known that the inverse to such a correlation matrix exists, provided $\mathbf{d}$ and $\mathbf{y}$ are distinct stochastic sequences, the computation becoming more accurate the closer they approach white noise. For example, suppose that a loop exists connecting $\mathbf{y}$ to $\mathbf{d}$, this may be drawn in a general fashion as in Fig. 3(a) where $\mathbf{y}$ is an input stochastic sequence. We then have:

$$\frac{d_t}{v_t} = \frac{B(z)C(z)}{1+G(z)C(z)} \tag{13}$$

or

$$d_t = H(z)v_t. \tag{14}$$

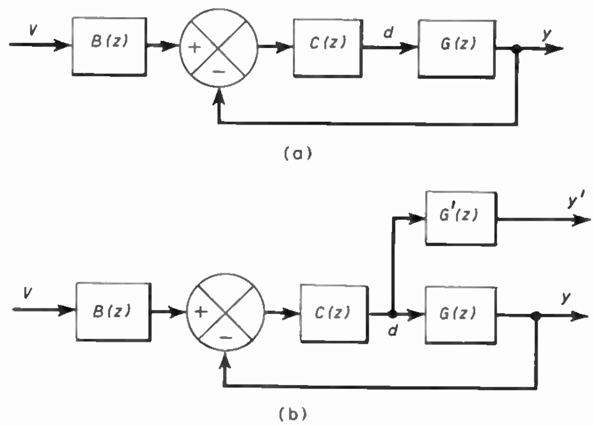


Fig. 3. Generalized loops.

It is apparent that the process G' in Fig. 3(b) can regard $\mathbf{d}$ as a stochastic input, and can be uniquely identified by comparison of $\mathbf{d}$ and $\mathbf{y}'$. The relation, $y_t' = G'(z)d_t$, holds regardless of the filter used to derive $\mathbf{d}$ from $\mathbf{v}$. Hence, of course, G can be similarly identified by comparison of $\mathbf{d}$ and $\mathbf{y}$ as proposed above. However the identification will be inaccurate if the loop or input variables are such that **d** has a low bandwidth. This error can be reduced by adding to **d** a random or pseudo-random sequence as in the standard cross-correlation identification method.

Unfortunately, in this form, the method requires excessive storage for the input/output records and various matrix operations. It may however be adapted into a recursive method as described by Hastings-James and Sage.[8] In this form it is quite feasible to implement the technique using a typical 4k store process control computer, although it would still occupy most of the storage space. Unfortunately though, using the recursive method, it is not possible to adjust the value of k in the model. Thus to accommodate possible decrease in the time delay, an initial value of k must be chosen such that $P(z^{-1})$ has several leading zeros, r say. This is wasteful of computer space and time as it enlarges the matrices as the square of r and may well make the procedure impractical. The only feasible approach then is to implement one of these techniques using time and space borrowed from a large computer. Such computers are becoming increasingly available, used for a variety of control and supervisory functions. There is then little to choose between the two methods, as they would both involve regular data-gathering and periodic transfers to and from the large machine. The method of Clarke is perhaps more attractive as it does not involve a complicated restart in the event of a divergent calculation or other disaster.

2.2 *Adaptation and Simulation*

Assuming that identification, (iii), is satisfactory, adaptation and simulation, (i) and (v) (a), present no problem. The z-transform model is already the ideal form for simulation. It is the z-transform of the pulse response in the form of the ratio of two short polynomials in z^{-1}. The polynomials are applied as weighting sequences to the simulation input and output. This is similar to, but more economical than the use of the pulse response itself, which is applied as a long weighting sequence to the input alone. Compensation, (ii)(a) and (v)(b), is also no problem as there are a number of techniques available which operate purely in the time-domain using a z-transform model.[9, 10, 11] On examination it is found that these techniques are very suited to the automatic computation of an open-loop control algorithm.[12] An algorithm can be found that provides optimal control (in some sense), for very little computing effort. The basic approach is to define a desired optimal response modified only to the extent that it is physically realizable. This criterion is readily met by inspection of the z-transfer function of the process.[10, 11, 12] Since the definition of optimal is to some extent arbitrary there will be a wide range of choice of possible responses. In general though the resulting overall transfer function will approximate to the time delay T_d, followed by a further delay of one or two sample intervals.

2.3 *The Predictor*

The only problem remaining is that of constructing a suitable predictor. The original approach by Wiener[13, 14] is not at all suited to the on-line calculation of a digital predictor. However a simplified method may in practice be used in which an mth order, moving-average model forms the predictor.[15] If the noise is n_t, and the predictor is represented by the mth-order polynomial $P(z^{-1})$, prediction k samples ahead may be written:

$$n_{t+k} = P(z^{-1})n_t + e_{t+k}. \tag{15}$$

On comparison it may be seen that this corresponds exactly to eqn. (2), and for a large number of samples we can write:

$$\mathbf{n} = \mathbf{X}\mathbf{u} + \mathbf{e} \tag{16}$$

where $\mathbf{X}$ is a matrix of values of $\mathbf{n}$, and $\mathbf{u}$ is $(P)^T$. Since it is used to predict the process output noise, the predictor is designed for a 'noiseless' situation, so the only error is due to the unknown future values of the postulated white noise sequence from which $\mathbf{n}$ is derived.[14] Thus $\mathbf{e}$ is totally uncorrelated with the values of $\mathbf{n}$ in $\mathbf{X}$ and there is no need for interaction. The solution may therefore be written down directly, to correspond to eqn. (6):

$$\mathbf{u} = (\mathbf{X}^T\mathbf{X})^{-1}\mathbf{X}^T\mathbf{n} \tag{17}$$

$\mathbf{X}^T\mathbf{X}$ is then the autocorrelation matrix of the noise for delays 0 to m, and $\mathbf{X}^T\mathbf{n}$ is the vector of autocorrelation values for delays k to $m+k$. Thus, as expected from Wiener theory, the predictor is completely defined by the autocorrelation function of the signal to be predicted.

Experiments have shown that using only a 2nd-order model near optimum prediction can be achieved for white noise filtered by two exponential lags, and the performance is still good for a three-lag noise model. The calculation and use of the predictor is extremely simple, the only difficulties being those of estimating the autocorrelation function of the noise.

3 Conclusion

This paper has summarized the theory underlying a proposal for the improved control of processes containing comparatively long pure time delay and subject to disturbances of time-varying power spectrum. It would appear that the proposed control scheme could be successfully implemented using a small digital computer with shared access to a machine of larger storage capacity.

4 References

1. Lang, G. and Ham, J. M., 'Conditional feedback systems—a new approach to feedback control', *Trans. Amer. Inst. Elect. Engrs.* 74, II, pp. 152–61, 1955, *Applications and Industry*, No. 19, July 1955.
2. Smith, O. J. M., 'Feedback Control Systems' (McGraw-Hill, New York, 1958).
3. Smith, O. J. M., 'Closer control of loops with dead time'. *Chem. Eng. Prog.*, 53, No. 5, pp. 217–19. May 1957.
4. Smith, O. J. M., 'A controller to overcome dead time', *Instrum. Soc. Amer. J.*, 6, No. 2, pp. 28–33 February 1959.
5. Lupfer, D. E. and Oglesby, M. W., 'Applying dead-time compensation for linear predictor process control', *Instrum. Soc. Soc. Amer. J.*, 8, No. 11, pp. 53–57 November 1961.
6. Petrovic, R., 'Prediction in control systems containing a pure time delay', *Control*, 13, No. 127, pp. 61–62, January 1969.
7. Clarke, D. W., 'Generalized least squares estimation of the parameters of a dynamic model', IFAC Symposium on Identification in Automatic Control Systems, Prague, 1967.

8. Hastings-James, R. and Sage, M. W., 'Recursive generalized least squares procedure for on-line identification of process parameters', *Proc. Instn. Elect. Engrs*, **116**, No. 12, pp. 2057–62, December 1969.
9. Williams, B. J., 'The Computer Control of a Distillation Column', N.P.L. Autonomics Report, No. NPL-COM-SCI-37, July 1968.
10. Tou, J. T., 'Digital and Sampled-data Control Systems', (McGraw-Hill, New York, 1959).
11. Box, G. E. P. and Jenkins, G. M., 'Some statistical aspects of adaptive optimisation and control', *J. R. Stat. Soc.*, B, **24**, No. 2, pp. 297–343 1962.
12. Swann, T. G., 'An Approach to the Control of Processes with Pure Time Delay,' Ph.D. Thesis, Reading University, 1970.
13. Wiener, N., 'The Interpolation, Extrapolation and Smoothing of Stationary Time Series', (Wiley, New York, 1949).
14. Bode, H. W. and Shannon, C. E., 'A simplified derivation of linear least square smoothing and prediction theory', *Proc. Inst. Radio Engrs*, **38**, No. 4, pp. 417–25, 1950.
15. Leviason, N. 'The Wiener r.m.s. error criterion in filter design and prediction', *J. Maths and Phys*, **26**, p. 110, July 1947.

Manuscript first received by the Institution on 15th May 1972 and in final form on 18th October 1972 (Paper No. 1512/ACS1).

UDC 621.372.54: 621.374

Binary transversal filters using recirculating shift registers

G. B. LOCKHART, M.Sc., Ph.D.*

and

S. P. BABARY, M.Sc.‡

SUMMARY

The use of recirculating shift registers in the implementation of binary transversal filters with quantized coefficients is discussed. Shift registers are employed in a serial processing mode to reduce circuit complexity at the expense of a lower maximum processing rate. An implementation of a 34-stage filter using standard m.o.s. shift registers is discussed in detail. Some alternative arrangements are considered.

* Formerly at the Department of Electronic and Electrical Engineering, University of Technology, Loughborough; now with the Department of Electrical and Electronic Engineering, The University of Leeds, Leeds, LS2 9JT.

† Formerly at the Department of Electronic and Electrical Engineering, University of Technology, Loughborough; now with Pye Business Communications, Wembley, Middlesex.

1 Introduction

A binary transversal filter[1, 2] (b.t.f.) consists of a binary shift register and a set of weighting resistors which determine its impulse response. The shift register does not suffer from the usual imperfections of the analogue delay line; delays between sections are controlled by clock-pulse timing and tap outputs are constrained to standard binary values. The b.t.f. input may be derived from any device or system which generates a binary-valued clocked output on which a linear filtering operation is required. Thus binary data streams or the digital output of delta or delta-sigma modulators or binary pseudo-random number generators may be usefully processed by the b.t.f.

Quantization of the resistor weighting coefficients[3] has certain advantages in the microcircuit implementation of the b.t.f. particularly in regard to l.s.i. techniques. If each weight is limited to a finite number of alternatives coefficients may be set digitally thereby reducing the number of different resistor values. The resulting restriction on the variety of impulse responses realizable in this way can be eased by feeding the sum of the weighted outputs to a single linear network such as an integrator. For example, binary coefficients are stored in shift register SR2 of the programmable b.t.f. in Fig. 1. If the logic levels associated with shift register SR1 are $\pm V$ volts, then multiplication by ± 1 is effected by the EXCLUSIVE-OR gates which present $-V$ volts to the integrator only when their input levels coincide and otherwise $+V$ volts. The impulse response of the filter is the response to the input sequence $(+V, 0, 0, \ldots)$ which will cause the integrator output to rise or fall by a fixed step according to the static bit pattern stored in SR2. (Although this input sequence cannot be applied in practice because input levels are constrained to $\pm V$ volts, the impulse response can be conceived as half the superposition of filter responses to the sequences $(+V, -V, -V, \ldots)$ and $(+V, +V, +V, \ldots)$ respectively.)

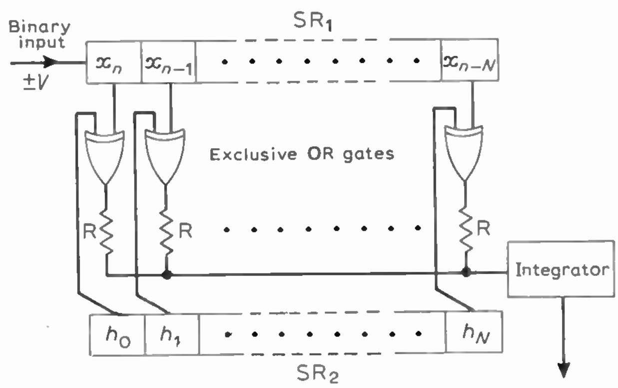


Fig. 1. Binary transversal filter with programmable binary coefficients.

The b.t.f.s to be discussed utilize recirculating shift registers in a serial mode so that only one coefficient multiplier per shift register pair is required, rather than $(N+1)$ as in Fig. 1. This advantage is offset by a lower maximum processing rate in comparison with parallel arrangements, but the availability of standard m.o.s. shift registers with maximum shifting rates in the megahertz range makes serial processing realistic for at least audio frequency filtering.

2 Use of Recirculating Shift Registers

It will be shown that the basic arrangement of Fig. 2 which utilizes two recirculating shift registers and a single EXCLUSIVE-OR/resistor combination performs essentially the same filtering operation as Fig. 1. If the bit patterns in SR1 and SR2 are as illustrated in Fig. 1 the nth output (before integration) is given by

$$y_n = \sum_{r=0}^{N} h_r x_{n-r} \tag{1}$$

where $h_r = \pm 1$ and $x_{n-r} = \pm V$.

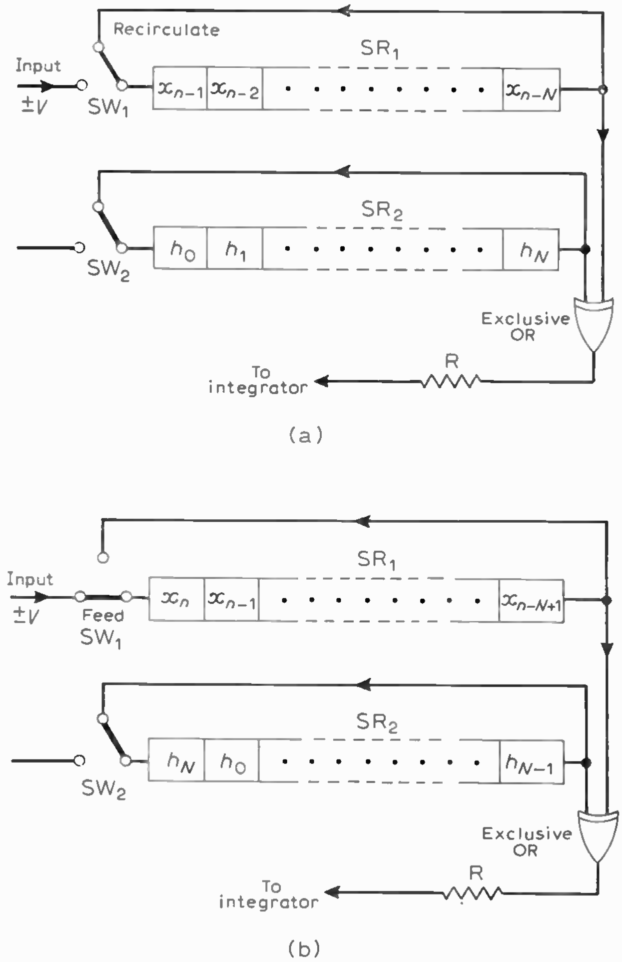


Fig. 2. Use of recirculating shift registers.

In Fig. 2, SR1 and SR2 again contain input and impulse response bits respectively but the contents of each are recirculated by N shifts after the occurrence of every new input digit so that the products of eqn. (1) are available serially from the last stage of each shift register. SR1 is one bit shorter than SR2 and both are clocked at a rate $(N+1)$ times the input bit rate.

Consider SR1 and SR2 bit patterns as illustrated in Fig. 2(a) for which the output product is $x_{n-N}\,h_N$ (i.e. $r = N$ in eqn. (1)). Let the next shift register clock pulse coincide with an input clock pulse in which case SW1 is switched (Fig. 2(b)) so that x_{n-N} is discarded, SR1 accepts a new digit, x_n, and the output product becomes $x_{n-N+1}h_{N-1}$ (i.e. $r = N-1$). At the next shift SW1 returns to the recirculate position and remains there for successive shifts, until the occurrence of the next input clock pulse.

The products of eqn. (1), $x_{n-r}h_r$, $r = N, N-1, \ldots, 0$, are therefore serially presented to the integrator in the order indicated above until $r = 0$ when a further shift causes the SR2 bit pattern to return to the $r = N$ position (Fig. 2(a)). Because SR1 is one bit shorter than SR2 the SR1 pattern is correctly advanced by one bit and the process can continue as before with x_{n-r+1} replacing x_{n-r}. The integrator forms the running sum of the products so that its output changes by

$$\left(\sum_{r=0}^{N} h_r x_{n-r}\right)$$

between input clock pulse at time positions corresponding to $r = 0$. The set of output samples taken at these times is therefore identical to the output samples from the parallel arrangement of Fig. 1.

3 Implementation

Two dual 16-bit m.o.s. shift registers were employed to implement a 34-stage b.t.f. using the technique described above. The operation of the circuit is illustrated in Fig. 3.

An effective b.t.f. length of 34 stages is achieved by cascading two 17-bit b.t.f.s, SR1(a)/SR2(a) and SR1(b)/SR2(b). The system is clocked by the generator G which is applied directly to all shift registers. Since a new digit must enter SR1(a) and SR1(b) every 17 shifts the input clock is derived by dividing the generator output by 17 before application to the feed/recirculate switches SW1 and SW3. (If necessary, the input clock may be used to synchronize analogue to binary conversion or other processes prior to filtering.) The four NAND gates in SW1 either feedback the output of SR1(a) during recirculation or enter a new input digit when an input clock pulse occurs. SW3 performs the same function with respect to SR1(b) but inputs digits which have been discarded by SR1(a). The first 17 bits of the impulse response bit pattern circulate through SR2(a), the extra stage D(a) and the switch SW2. The last 17 bits circulate through SR2(b) and SW4 in the second section of the filter. During operation SW2 and SW4 are set to the recirculate position but initially can be switched to feed in a required impulse response bit pattern via a manually-operated paper tape reader designed for this purpose.[4] Binary digits from SR1(a)/SR2(a) and SR1(b)/SR2(b) are multiplied by the two EXCLUSIVE-OR gates and the outputs added and integrated by the RC-integrator. The potentiometer P is adjusted so that the levels presented to the integrator are $\pm V$ volts. The integrator itself is a simple RC-type and therefore insensitive to small residual 'd.c.' voltages caused by fluctuations in the logic levels or slight unbalance of P.

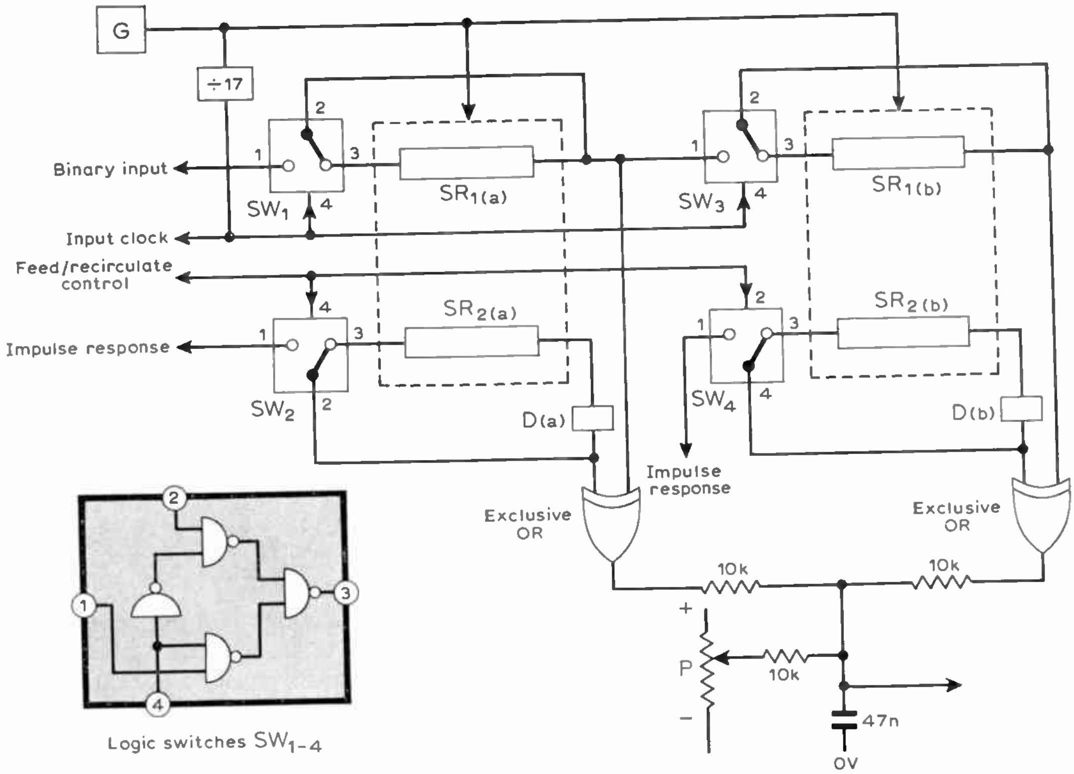


Fig. 3. Implementation of a 34-stage filter.

4 Alternative Arrangements

Configurations other than Fig. 3 are feasible and vary according to the required 'trade-off' between circuit complexity and maximum processing rate. For example, the operation of SR1(a) and SR1(b) in series leads to a single 33-stage b.t.f. This is the ultimate serial form. Circuit complexity is reduced at the expense of halving the maximum processing rate.

Binary coefficients are employed in the b.t.f.s described above. For improved coefficient characterization two or more impulse response registers can be placed in parallel to increase the number of available weights. Alternatively the shift register lengths can be increased at the expense of higher shifting rates.

5 Applications

The technique described can be applied in the realization of any b.t.f. provided processing rates are significantly less than the maximum shifting rate of the m.o.s. registers. This applies to the prototype[4] which was designed for use in a digital filter based on delta modulation[5] operating at input frequencies of less than 3 kHz.

6 Acknowledgment

The writers are indebted to Professor J. W. R. Griffiths for the use of facilities in the Department of Electronic and Electrical Engineering, Loughborough University of Technology.

7 References

1. Leuthold, P., 'Filternetz mit digitalen Schielberegistern', *Philips Res. Repts*, Suppl. 5, 1967.
2. Voelcker, H. B., 'Generation of digital signalling waveforms', *IEEE Trans. on Communications*, **COM-16**, pp. 81–93, 1968.
3. Lockhart, G. B., 'Binary transversal filters with quantised coefficients', *Electronics Letters*, **7**, pp. 305–7, 1971.
4. Babary, S. P., 'Digital Filters Based on Delta Modulation', M.Sc. Dissertation, Loughborough University of Technology, September 1971.
5. Lockhart, G. B., 'Digital encoding and filtering using delta modulation', *The Radio and Electronic Engineer*, **42**, pp. 547–51, December 1972.

Manuscript first received by the Institution on 12th June 1972 and in final form on 8th November 1972. (Short Contribution No. 162/CC163.)

LETTERS

From: Dr. J. C. Majithia,
Dr. A. Pugh, C.Eng., F.I.E.R.E.
and Mr. H. Dunderdale, B.Sc.Tech.

Designing Asynchronous Counters

In his recent paper,[1] Mr. H. Dunderdale presents a technique for designing asynchronous counters. This technique of using transition maps together with generalized transitions to design clocked sequential circuits is well known and has been in use for several years. The design of synchronous and asynchronous counters is a special case of this and is in fact a variation of the procedure described by Marcus.[2,3] The design approach outlined by Dunderdale is a standard one and has been taught at undergraduate levels. (Both the authors have taught this material in the undergraduate courses in Digital Systems at the University of Waterloo.) Several aspects of this technique which are not obvious in Ref. 1 are now outlined.

In general the transitions are defined as follows:

$0 \rightarrow 0$ φ transition
$1 \rightarrow 1$ I ,,
$0 \rightarrow 1$ α ,,
$1 \rightarrow 0$ β ,,

Counters may be regarded as a special case of sequential circuit synthesis. Since the counting sequence is given, it becomes a relatively simple procedure to generate the transition table and the excitation table and complete the synthesis. The use of generalized transitions in this procedure allows one to proceed with the synthesis without initially selecting the type of flip-flop. Hence the selection of flip-flop can be left as the final step in the design. In counting sequences some conditions or count sequences never occur and these are considered as 'don't care' conditions. The conditions required at the inputs of a flip-flop can be easily derived in terms of the generalized transitions. For example, in the case of the D-type flip-flop D must be '1' for output to obey either α or I transition. These conditions are listed in Table 1.

Table 1
Input terminal conditions in terms of transitions

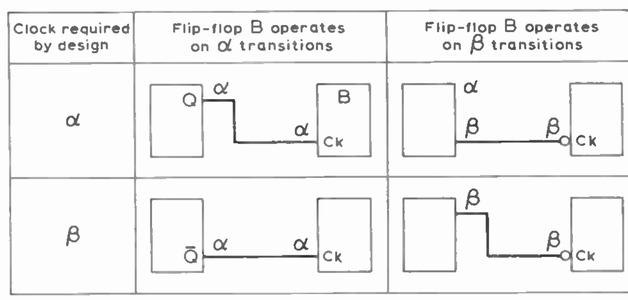
Flip-flop input terminal	Essential transitions	Optional or 'don't care' transitions
R	α	I, X
S	β	φ, X
J	α	β, I, X
K	β	α, φ, X
T	α, β	X
D	α, I	X

'X' refers to 'don't care' conditions, i.e. sequences which do not occur.

In synchronous counter design, only one aspect of design need be considered, i.e. determine the inputs logic for the various input terminals of the flip-flops. The transition maps of various stages have no interaction and therefore the design procedure is straightforward. For asynchronous design, this is not the case. The first step is to find transitions from other stages, of a single type (i.e. all α or all β), which provide the clock input for a particular stage. The clock input must be provided for the essential transitions. This leads to a simplification that all other entries on the transition map become 'don't care' conditions. This aspect has been used by Dunderdale in his design examples although it is not explicitly stated.

The operating characteristics of a particular flip-flop must must also be taken into account in the final design. In general, a flip-flop may change state on either a α transition or a β transition. Thus, the commercial J-K flip-flop (e.g. SN7473) triggers on a β transition. If the design procedure requires an α clock source from another stage then these must be derived from the $\bar{Q}$ output of that stage. This leads to four possibilities shown in Table 2.

Table 2
Derivation of clock sources for the two possible types of flip-flops.



Several disadvantages are inherent in the design of asynchronous counters using transition maps. Except for small systems and well-known counting sequences, it is not always possible to derive the required clock transitions. This makes a computer-aided design approach for asynchronous counters very impractical. In contrast, the transition map method for synchronous counters has been found easy to program. The method and the associated minimization procedure has been programmed using APL[4].

J. C. MAJITHIA
ALAN PUGH*

Department of Electrical Engineering,
University of Waterloo,
Waterloo, Ontario, Canada.
23rd November 1972

With reference to comments made by Majithia and Pugh, I do not claim the 'technique of using transition maps together with generalized transitions to design clocked sequential circuits' to be other than well-known and indeed, in the opening sentence of the introduction I refer the reader to published work on this subject[5] as the basis for the design method. The material referred to by Majithia and Pugh in references 2 and 3 is only relevant in that it deals with the derivation of the sufficient conditions to achieve specified transitions and this point is more relevantly dealt with in reference 1.

The approach to asynchronous counter design I described is one I developed about two years ago and I have taught this material in undergraduate and other courses at the University

* Visiting Professor at University of Waterloo during 1972; now returned to Department of Electrical and Electronic Engineering, University of Nottingham.

of Salford.[6] The novelty of the method lies in the selection of internal clock sources to enable the asynchronous counter to be realized from synchronous flip-flops. If step 3 of the design method is followed, the availability of a suitable clock source for a stage can rapidly be ascertained by scanning the columns defining the sequence at the points where change occurs in the column relevant to the stage under design. If at these points of a particular column a unipolarity change occurs, i.e. either $1 \rightarrow 0$ or $0 \rightarrow 1$ then this column presents a suitable clock source. Of course the overall system should not loop round such that a stage used as a clock source is itself clocked by stages driven directly or indirectly from this stage.

The point made by Majithia and Pugh concerning the additional 'don't care' entries is specifically dealt with in the second paragraph on page 418 of my paper. J and K entries are only made for values of the relevant state variables immediately before an active clock edge transition, other values of these variables giving 'don't care' entries. The elimination of the clock pulse source variable simplifies the Karnaugh maps even further. Thus these maps are normally less complex than for a synchronous design, requiring fewer variables to define them and containing proportionately more don't care entries.

Quite clearly whether a flip-flop switches on the $1 \rightarrow 0$ clock edge of $0 \rightarrow 1$ clock edge must be taken into account in the final design. Equally clearly a clock source could be derived from a Q or $\overline{Q}$ terminal. Provided the truth table specifying flip-flop response is the same for each type of flip-flop the selection of a $0 \rightarrow 1$ active clock edge or $1 \rightarrow 0$ active clock edge flip-flop could be left to the final design stage. However, since either edge is available if both Q and $\overline{Q}$ terminals are present on all flip-flops one may as well standardize on one type of flip-flop right at the start.

With relevance to computer-aided design a research student at the University of Salford, E. Krumm, has written programs for synchronous counter synthesis in terms of J–K flip-flops using KDF9 *Algol*. It should be fairly easy to program a computer to scan through a sequence to determine whether suitable internal clock-sources are available. If so, the routines used in the synchronous design program could be applied to the relevant stages of the asynchronous counter design. If not, then a straight synchronous design would be available anyway.

H. DUNDERDALE

Department of Electrical Engineering,
University of Salford,
Salford, M5 4WT.
8th December 1972

References

1. Dunderdale, H., 'A formal method for designing sequential circuits from synchronous flip-flops such as the J–K type using a hybrid synchronous-asynchronous technique', *The Radio and Electronic Engineer*, 42, No. 9, pp. 416–8, September 1972.
2. Marcus, M. P., 'Switching Circuits for Engineers', pp. 273–98. (Prentice-Hall, Englewood Cliffs, N.J., 1967.)
3. Givone, D. D., 'Introduction to Switching Circuit Theory', pp. 371–87. (McGraw-Hill, New York, 1970.)
4. Volker, M. and Majithia, J. C., 'The Use of APL Software for Sequential Counter Designs', Department of Electrical Engineering, University of Waterloo, April 1972.
5. Dean, K. J., 'The design of parallel counters using the map method', *The Radio and Electronic Engineer*, 32, No. 3, pp. 159–62, September 1966.
6. Dunderdale, H., 'An Introduction to Formal Methods in Logical Design', Part II, Sequential Systems', University of Salford, April 1972.

STANDARD FREQUENCY TRANSMISSIONS—January 1973

(Communication from the National Physical Laboratory)

January 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	−0·1	0	+0·2	736	657·2
2	+0·2	−0·1	+0·2	734	658·0
3	0	0	+0·2	734	658·0
4	+0·1	+0·1	+0·2	733	657·3
5	0	+0·1	+0·3	733	656·4
6	0	0	+0·1	733	656·2
7	0	0	0	733	656·2
8	0	+0·1	0	733	655·7
9	0	0	0	733	655·4
10	0	0	0	733	655·2
11	0	0	0	733	655·4
12	−0·1	0	0	734	655·6
13	−0·1	−0·4	0	735	656·8
14	0	0	−0·1	735	657·5
15	−0·2	−0·1	−0·1	737	658·4
16	0	0	−0·1	737	658·6
17	−0·1	0	−0·1	738	658·6
18	0	0	0	738	659·5
19	−0·1	0	−0·1	739	659·8
20	+0·1	−0·1	−0·1	738	660·6
21	−0·1	0	−0·1	739	660·4
22	+0·2	+0·1	−0·1	737	659·1
23	0	0	0	737	658·6
24	+0·1	+0·1	0	736	657·7
25	+0·1	+0·1	0	735	656·6
26	+0·1	+0.1	0	734	656·0
27	+0·1	+0·2	−0·1	733	654·5
28	0	0	0	733	654·3
29	+0·1	+0·1	0	732	653·5
30	+0·2	+0·1	0	730	652·1
31	0	0	0	730	651·7

All measurements in terms of H.P. Caesium Standard No. 334, which agrees with the N.P.L. Caesium Standard to 1 part in 10^{11}.

* Relative to UTC Scale; $(NTC_{NPL} - \text{Station}) = +500$ at 1500 UT 31st December 1968.

† Relative to AT Scale; $(AT_{NPL} - \text{Station}) = +468{\cdot}6$ at 1500 UT 31st December 1968.

IERE News and Commentary

Education and Training for Multidisciplinary Subjects

In conjunction with the Department of Trade and Industry, CEI is sponsoring a discussion meeting on 'Education and Training for Multidisciplinary Subjects' for all members of the Engineers' Registration Board and other interested people. The meeting will be held at the Institution of Mechanical Engineers, 1 Birdcage Walk, Westminster, London SW1, on Thursday, 29th March at 5.30 p.m.

This is the first of a series of events dealing with the multidisciplinary subjects described as Industrial Technologies, which include Tribology, Terotechnology, Corrosion Technology and Materials Handling. It has been estimated that greater awareness of these four technologies could bring benefits of up to £1500 million in the manufacturing services and distributing industries.

The key to a proper appreciation of the potential savings is education and training. Mr. Gilbert A. Hunt, C.B.E., Chairman of the Committee for Industrial Technologies, will introduce the general aims before handing over to specialist speakers, including Dr. H. Peter Jost, C.B.E., Maj.-Gen. Sir Leonard Atkinson, K.B.E. (Past President of the IERE), Mr. J. B. Cotton and Dr. A. H. Chilver, Vice-Chancellor of Cranfield Institute of Technology. The speakers will discuss how government, industry, learned societies and educational bodies can foster or provide education and training in these multidisciplinary subjects.

Mr. Philip Horton, M.V.O.

The many members who have supported the Institution's Benevolent Fund will know that over the years a close connexion has been built up with Reed's School. The appointment of the recently-retired Secretary of the School, Mr. Philip Horton, as M.V.O. (4th Class), which was announced in Her Majesty's New Year's Honours List, is therefore a cause for pleasure.

Since her wedding in 1947, the Queen has been nominating two boys each year to receive Queen's Bursaries for entry to Reed's. The boys chosen are usually the sons of Servicemen killed in action, or boys in whom Her Majesty has some personal interest.

With the always ready co-operation of Mr. Horton, several children who were dependants of beneficiaries of the Institution's Fund have been nominated to attend the School and, in addition, Bursaries for future nominations have been donated.

Reed's School, which has been at Cobham, in Surrey, since soon after the 1939–45 war, is a public school with extensive facilities for science teaching at sixth-form level. Mr. Horton's retirement on January 1st this year completed nearly 30 years' service to the School.

MacRobert Award Lecture

Mr. Godfrey Hounsfield, the winner of the 1972 MacRobert Award, will deliver a lecture on 'The New X-Ray Technology for Sensitive Discrimination of Soft Tissue' at the Institution of Electrical Engineers, Savoy Place, London WC2 on Thursday, 12th April, at 5.30 p.m. The lecture is being sponsored by the Council of Engineering Institution.

The EMI Scanner System developed by Mr. Hounsfield at the Central Research Laboratories of EMI was described as 'epoch-making', because it broke away from the photographic techniques for recording X-ray pictures which had remained unchanged in principle since Röntgen's day. A brief description of the techniques was given in the January-February issue of *The Radio and Electronic Engineer*.

The MacRobert Award Lecture is open to all. Tea will be served from 5 p.m. and a film will be shown during the course of the lecture.

Student Recruitment—Evidence Needed

The CEI Working Party on Student Recruitment, recently set up under the chairmanship of Sir Leonard Atkinson (Past President of the IERE), would be glad to hear from anyone who has knowledge of articles, research work or papers on the problem of student recruitment and the use of qualified manpower.

The Working Party has access to the annual statistical returns from Government and other official sources. Additionally, it would like to know of any local surveys or reports which are available and which give information on the following subjects:

Numbers entering degree courses in engineering; entry qualifications of these students; numbers and types of degrees awarded; numbers of HND and HNC candidates and numbers of awards made; and information on the first employment of university graduates HND and HNC holders.

Any member having information of this nature is invited to write to the Secretary of the Working Party, Miss R. Winslade, O.B.E., at the Council of Engineering Institutions, 2 Little Smith Street, London, SW1P 3DL.

SERT Elects its First Vice-President

The Society of Electronic and Radio Technicians has elected Sir Cyril English, C.Eng., F.I.Mech.E., Director General of the City & Guilds of London Institute, as its first Vice-President.

Sir Cyril trained as a mechanical engineer and was for six years a technical teacher. Following war service in the Royal Navy, where he reached the rank of lieutenant commander (E), he joined the Inspectorate of the Ministry of Education and became Senior Chief Inspector of Further Education in 1965; in 1968 he was appointed to his present post. He has served on many national committees concerned with education and training and for the last two years he has been Chairman of the British Association for Commercial and Industrial Education (BACIE).

The President of SERT is Mr. James Redmond, C.Eng., F.I.E.E., Director of Engineering of the British Broadcasting Corporation.

Television Standards Converter Using Digital Techniques

Conversion of television pictures from one international standard to another has proved an extremely complex problem, because the various television standards differ in the number of lines, the number of fields and in the colour systems. It has become clear that the only really satisfactory way of achieving high-quality conversion is to process each picture point separately, and it is only recently that computer-type digital logic has been able to work fast enough to operate on the signal in real time.

What is claimed to be the world's first television picture converter to use digital techniques for changing American or Japanese television signals into European television signals has recently been demonstrated by engineers of the Independent Broadcasting Authority. The new equipment shows marked improvements in the quality of the output colour pictures, is smaller than existing equipments, has no line-up adjustments and is confidently expected to prove itself completely stable in operation.

Last year the IBA completed its first digital standards converter which changes 625-line colour television signals into 405-line black-and-white signals, and this line-frequency converter was used operationally for a trial period at the main IBA London v.h.f. transmitter. The latest extension of this work solves the much more complex problem of changing from one colour television system using 30 pictures per second—as in America—to another using only 25 pictures per second—as in Europe. With its negligible picture distortion in the conversion process, the new equipment will satisfy a global demand for better quality in the satellite exchange of programmes between countries.

The new IBA digital field-rate converter is a small device occupying no more than two six-foot equipment racks, which is to be compared with the seven racks of earlier electronic converters. It converts 525-line, 30 pictures-a-second, NTSC colour television signals into 625-line, 25 pictures-a-second, colour-separation (red, green and blue) signals which are then coded into either PAL or SECAM signals in a normal, analogue, coder.

The new converter uses digital television techniques and the standards conversion process is to all intents and purposes free of distortion. It uses a number of novel techniques for decoding the NTSC signal and for interpolating between fields and lines. The converter will eventually be bi-directional (i.e. 625/50 to 525/60) but this facility has not yet been added.

Historical

The earliest type of field converter to be used for changing 60-field signals into 50-field signals was an optical converter which used a camera, operating on one scanning standard, to look at a cathode-ray-tube picture display operating on the other scanning standard. A later development for colour television has been to use two such arrangements, with one camera-to-display unit converting the luminance picture and a second camera-to-display unit converting the coded chrominance signal. In general, optical standards converters have problems with halation, spot-size, lag, phosphor grain pattern blemishes, the compromise between phosphor decay-time and movement judder, chrominance resolution and registration between luminance and chrominance.

All-electronic analogue converters avoid some of the problems of optical converters but have been, to date, much more expensive, and much larger. If they use quartz delay-lines then the delay-lines need to be operated in a controlled environment to avoid variations in delay-time with temperature, while starting from cold is not practicable. The write-in and read-out rates have to be the same, and organizational problems arise in continually switching between numerous lines of different delay-times. Spurious signals in the delay-lines cause picture defects, especially on high-purity colours, and attenuation in the lines, coupled with the need for amplitude and group-delay equalization, make the devices electrically noisy and tedious to line-up. Since the subcarrier which emerges from the delay-line is used to judge the time delay, the noise on the subcarrier signal limits the timing accuracy of the output. The analogue separation of chrominance and luminance is imperfect and gives rise to further picture defects.

Advantages of Digital Technology

Once the television signal has been converted from its normal analogue waveform to a data stream of binary digital pulses, considerable improvements can be made in the conversion process. The cost of such complex data processing depends upon the sophistication and technical elegance of the computation logic which is developed, and the present IBA converter is at once notable for the economy of its means and for the precision of its numerical calculation. Considered as a special-purpose computer it is believed to be the fastest in the world so far: its operating speed represents 3×10^{12} digit movements per second!

The storage elements are—as for the line converter—computer shift registers using m.o.s. f.e.t. technology. Such storage of digital data in shift registers is impeccable and there is no degradation of the signal in the store. At the same time, the write-in and read-out rates can be different, and the delay between writing and reading is flexible and can be precise.

The machine has an instantaneous warm-up and there are no preset controls in the digital conversion circuits. An ingenious feature of the design is the type of spatial filter which is used, among other operations, to produce the luminance, and the I and Q chrominance signals, from the incoming composite data stream. Inherently, the spatial filters have perfect tracking and there are no adjustable controls.

The digital parts of the equipment are designed mathematically for a given performance and they calculate to their design accuracy. Although—as in any equipment—parts can fail, it is not possible for the performance to drift away from its design tolerance, and the converter will, for example, always give a k-rating factor of better than 1%.

Although the converter uses nearly 8000 integrated circuits and the shift registers represent some 15 million transistors, an important element of the design is the repetition of circuits, i.c.s and boards, which has reduced costs and simplifies repair work. Most of the circuit boards can be tested on plug-in go/no-go test sets which not only show if a fault is present but also indicate its location. The cost of integrated circuits is falling steadily and the machine should become even cheaper.

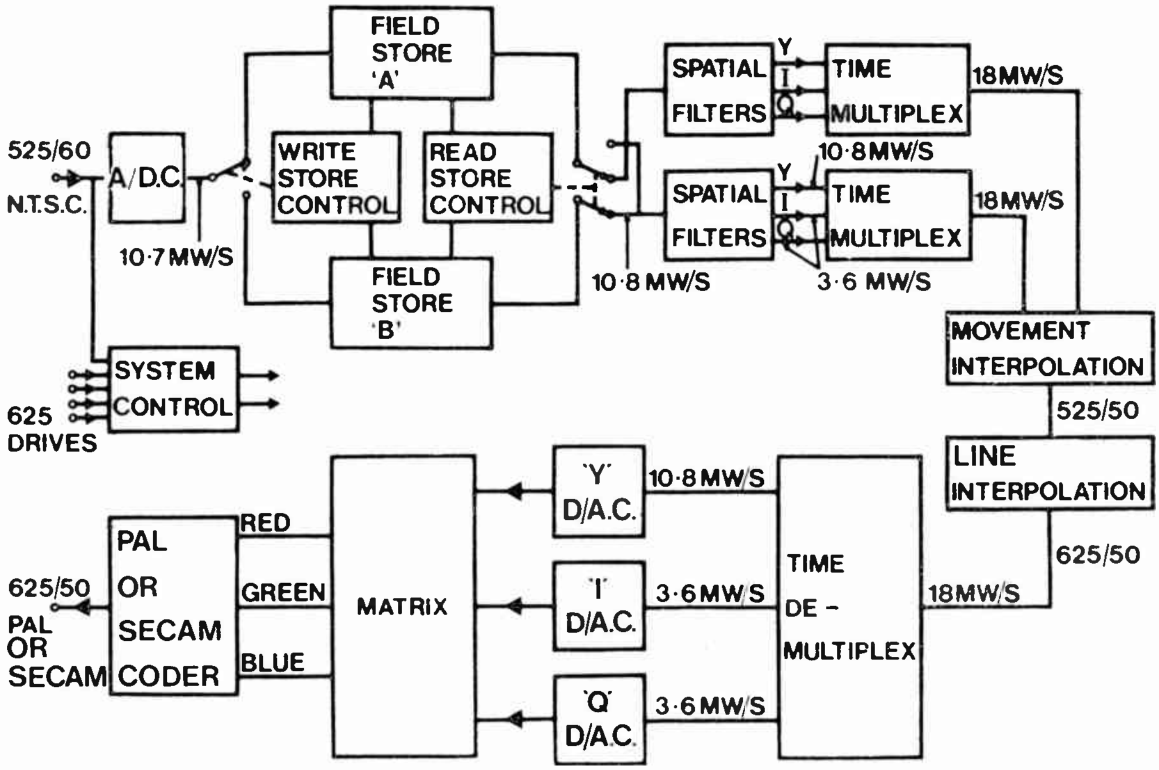


Fig. 1. Block schematic of the converter.

Principles of the Field-Rate Converter

There are three main functions which such a converter has to carry out. It must change the field rate from 60 Hz to 50 Hz (and lengthen the duration of the field period), it must change from a 525-line scanning rate to a 625-line scanning rate, and it must change both the colour subcarrier frequency and the modulation of the subcarrier.

The operation of the digital field-rate converter will be described with reference to the block schematic of Fig. 1. After synchronizing signals have been fed to the timing control circuits, the first operation of the converter is to change the incoming coded analogue video signal into a digital signal in an analogue-to-digital converter. The sampling rate chosen is three times the subcarrier frequency (10·7 MHz) since this rate is higher than the Nyquist requirement for waveform accuracy and enables several novel techniques to be used in the subsequent processing. Each picture element so sampled is then analysed into one of 256 equally-spaced amplitude levels to produce an eight-bit binary-coded output pulse-train, or word. To reduce the maximum operating-speed requirement of the solid-state devices in the following circuits, eight-wire parallel circuit feeds are used from this point onwards—and further down the chain this is increased to 24 parallel feeds, and later to 32.

Now consider two cameras panning, together, round a scene at an angular rate of 6° of arc per second of time. Suppose that one camera is operating on the American 525-line system, whilst the second camera is on the UK 625-line standards. Presume that suitable optical arrangements ensure that both cameras see the same view. Then the American camera with a 60 Hz field frequency will be panning at the rate of 0·1° per television field, while the British camera with the lower field frequency of 50 Hz will be panning at the rate of 0·12° per television field. That is to say, the amount of horizontal movement per television field is different in the two television systems and it is this difference which the movement interpolator must create. It does this by combining, in appropriate proportions which differ from one field to the next, the signals from two sequential fields of the incoming signal. It is thus necessary to provide storage facilities in shift registers for two complete fields.

When the fields on both input and output systems coincide, then picture information from field one of the incoming signal may be used directly to produce the first field of the outgoing signal. However, the second field of the outgoing signal must show a greater change of position of moving objects than the second field of the incoming signal. To achieve this, the outgoing signal is made up of a (large) proportion of picture information from the second incoming field and a (small) proportion of picture information from the third incoming field. Similarly, the third outgoing field consists of a combination of information from the third and fourth incoming fields, and so on. The whole process repeats 10 times a second, which is the approximate common repetition frequency of the two standards involved.

Of course, in order for such a combination of signals to be satisfactory, the signal from two sequential incoming fields must at any moment relate to identical picture points. This is not the case when the signal is generated because of the interleaved relationship between succeeding picture fields.

Information is read out of the two field stores simultaneously, at a rate depending on the required outgoing line frequency. Each field store is made up of 48 modules arranged in groups of four and each group of modules stores 21 lines of picture information using a total of 96 shift registers each of 1024 bit capacity. The outgoing line-time is very nearly the same as the incoming line-time, and the rate is now 10·8 megawords a second. However, because of the difference in field rate required, there are at this stage some line gaps so that 21 lines occupy the time required for 25.

The signal from each field store then goes to a spatial filter, which has three functions. Firstly, it modifies the output signal in such a way as to cancel out the effects of interlace between sequential incoming fields. Secondly, it separates the luminance and the chrominance components of the signal without any loss of either vertical or horizontal resolution—at any rate for still pictures. Thirdly, it produces from the chrominance signal obtained, I and Q components. The output signal from the spatial filters consists then of three separate parts: a train of luminance signal values Y; a train of chrominance signal components I; and a train of chrominance signal components Q. These three trains of pulses are interleaved in the time multiplexer. Because the I and the Q signals are of narrower bandwidth than the luminance signal, a slower sampling rate is sufficient and the time multiplexer produces a sequence of output signal pulses in the form of Y, I, Y, Q, Y; Y, I, Y, Q, Y, ... The data rate at this point is approximately 18 megawords per second.

As shown in the block diagram, the spatial filters, which are arranged as complementary pairs, are fed in turn from the field stores A and B. In the spatial filters, separate matrices are used to generate the luminance and the chrominance signal. The matrices examine at one time a series of 15 sampling points arranged as five points in three sequential lines. Appropriate proportions of the signals at each of these 15 points are taken to produce either a luminance signal without a chrominance component or a chrominance signal without a luminance component. The chrominance signal is separately demodulated to produce the I and Q components.

The movement interpolator takes proportions of each of the multiplexed signals and for every six incoming fields produces the appropriate five outgoing fields. The ratio of the two standards is not precisely 5/6 and the actual operation is somewhat more complex. At the output of this movement interpolator the signal consists of 525 lines of picture, but running at the 50 field rate.

The function of the line interpolator is to generate from these 525 lines a sequence of 625 lines each with the information appropriate to its position in the picture. This fills in the line gaps mentioned above. An interpolation process is again used. The picture points in each outgoing line are produced by selecting appropriate proportions of the corresponding signal points from three incoming lines.

The time demultiplexer separates the luminance pulses from the I and the Q information and each is separately converted back from the digital to the analogue form. The three signals obtained are then matrixed to produce standard R, G, B, signals and the composite colour signal is finally obtained using a standard colour coder, either a PAL coder or a SECAM coder.

As was mentioned above, the reading rate for the field stores depends directly on the outgoing line frequency required. The converter is designed to be locked by station synchronizing signals in the same way as a camera or any other picture source.

Commenting on the importance of the new picture converter, Howard Steele, IBA's Director of Engineering, stated: 'Although the work has been undertaken as part of a long-term investigation into digital techniques for television, it opens the way to operational applications which could improve and make less critical the interchange of programmes and programme material between countries and continents. The use of computer-type metal-oxide semiconductor integrated circuits and simple on/off digital waveforms is expected to eliminate many of the critical adjustments and maintenance problems of existing analogue-electronic and opto-electronic standards converters. It is a significant step forward in television engineering.'

This latest application of digital techniques clearly shows the way ahead for the next decade or so, as far as the broadcasting authorities are concerned. As well as their line standard converter using digital techniques, the IBA have put into service a system for transmitting data in digital form in the vertical interval of the television picture, while the BBC for their part are actively developing p.c.m. techniques for network transmissions and for a variety of applications within the studio complexes.

References

1. Baldwin, J. L. E., 'New techniques in television standards conversion', International Broadcasting Convention 1969. (IEE Conference Publication 1969, pp. 198–200.)
2. Baldwin, J. L. E., Stalley, A. D. and Kitchin, H. D., 'A standards converter using digital techniques', *R. Telev. Soc. J.*, 14, No. 1, pp. 3–11, 1971.
3. Baldwin, J. L. E. and James, A., 'Television standards conversion using digital processing'. Asian Broadcasting Union Engineering Committee meeting, Shiraz, Iran, October, 1971.

Chairmen of D.T.I. Requirements Boards Appointed

The names of Chairmen to head five of the six Requirements Boards which are being set up to implement the customer/contractor principle in industrial research and development supported by the Department of Trade and Industry have been announced.

The appointments are:

Chemical and Mineral Processes and Plant	Mr. Donald Clark (Under Secretary DTI)
Computers, Systems and Electronics	Mr. John Nichols (Under Secretary DTI)
Engineering Materials	Mr. John Crane (Director, Imperial Metal Industries Ltd.)
Mechanical Engineering and Machine Tools	Mr. John Atwell (Director, Weir Group Ltd.)
Ship and Marine Technology	Mr. Nigel Broakes (Chairman, Trafalgar House Investments Ltd.)

The Chairman of the Standards and Metrology Board will be announced later.

Mr. John Winfrith Nichols, B.Sc., C.Eng., F.I.E.E., who is appointed Chairman of the Computers, Systems and Electronics Requirements Board, is the Under Secretary in charge of the similarly named Division of the Department of Trade and Industry. He previously served in the Systems and Automation Division of the Ministry of Technology, and from 1959 to 1965 was with the UK Atomic Energy Authority. From 1955 to 1959 he was Chief Research Officer, Corporation of Trinity House; he was co-author of a paper, with Mr. A. C. MacKellar, on 'Some recent developments in marine navigational aids', published in the Institution's Journal in July 1962.

In its White Paper 'Framework for Government Research and Development' (Cmnd 5046, July 1972) the Government reaffirmed its intention to implement the customer/contractor principle recommended by Lord Rothschild for the control of Government R & D in the Green Paper with the same title (Cmnd 4814, November 1971).*

The White Paper said '. . . departmental customers must work in partnership with their research and development "contractors", whether inside or outside the Department. Responsibilities are then clear. Departments, as customers, define their requirements; contractors advise on the feasibility of meeting them and undertake the work; and the arrangements between them must be such as to ensure that the objectives remain attainable within reasonable cost. This is the customer/contractor approach.'

Accordingly, the DTI is setting up the six Boards referred to above, each concerned with customer interests in its appropriate technological field. The function of the Boards will be to determine, in agreement with the Minister for Aerospace, the objectives and balance of the Department's intra-mural and, where appropriate, extra-mural R & D programmes in the relevant technical field within the funds available.

R & D programmes funded by DTI are of two kinds: those undertaken to support Government's own functions and those to support industry. They cover a wide area of technology outside the special areas of Aerospace and Nuclear Energy. Initially each Board will be responsible for research expenditure of about £2M.

The contractors for this R & D include the DTI's own industrial research establishments, namely

National Physical Laboratory, Teddington
National Engineering Laboratory, East Kilbride
Warren Spring Laboratory, Stevenage
Safety in Mines Research Establishment, Sheffield and Buxton
Laboratory of the Government Chemist, London

as well as the UK Atomic Energy Authority (for non-nuclear applications)

the grant-aided Research Associations, industrial companies, independent consultant laboratories, universities and Research Councils.

Boards will generally consist of 12 members and will include six independent members from industry, two independent scientists, and appropriate official representatives of DTI main customer divisions, other Government departments, the main contractors (including Research Councils) and staff from DTI research divisions. A new Research Requirements Division under the Department's Chief Scientist will provide technical, management and financial back-up services for the Boards.

* See *The Radio and Electronic Engineer*, 41, p. 525, December 1971, and 42, p. 349, August 1972.

Microwaves across the Roof of Europe

British Equipment for Swiss Television Network

GEC Telecommunications Limited of England is supplying 2 GHz and 6 GHz microwave-radio equipment to the Swiss Posts Telephones and Telegraph Department for its television network. It will be installed at 16 locations in various parts of Switzerland. Through Hasler AG of Berne, GEC has supplied most of the transmission equipment for the Swiss television network and this order represents the company's eighth extension to the system. Thirty-nine GEC-equipped microwave-radio stations have now been installed covering a route length of more than 6437 km (4000 miles).

GEC microwave-radio equipment was first installed in Switzerland in 1954 as part of the Eurovision link. The equipment linked Germany and Italy and was the first microwave-radio route across the Alps: the repeater station on the 3658 m (12 000 ft) Jungfraujoch was at the time the highest radio station in the world. The adjoining picture shows the installation on the Santis Mountain (3352 m, 11 000 ft), which is covered with snow and ice for most of the year.

The number of television sets in Switzerland has grown from less than 100 000 in 1954 to today's figure of well over 1·4 million. Swiss television has programmes in French, German and Italian to cater for the country's three main languages.

Convention of National Electrotechnical Societies of Western Europe

In recent years there have been a number of moves to increase the collaboration between professional engineers both at national and international level. Thus in the United Kingdom there is the Council of Engineering Institutions, in Europe there is the Fédération Européenne d'Associations Nationales d'Ingenieurs (FEANI) while at trans-continental level there is the World Federation of Engineering Organization (WFEO). All of these organizations are multi-disciplinary and, while they have an extremely important role to play, they cannot, by their nature, meet all the detailed needs peculiar to individual disciplines.

It was against this background, and coupled with the increasingly close ties between European countries that, early in 1971, the Presidents of several European Electrical and Electronic Engineering Institutions met informally, at the invitation of the Association Suisse des Electriciens, to discuss ways and means of improving collaboration between their Institutions.

As a result a small Working Party was formed to draft a set of Articles for a 'Convention of National Electrotechnical Societies of Western Europe'. Membership of the Convention is open to all professional electrical/electronic institutions in Western Europe and the primary objective of the Convention is to facilitate the exchange of information and the fostering of multi-lateral collaboration between the member Societies, including the organization of joint events between two or more of the members.

A Constituent Assembly of the Convention was held in Zurich on November 24th last and representatives from professional electrotechnical institutions in the following countries were present:

Austria, Belgium, Denmark, Finland, France, Germany Italy, Netherlands, Norway, Spain and Switzerland.

The United Kingdom was represented by:

The Institution of Electronic and Radio Engineers:
Professor W. Gosling, Vice-President
R. C. Slater, Deputy Secretary

The Institution of Electrical Engineers:
Professor J. F. Coales, Past President
Dr. G. F. Gainsborough, Secretary

The Constituent Assembly was presided over by M. R, Richard, President of the Association Suisse des Electriciens and at this meeting the Articles of the Convention were agreed and signed by all the Institution representatives present.

Following the Constituent Assembly the first General Assembly was held. At this meeting it was agreed that M. Richard should continue as President for the remainder of 1972 and the President of the Société Royale Belge des Electriciens, M. J. Schrans, should be President of the Convention for 1973.

It is envisaged that a General Assembly of the Convention will be held annually and in 1973 this will take place in Brussels. In the meantime the aims of the Convention will be furthered by a seven-man Executive Committee on which the UK is represented by Professor J. F. Coales. Among the more immediate objectives are:

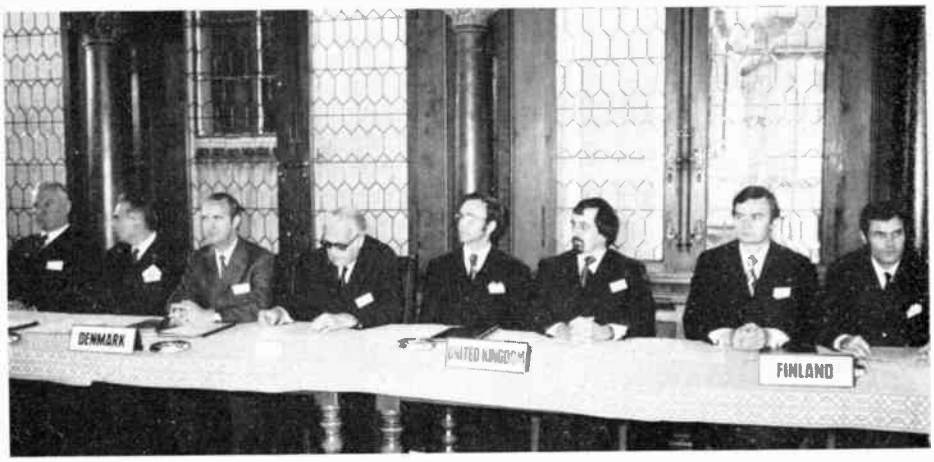
(a) The drafting of Bye-Laws for the working of the Convention.

(b) Consideration of activities to encourage the interest of younger members on an international basis.

(c) The organization of international events and particularly Eurocon 73 and Eurocon 74.

The first technical meeting to be held under the auspices of the Convention will take place in Geneva on 7th May, 1973 and has been organized by the European Organization for Nuclear Research (CERN). The subject will be 'The electrical and electronic environment of new bubble chambers' and following the presentation of papers by members of CERN in the morning there will be visits to some of the bubble chamber and accelerator installations at Meyrin.

Further information may be obtained from the Deputy Secretary of the IERE, 3 Bedford Square, London WC1B 3RG, or from the Association Suisse des Electriciens, Seefeldstrasse 301, 8008 Zurich.



On the left the President of the Association Suisse de Electriciens, M. R. Richard, addresses the Constituent Assembly of the Convention in the medieval Guildhall Zunfthaus zur Zimmerleuten at Zurich. In the other photograph, some of the 29 representatives are shown, among them the IERE representatives Professor Gosling (third from the right), and next to him Mr. Slater.

Members' Appointments

CORPORATE MEMBERS

Mr. Peter L. Mothersole (Fellow 1961, Member 1957. Graduate 1952) has been appointed Commercial Chief Engineer of Mullard Ltd. Mr. Mothersole joined Mullard Research Laboratories in 1953 and moved to the Central Application Laboratory at Mitcham eleven years later as group leader and head of the consumer applications activity. From 1969 until his present appointment he was engineering manager and a member of the executive management team of Pye TV. He has contributed numerous papers to the Institution's Journal and other publications.

Mr. Mothersole has represented the IERE on the Management Committee of the International Broadcasting Convention since 1968 and is currently chairman of its Programme & Papers Committee. He has served on the Communications Group Committee since its formation as the Television Group Committee in 1960 and was chairman from 1964 to 1969. He serves on the IEE Electronics Divisional Board and chairs one of its Professional Groups.

Mr. R. A. Bent (Fellow 1957, Member 1951) has recently been appointed Managing Director of W. F. Parsonage & Co. Ltd., radio and electrical engineers of Walsall. He was apprenticed to the English Electric Company and subsequently was concerned with automatic weapon control and early work in industrial electronics. In 1946 Mr Bent founded British Electronic Products and, following its association with a succession of larger groups, served as Managing Director of Lancashire Dynamo Electronic Products and, more recently, of Thorn Automation. He has taken an active part in the affairs of the West Midlands Section and was for several years Local Honorary Secretary.

Mr. J. M. Peters, M.Sc.(Eng.), (Fellow 1963, Member 1964, Graduate 1951) has recently been promoted to Senior Principal Production Engineer in the Ministry of Defence and appointed Deputy Director in the Directorate of Standardization.

Mr. Peters was Chairman of the Southern Section in 1964 to 1965 and is currently a member of the Institution's Professional Activities Committee, having served on the Technical Committee for several years.

Mr. K. A. Baker (Member 1971) formerly Senior Electronic Engineer with Cartner Group Ltd., Kenton, Middlesex, has joined STC Ltd., Electronic Switching Division, as Senior Design Engineer, in their Test Equipment Group at New Southgate.

Mr. G. H. Boghossian (Member 1965, Student 1959) has joined Manarp Electronic Instruments Ltd., Hayes, Middlesex, as Technical Director.

Mr. R. Carrier (Member 1971, Graduate 1966, Student 1965) has joined Perkins Engines Company, Peterborough, as Head of Instrument Facility. He was formerly Head of Instrumentation Group, CAV Ltd., London.

Major W. J. Cleeve, REME, (Member 1970, Graduate 1969) has been appointed Officer Commanding 94 LOC Regiment Workshop, REME, BAOR, Germany.

Mr. D. K. Craig (Member 1972) has been awarded the Wakefield Gold Medal of the Royal Aeronautics Society jointly with two colleagues. This Medal, which is awarded 'for contributions towards safety in aviation' was presented on 7th December last in recognition of the work of the three recipients in the development of the Autoland system in *Trident* aircraft, which led to its official acceptance for airline services to category 3A weather standards. Mr. Craig has been with British European Airways for the whole of his professional career. He is currently a Senior Development Engineer in the Design and Development Branch.

Mr. R. Cropper, B.Sc., A.R.C.S. (Member 1971), previously a Senior Engineer with Plessey Company at Roke Manor, has moved to Smiths Industries Ltd. as a Project Engineer in the Access Systems Unit at Hemel Hempstead.

Captain W. J. Crossley, RAN (Member 1963, Graduate 1958) has been posted to H.M.A.S. *Watson*, Watson's Bay, Sydney, as Commanding Officer.

Mr. M. R. Jackson (Member 1958, Graduate 1951) formerly Principal Engineer, Magnetic Recording Head Design with Data Recording Instrument Co., is now a Lecturer at Southall College of Technology.

Sqn. Ldr. J. H. Goodfellow, M.A., RAF (Member 1972, Graduate 1964), formerly Peripheral Services Engineer, Directorate of Data Processing. National Air Traffic Service, has been posted RAF Buchan, Aberdeenshire as Senior Engineering Officer.

Sqn. Ldr. W. A. Gossage, RAF (Member 1971, Graduate 1969, Student 1959) has been appointed to Electrical Engineering 1c (RAF) with the Ministry of Defence, London. Sqn. Ldr. Gossage was formerly Officer-in-Charge Radar Support Squadron, RAF Staxton Wold.

NON-CORPORATE MEMBERS

Mr. A. A. Ajilogba (Graduate 1970), formerly a Telecommunications Engineer with Shell-BP Co. Ltd. in Warri, Nigeria, has now joined Siemens Nigeria Ltd. as Telecommunication Engineer at Lagos.

Sub. Lieut. B. G. Carter, RN (Graduate 1970, Student 1966) Weapons Electrical Officer in HMS *Forth*, has been appointed Weapons Electrical Officer, HMS *Onyx*, Portsmouth, in the rank of Lieutenant.

Sub. Lieut. G. S. Clark, RN (Associate 1970, Student 1965) has been appointed Supply Officer, Supply Squadron, RAF Upavon, HQ 46 Group, Wiltshire, in the rank of Flight Lieutenant, RAF.

Mr. I. S. I. Ezerendu (Associate 1972) has been appointed an Electronics Technologist in the Department of Physics, University of Nigeria, Nsukka. He was previously Assistant Technical Officer, P & T Department, Makurdi, Nigeria.

Mr. P. A. Foster (Graduate 1967) has joined Welding Alloys Ltd., Royston, Herts, as Product Manager. He was previously a Project Engineer with Metals Research, Melbourn, Herts.

Mr. M. R. Jones (Graduate 1971) has been appointed a Principal Engineer, Process and Communications Development, International Computers Ltd., Stevenage. He was with Cincinnati Milacron, Process Controls Division, Biggleswade, Bedfordshire as a Senior Electronic Design Engineer.

OBITUARY

The Council has learned with regret of the death of the following member:

Henry Armstrong (Member 1940, Associate 1939, Student 1933) died suddenly on 13th December at the age of 63, following a heart attack in Newcastle. After studying at the North Eastern School of Wireless at Jesmond, Mr. Armstrong was engaged by the Marconi Company as a Marine Radio Officer from 1929 up to 1937, when he became an Air Ministry Radio Instructor at Cranwell. In 1939 he took over the Principalship of his former School, which was a Government-licensed establishment, and from 1949 to late 1951 he was Principal of the Northern Counties Wireless School at Preston, Lancashire. For a short time Mr. Armstrong was with the Rediffusion Group of Companies. He joined the Inspectorate of Electrical and Mechanical Engineering, Ministry of Supply, in Newcastle in 1951, and then in 1956 rejoined the Marconi Company at Gateshead as a Technical Author. He continued in similar work to the time of his death, although indifferent health prevented full-time activity towards the end of his life.

He was a tower of strength in the North Eastern Section in its early days, being Section Secretary from 1940 to 1945. He took a keen interest in the activities and the advancement of recognition of the Institution and attended Section meetings whenever possible.

New Books Received

All the books which are described below are available in the Library and may be borrowed by members in the United Kingdom. A postal loan service is available for those who are unable to call personally at the Library

Theory and Design of Digital Computers

DOUGLAS LEWIN. Nelson, London, 1972. 23·9 cm × 15·9 cm. 383 pp. £4·95.*

CONTENTS: The stored program principle. Principles of machine-code programming. Input compilers and procedure orientated languages. Control unit organization. The arithmetic unit. Storage systems. Input/output systems. Engineering and systems aspects. Highly parallel processing systems.

Professor Lewin has chosen a comprehensive title and topic and he has certainly produced a book worthy of the name. It contains a great deal of meat in nearly 400 packed pages making it difficult to do justice to in a brief review.

In trying to be comprehensive it is often difficult to know where to start and stop, and Professor Lewin has wisely not included a section on circuitry as integrated circuits provide this boundary nowadays. There is very good cover of both logical design and machine organization structures. Programming techniques are also well dealt with at a fairly basic level.

However, a book that is intended and will undoubtedly be used, for teaching has certain responsibilities, particularly in giving the right balance. In this context the Iverson notation is presented in a depth that really is not justified by its acceptance in practice, however elegant the technique may be. The student should also be warned that although there is a good chapter on input/output systems, the example given is by no means typical and indeed represents one of the poorer designs. Finally, while still on the critical note, the examples of Parallel processing are all from the U.S.A. and the very interesting Manchester *MU5* approach is not even mentioned.

Inevitably a review tends to highlight adverse comment and anything said here should not be considered as overall criticism of what is a very well thought out book. Its general balance and depth of treatment is excellent. Certainly anyone who has thoroughly understood the material can consider himself well qualified in the subject.

A. ST. JOHNSTON

(Professor Lewin (Member 1960) is in the Department of Electronics at Brunel University.)

Modern Data Communication

WILLIAM P. DAVENPORT. Pitman Publishing, London, 1972. 22·7 cm × 15 cm. 200 pp. £2·75.*

CONTENTS: Introduction to data transmission. The language of data. Coding for communications. Characteristics of transmission media. Efficiency and error control. Modulation and multiplexing. Commercial communications channels and services. Switching and network concepts. Data-set uses and characteristics.

This book is written to explain the theories and concepts of data transmission to the operators and managers of information systems, and the students of information handling. An attempt has been made to reduce intricate theories to clear and understandable terms without losing the essence of their meanings. To this end, only simple arithmetic is used, and all terms, calculations and assumptions have been explained in some detail.

(Mr. Davenport is Vice-President and General Manager of Scott-Buttner Communications Inc., Alameda, California.)

Physical Electronics

J. SEYMOUR. Pitman Publishing, London, 1972. 21 cm × 14·4 cm. 438 pp. Paperback. £3·25*

CONTENTS: Electrons in atoms. Electrons in crystals. Contacts between materials and p-n junctions. Junction transistors with uniform and graded bases. Field-effect transistors and other semiconductor devices. Electron emission and vacuum devices. Electrons in gases. Microwave devices and electrical noise. Masers and lasers. Travelling waves. Wave mechanics, an introduction. Density of energy levels in a semiconductor. Relativistic mass increase. Electric field between a cylinder and a coaxial wire.

Based on a series of lectures, the level of this book is about the second or third year of an undergraduate course and it is also appropriate for CEI or HNC students. The emphasis is on solid-state devices and the author in general follows the approach of first giving a simple physical theory for a device, then deriving an equivalent circuit taking into account external factors, and finally discussing applications.

(Dr. Seymour is a Principal Lecturer at the School of Electrical and Electronic Engineering, Thames Polytechnic, London.)

Inventions, Patents and Trademarks

P. MEINHARDT. Gower Press, London, 1971. 24 cm × 16 cm. 397 pp. £6·50.*

CONTENTS: Part One: Invention and Inventions; Part Two: Patent Law and Practice; Part Three: Trade Mark Law and Practice; Part Four: Management of Patents and Trade Marks in Companies.

This book analyses the management problem of establishing patent policy, controlling patent and trademark finance and effectively organizing a patent department. Importance is placed on the need for patent protection in a competitive business, and it is stressed that companies ignore this at their peril. The book should help companies to avoid infringing other patents and at the same time enable the company to protect its own patents.

(Mr. Meinhardt is a member of the English Bar and an Associate of the Chartered Institute of Patent Agents.)

The Radio Amateur's V.H.F. Manual (3rd edition)

EDWARD P. TILTON and DOUGLAS A. BLAKESLEE. The American Radio Relay League, Newington, Connecticut, 1972. 24cm × 16·3cm. Paperback. 352 pp. $3.C0.

CONTENTS: Reception above 50 MHz. V.h.f. receivers, converters and preamplifiers. V.h.f. transmitter design. V.h.f. exciters and amplifiers. V.h.f. stations. Antennas and feed systems. Building and using u.h.f. antennas. F.m.—theory and techniques. F.m. transmitters, receivers and accessories. Repeaters —theory and practice. U.h.f. and microwaves. Test equipment for the v.h.f. station. Interference causes and cures.

The basic form of the previous edition of this popular book is retained but it has been completely revised and a large amount of new material is added to take account of the developments in the v.h.f. and u.h.f. region. These include single-sideband and solid-state converter projects for the amateur. The emphasis, as always, is on the practical engineering aspects.

(The authors are on the staff of the American amateur radio journal QST.)

An Introduction to Logical Design of Digital Circuits

C. M. REEVES. Cambridge University Press, London, 1972. 22·2 cm × 15 cm. 192 pp. Paperback. £1·60.

CONTENTS: Synthesis of a computer. Boolean algebra. The design of combinational circuits. Sequential networks. Computer circuits.

This book introduces a range of mathematical concepts involving discrete-valued variables and presents them in terms of their role in the description and design of digital circuits.

Computing systems are described both from the functional point of view and in terms of the subcircuits and individual electronic components from which they are built. An account of Boolean algebra introduces algebraic, tabular, and graphical procedures and their application to the design of combinational circuits. The extensions necessary to cater for time dependence are also developed. Exercises, many of which are capable of solution by means of a logic-tutor, are included and an appendix describes a computer simulation of a suitable logic-tutor.

(Dr. Reeves is Director of Graduate Students, Centre for Computer Studies, University of Leeds.)

Book Supply Service

As a service to members, the Institution can supply copies of most of the books reviewed in the *Journal* at list price, plus a uniform charge of 25p to cover postage and packing.

Orders for these books, which are denoted by an asterisk (*) after the price, should be sent to the Publications Department at Bedford Square and **must** be accompanied by the appropriate remittance.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 9th January and 1st February 1973 recommended to the Council the election and transfer of 76 candidates to Corporate Membership of the Institution and the election and transfer of 34 candidates to Graduateship and Associateship. In accordance with Bye-law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

Meeting : 9th January 1973 (Membership Approval List No. 153)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Direct Election to Fellow

LAWRENCE, Edward David, B.Sc. *Ulcombe, Maidstone, Kent.*

Transfer from Graduate to Member.

BROCK, David John. *Ash Vale, Surrey*
BROWN, John Edward. *Sawston, Cambridgeshire*
CLARIDGE, Peter Gordon, B.Sc., Sqn. Ldr., RAF *Locking, Somerset.*
COLEMAN, Raphael Frederick *Llandaff, Cardiff.*
DEEFHOLTS, Mervyn Gordon, *Ottershaw, Surrey.*
FOSTER, William Ronald *North Baddesley, Hampshire.*
FRY, Clifford Bertram *Cowes, Isle of Wight.*
GIBBS, Graham John *Chelmsford, Essex.*
GRADY, John, Inst. Lt., RN *Helensburgh, Dunbartonshire.*
HUDSON, Alan Sidney *Airdrie, Lanarkshire.*
HUSBAND, Anthony John, Flt. Lt., RAF. *Abingdon, Berkshire.*
IRWING, Edward Alexander *Newbury, Berkshire.*
JOSEPH, Michael, Sqn. Ldr., RAF. *Newton, Nottingham.*
JOWERS, James George *Kingsclere, Berkshire.*
KING, Noel Robert Bruce *Crewe, Cheshire.*
LEE, George Anthony, Flt. Lt., RAF. *Bushey Heath, Hertfordshire.*
LONGMORE, Keith Leslie *Coventry, Warwickshire*
MACLEAN, Duncan Clark *Glasgow*
MAJOR, William John, Lt., RN *Titchfield Common, Hampshire.*
MATIMONG, Frank Nigel *Chesham, Buckinghamshire.*
PUSEY, Ernest Albert *South Wonston, Hampshire.*
READ, John Harvey *Hitchin, Hertfordshire.*
SARGENT, Arthur John *Carshalton, Surrey.*
SCOTT, Terence Leslie, Lt., RN *Eggbuckland, Devon.*
TIDSWELL, David Alan *Wigginton, Hertfordshire.*
TREWIN, Ian Alfred Flt. Lt., RAF *Medmenham, Buckinghamshire.*
UTTERRIDGE, Edward Joseph *Newport, Isle of Wight.*

Direct Election to Member

HIX, Martin, B.Sc.(Eng,.) Lt. Cdr,, RN *Tavistock, Devon.*
JAMESON, Walter Stewart, *Malvern Link, Worcestershire.*
NEWSAM, William, *Ferndown, Dorset.*
VARNEY, Arthur George, *Bexley, Kent.*

NON CORPORATE MEMBERS

Transfer from Student to Graduate

ANDREWS, Ian Michael, B.Sc. *Cyncoed, Cardiff.*
BENNETT, Philip Alan *Cumbernauld, Glasgow.*

Direct Election to Graduate

LOGAN, John Howard, B.Sc. Flg. Off, RAF *Norwich,*
McGINN, Peter, B.Sc. *Perivale, Middlesex.*
MANNING, Timothy John *Maidenhead, Berkshire.*
MURPHY, John *Bletchley, Buckinghamshire.*
NEEDHAM, Richard James, B.Eng. *Henley on Thames, Oxfordshire.*
WALSH, Martin Lee, B.Sc. *Tredegar, Monmouthshire.*

Direct Election to Associate

HONEYBALL, Keith Arthur, *Marlow, Buckinghamshire.*
KEHOE, Owen John *Renmore, Galway.*
PASCOE, Derek Robert *Ilford, Essex.*

STUDENTS REGISTERED

GALANIS, Creon *London, N,14.*
JONES, Stephen *Thurmaston, Leicester.*
LANGRIDGE, Philip *Ramsgate, Kent.*
LYNCH, James Francis *London, NW3.*
NELLIGAN, Richard Peter *Bromley, Kent.*
SHARPE, James Lester *Narborough, Leicestershire.*

OVERSEAS

CORPORATE MEMBERS

Direct Election to Fellow

POTTER, William Francis, *Manotick, Ontario, Canada.*

Transfer from Graduate to Member

GEORGE, Kaniyarassaril Varghese, Flt. Lt., IAF *Adampur-Panjab, India.*
GILPIN, Humphrey Collin Jartue Milford *Freetown Sierra Leone.*
HORDATT, Clarence Fitzpatrick Alexander *St. James, Trinidad.*
OBO, Joab M. Alupakusadi *Kampala, Uganda.*
SAUZIER, Joseph Robert Denis, Flt. Lt., RAF *Akrotiri, Cyprus.*

Direct Election to Member

BAHIL, Vinod Kumar, Major, *Bangalore.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

BHAKAR, Amar Singh *Jamnager, Gujarat, India.*

Direct Election to Graduate

DAS GURTA, Shubhashish *Calcutta.*
PRASAD, Vadlamudi Sivarama *Hyderabad.*
RATAN, Bhakt *Sindri, Bihar.*
WONG, Wai Hung *Kowloon, Hong Kong.*

Transfer from Student to Associate

LEE, Yew Kwan *Singapore.*

STUDENTS REGISTERED

CHAN, Kwok Chee *Kowloon, Hong Kong.*
HAMEED, Husni Abdel Hassan *Khartoum.*
KWONG, Wan Kay *Singapore.*
KURUPPU, Don Kumarasiri *Bandaragama, Sri Lanka.*
LIM, Kee Choon *Kedah, Malaysia.*
LIU, Chi Kin *Shaukiwan, Hong Kong.*
SENEVIRATNE, Rohan Dushyand de Silva *Nugegoda, Sri Lanka.*

Meeting : 1st February 1973 (Membership Approval List No. 154)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

BANHAM, Keith Rudd *Walton on Thames, Surrey.*
BRAYFORD, Neil *Sheffield.*
BURNS, Robert Francis *Old Coulsdon, Surrey.*
COTTRELL, David Edward *Camberley, Surrey.*
DANE, Alan Arthur *Carlisle.*
DENNAY, Charles William *Evesham, Worcestershire.*
DEXTER, Clive Louis *North Harrow, Middlesex.*
HALL, David Anthony *Tunbridge Wells, Kent.*
HELSBY, Indira Patricia (Mrs.) *Ripley, Derby.*
HUNTER, Ian James *Stonehaven, Kincardineshire.*
LIGHTBOURNE, John Vernon *Walsall, Staffordshire.*
LUCAS, William Stanley *Potters Bar, Hertfordshire.*
LUDDEN, Gerard John Anthony *Moreton, Cheshire.*
McMASTER, John *Dalry, Ayrshire.*
MARSHALL, Peter Julian *Twyford Moors, Winchester, Hampshire.*
MORLEY, John David *Morecambe, Lancashire.*
MUCKLESTON, Alan John *Eridge, Kent.*
PAYNE, John Lincoln *Eynsham, Oxfordshire.*
PEBERDY, William Thomas *Thatcham, Berkshire.*
PERKINS, John Frederick *Fareham, Hampshire.*
POLLARD, Peter John *Polegate, Sussex.*
POTTINGER, Terence David *Solihull, Warwickshire.*
PROSSER, Averyn Graham *Newport, Monmouthshire.*
READING, Norman John *Dunstable, Bedfordshire.*
REYNOLDS, Charles Patrick Joseph *Helensburgh, Dunbartonshire.*
STRIKE, Roger James *Maidenhead, Berkshire.*
TAIT, Brian *Bedlington, Northumberland.*
WADE, Maurice James *West Wickham, Kent.*
WHITESIDE, Leslie Terence *Seven Kings, Essex.*
WINTERTON, Ian Thomas, Flt. Lt., RAF. *Chivenor, Devon.*

Direct Election to Member

CONGALTON, William, B.Sc. *London, SW18.*
FIRTH, Eric Walter, B.Sc., Ph.D. *Yeovil, Somerset.*
HOWE, Kenneth *Ripley, Surrey.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

MORRIS, Stephen John, B.Sc. (Tech.) *Pontypridd, Glamorgan.*

Direct Election to Graduate

HILLYER, Neville, B.Sc. *Reading, Berkshire.*
LEWIS, William Peter *Stockwood, Bristol.*
MARSHALL, Victor Alan *Great Dunmow, Essex.*
PALMER, Geoffrey Ian *Sherborne, Dorset.*
PLANT, David Jervis *Middlewich, Cheshire.*
PRESCOTT, David John *Worcester Park, Surrey.*
TAYLOR, James Cedric *Heywood, Lancashire.*
WALL, Derek John *Sutton Coldfield, Warwickshire.*

Direct Election to Associate

BURKE, Edmund *Wigston Fields, Leicester.*
SMITH, John William *Sunderland, Co. Durham.*
TULIP, Lawrence Wilfred *Throckley, Newcastle upon Tyne.*

STUDENTS REGISTERED

ATHWAL, Gurdev Singh *London, E12.*
LANE, Malcolm *London, N7.*
MOGER, Colin Stanley *Rayleigh, Essex.*
MURRAY, David Stuart *Newcastle upon Tyne.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

BARNETT, Ian James *Geneva, Switzerland.*
CRABTREE, David, M.Sc. *Montreal, Canada.*
HOWARD, Trevor Neville *Salisbury, Rhodesia.*

Direct Election to Member

WATERS, George William, B.Sc., D.Phil. *Maple Shade, N.J., U.S.A.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

CHEUNG, Kwok For, M.Tech. *Kowloon, Hong Kong.*
ONG, Hong Eng *Kuala Lumpur, Malaysia.*

Direct Election to Graduate

CHAN, Man Woon *Kowloon, Hong Kong.*
JONG, Boen Hin *Randwick, N.S.W. Australia.*
PALTEKIAN, Khadjik Nazareth *Amman, Jordan.*

STUDENTS REGISTERED

CHOCK, Kek Ling *Kuala Lumpur, Malaysia.*
TAN, Kim Shah *Kuantan, Pahang, Malaysia.*
TEYO, Ming Kim *Gopeng, Perak, Malaysia.*
WONG, Kiong *Selangor, Malaysia.*

Notice is hereby given that the elections and transfers shown on Lists 150, 151 and 152 have now been confirmed by the Council.

Correction: List No. 151 published in the Jan/Feb 1973 issue should be dated 29th November 1972.

Forthcoming Conferences

Noise and Vibration Control

The Third Conference on Noise and Vibration Control, organized by the IERE, SEE, University of Bath, and UWIST, will be held at Traherne Hall, Cardiff (UWIST Hall of Residence) from 16th–18th April. The following papers will be given:

'Theory of Noise Control'—Dr. S. A. Petrusewicz (Bath University).

'Theory of Vibration Control'—J. Grosjean, (Bath University).

'Practical Aspects of Noise Control'—C. G. Gordon (Institute of Sound and Vibration Research, Southampton University).

'Practical Aspects of Vibration Control'—Dr. D. K. Longmore (Bath University).

'Criteria and Standards'—D. R. Hub (UWIST).

'Code of Practice for Reducing the Exposure of Employed Persons to Noise'—Miss L. A. Pittom (HM Deputy Chief Inspector of Factories).

'The Role of the Local Authorities in Noise Control'—W. Bate (Association of Public Health Inspectors).

'Legal Aspects of Noise Control'—C. Kerse (University of Manchester).

'Control of Neighbourhood Noise'—M. J. Cockerell (Atkins Research & Development Ltd).

'Traffic Noise and its Control'—Dr. P. T. Lewis (UWIST).

'Internal Noise in Buildings'—D. Walters (University of Aston).

'Vibration and Services Noise in Buildings'—D. J. Croome (Loughborough University).

Fees are £25 for the whole conference, £10 for attendance either Monday or Wednesday, and £12 for Tuesday which includes the Conference Dinner and Exhibition. These charges include meals and refreshments but not accommodation.

Applications should be made to the Conference Secretary, (Mrs. J. Jackson), UWIST, King Edward VII Avenue, Cardiff, CF1 3NU. (Tel. Cardiff (0222) 42522, ext. 211).

The 1973 London Electronic Component Show

The London Electronic Component Show will be held from 22nd to 25th May, 1973, at Olympia, London. This biennial show, one of the longest-running of its type (the first was held as long ago as 1943—in war time and with restricted admittance), will be the 23rd of the series. Representing both the passive and active sides of the industry, it features instruments and equipment used in manufacturing and testing.

The previous show in May 1971 included a record number of overseas participants, with 235 firms from abroad among the total of 611 exhibitors. Because of industrial reorganizations, among other factors, the final total number of exhibitors in 1973 is likely to be rather less but, in compensation, an increasingly large number of exhibitors, both UK and European, will be taking part for the first time. Foreign attendance in 1971 was stated to be the biggest ever, representing over 13% of the 57,279 visitors, and the organizers are hoping to improve on these figures and on the stimulation of trade—the more so since it will be the first electronics exhibition since British entry to the EEC.

Members of the IERE may obtain complimentary tickets for the Exhibition on application to the Meetings' Secretary, IERE, 8-9 Bedford Square, London WC1B 3RG (Telephone 01-637 2771, extension 20).

The London Electronic Component Show is sponsored by the Radio and Electronic Component Manufacturers' Federation and is organized by Industrial Exhibitions Limited, a wholly-owned subsidiary of Industrial and Trade Fairs Holdings Limited. Further information regarding participation or attendance may be obtained from Industrial Exhibitions, Commonwealth House, New Oxford Street, London WC1A 1PB.

For the first time the IERE is associated with the Show through joint sponsorship, with the RECMF, of a Seminar which will be held at a nearby hotel. This Seminar will occupy three morning sessions, on 23rd, 24th and 25th May and its theme will be a survey of European components. The three sessions will be on 'European Components Research', 'Component Testing and Evaluation' and 'The Evolution of Harmonized Components in Europe'. Details of the papers and speakers will be published in the April issue of *The Radio and Electronic Engineer* together with registration information.

Colloquium on Remote Control System Organization

A one-day Colloquium will be held in London in October 1973 on 'Remote Control System Organization'. Systems of remote control and telemetering of necessity incorporate data transmission but the philosophy of system design to make efficient use of the communication path and of data transmission technique has not received due attention despite many years of successful practice in the industrial field.

The Colloquium is to be primarily concerned with the philosophy of system design and the choice of a particular technique to meet a system need. Relevant practical applications both with and without computer control will be discussed. It will be of interest to control engineers and all involved in the design, development, and operation of remote control systems, both in existing areas, and in other industries where the technique of telecontrol may become applicable.

The detailed scope of the Colloquium is as follows:

Design Choice made in Practical Schemes. The quantification of requirements for the implementation of a given system. The assessment of available communication paths, and the choice of established or new telecontrol techniques to meet specific operational needs.

Security of Command. The degree of security necessary in a command which may be inherent in the command signal itself or in revertive checks. The balance that must be struck between failure to act and false actions.

Scanning Patterns and Data Sequences. Sequential scanning or random 'on demand' access is an important system choice. Schemes with a high telemeter content may demand a simple scanning pattern, whereas schemes with a high alarm and command content require complex scanning patterns. Address/reply methods have similar balances to be struck.

Signalling Speed. The characteristics of block length, element speed and so on, effect subtly the effective correct information transmission rate. The basic error rate increases with signalling speed and hence a law of diminishing returns may apply.

Coding. The economic justification of different error detection and error correction methods and its dependence on different classes of transmission interference and system requirement.

Offers of papers for presentation at this Colloquium are invited and in the first instance a synopsis of 100–200 words should be sent to Colloquium Secretariat, IERE, 8–9 Bedford Square, London WC1B 3RG.

Forthcoming Institution Meetings

Tuesday, 27th March

JOINT IERE-IEE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP

Colloquium on
ARRHYTHMIA RECOGNITION AND DETECTION

IERE Lecture Room, 2.30 p.m.

Advance Registration necessary. No charge to members of IERE and IEE; £1.50 to non-members.

'The Clinical Importance of Arrhythmia Detection'
By Dr. D. A. Chamberlain (*Royal Sussex County Hospital*)

'Development of a "Bedside" Arrhythmia Monitor'
By Dr. J. M. Neilson (*Royal Infirmary, Edinburgh*)

'Out-Patient Monitoring Techniques'
By Dr. F. Stott (*Clinical Research Centre*)

'Multi-Dimensional Analysis in Arrhythmia Detection'
By Professor E. McA. Sayers (*Imperial College of Science and Technology*)

'The Relevance of some Methods of Detection, Recognition and Monitoring of Cardiac Arrhythmias'
By Dr. J. A. Bushman (*Royal College of Surgeons*)

Wednesday, 28th March

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Colloquium on
SECONDARY RADAR IN MARITIME APPLICATIONS

IERE Lecture Room, 2.30 p.m.

Advance Registration necessary. No charge to members of IERE and IEE; £1.50 to non-members.

'A General Purpose Marine Secondary Radar System'
By A. Allen (*Cossor Electronics*)

'Secondary Surveillance Radar IFF Mk 10 (SIF) Systems in Naval Ships and Aircraft'
By G. Morris and M. Peachey (*Cossor Electronics*)

'Possible Future Developments of the Marine Radar Transponder'
By F. Marshall (*ASWE*)

Wednesday, 4th April

COMMUNICATIONS GROUP

Facsimile—A Review

By M. J. Bowden and J. Malster (*Rank Xerox*) IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

The paper reviews the growth and potential of facsimile as an alternative communication medium over voice grade public switched and leased telephone lines.

The problems confronting facsimile manufacturers in relation to the common carriers, standardization, and economics are discussed. The relative merits of current techniques are considered and future trends are outlined.

Tuesday, 10th April

AUTOMATION AND CONTROL SYSTEMS GROUP

Colloquium on
RECENT DEVELOPMENTS IN SYSTEMS PERFORMANCE MEASUREMENT

Botany Theatre, University College London, 10 a.m. to 5.30 p.m.

Advance Registration necessary

Principles of Correlation

'Noise Rejection Properties of Correlation Measurements of Dynamic Response in Control Engineering'
By Dr. J. D. Lamb and Dr. P. A. Payne (*UWIST*)

Time Domain Testing

'The Application of Pseudo-Random Binary Sequences to a Gun-Mounting System'
By Cdr. D. J. Kenner (*Formerly Royal Navy*)

'A Special Purpose Digital Computer for On-Line Determination of System Impulse Response'
By H. Banasiewicz, Dr. S. E. Williamson and Professor W. F. Lovering (*University of Surrey*)

Fault Diagnosis

'Fault Diagnosis—A Pragmatic Approach'
By R. D. Woodward and E. W. Carr (*Honeywell*)

'Fault Diagnosis using Time Domain Measurements'
By Professor D. R. Towill (*UWIST*)

Frequency Domain Techniques

'Practical Problems of Digital and Analogue Measurement of Frequency Response (including Homodyne Detector Technique)'
By R. Kennedy and S. Urbanski (*SE Laboratories (Engineering)*)

'The Application of a Commutated Filter to the Design of a Frequency Response Analyser'
By C. J. Paull and W. A. Evans (*University College of Swansea*)

'Two-Channel Frequency Response Analysis by Special Frequency Test Methods'
By Dr. A. M. Fuchs (*Bafco*)

Test Automation

Paper by R. Passmore (*RRE*)

'The Desirable Characteristics of a Frequency Response Analyser for use in a Practical Measurement Environment and in Automatic Test Equipment'
By A. J. Martin (*Solartron Electronics Group*)

'The Software-Hardware Trade-Off in Automatic Dynamic Response Testing'
By J. B. Izatt and A. J. Ley (*Solartron Electronics Group*)

Wednesday, 25th April

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Electronic Aids to Position Fixing

By D. J. Phipps (*Decca Survey*)

MEETING POSTPONED: NEW DATE TO BE ANNOUNCED.

Kent Section

Thursday, 5th April

Instrumentation in a Nuclear Power Station

By A. Bewick (*CEGB*)
Medway College of Technology, Chatham, 7 p.m.

The lecture will deal with the control and monitoring aspects, namely temperature scanning; health physics instrumentation; reactor safety equipment; reactor control; and aspects of the management of instrument maintenance and calibration services.

Wednesday, 11th April

Afternoon visit to Dungeness Nuclear Power Station. Applications to attend to Honorary Secretary, Kent Section.

East Anglian Section

Wednesday, 21st March

The Transistor: its History and Consequences

By E. Wadham (*Mullard*)
Lecture Theatre 2, Civic College, Ipswich, 6.30 p.m. (Tea 6 p.m.)

The various stages of transistor development over the past 25 years will be described with special emphasis on those aspects which have generated major advances. The contrast between the early germanium period which established the industry and the present situation of large scale integration involving computer-aided design techniques will be highlighted by the showing of two Mullard films representative of these periods.

Thames Valley Section

Thursday, 5th April

ANNUAL GENERAL MEETING at 7 p.m.

Followed by

Fault-Tolerant Computing Systems

By L. A. Crapnell (*Ferranti*)

The J. J. Thomson Physical Laboratory, University of Reading, Whiteknights Park, Reading, 7.30 p.m.

The lecture will review some of the techniques which have been applied in fault-tolerant computing systems, and will describe in more detail one such technique which is being developed at Ferranti. The advantages and disadvantages of the proposed technique will be discussed.

North Western Section

Wednesday, 28th March

JOINT MEETING WITH IEE

Satellite Communication Systems

By Lt. Cdr. B. E. Collins, RN
Lecture Theatre RG7, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.)

Thursday, 3rd May

ANNUAL GENERAL MEETING at 6.15 p.m.

Followed at 6.45 p.m. by

Facsimile—A Review

By M. J. Bowden and J. Malster (*Rank Xerox*)
Lecture Theatre RG7, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.)

North Eastern Section

Wednesday, 11th April

A Country-Wide Data Transmission Network

By W. A. Ellis (Post Office) at 6 p.m.

Followed by ANNUAL GENERAL MEETING
Main Lecture Theatre, Ellison Building, Newcastle upon Tyne Polytechnic, 6 p.m. (Tea in Staff Refectory 5.30 p.m.)

Merseyside Section

Wednesday, 11th April

The Polytechnic and the Professional Engineer

By H. Currie (*Liverpool Polytechnic*)

Department of Electrical Engineering and Electronics, University of Liverpool, 7 p.m. (Tea 6.30 p.m.)

Yorkshire Section

Thursday, 29th March

Radio Communication within the North Eastern Gas Board

By R. Grant (*NE Gas Board*)
NE Gas Board, New York Road, Leeds, 7.30 p.m.

Friday, 27th April

(PLEASE NOTE CHANGE OF DATE)

ANNUAL GENERAL MEETING at 7 p.m.

Followed by Ladies Evening and Buffet Supper. Leeds University.

East Midland Section

Tuesday, 20th March

Application of Digital Logic

By I. D. Brown (*Rolls Royce*) and S. L. Norman (*BPB Industries Instruments*)

Lecture Theatre A, Physics Block, Leicester University, 6.45 p.m. (Tea 6 p.m.)

Basic logic families are illustrated, with working models, culminating in transistor-transistor logic medium scale integration (TTL-MSI). This family is then used to produce a digital frequency meter having a 4½ digit readout and 20 MHz operating frequency.

Thursday, 5th April

ANNUAL GENERAL MEETING

Room 0.8, Hawthorn Building, Leicester Polytechnic, 7.30 p.m.

West Midland Section

Monday, 26th March

Sonar and Underwater Acoustic Communications

By Dr. V. G. Welsby (*University of Birmingham*)

MEB Offices, Summer Lane, Birmingham, 6 p.m. (Tea 5.30 p.m.)

A review will be presented of modern techniques based on the use of sound waves in the sea and in lakes, rivers, etc. Systems for diver communication and navigation are described. High resolution sonars, sometimes using focused acoustic arrays, have uses which range from the study of the behaviour of fishshoals to aiding police searches in muddy canals. Acoustic telemetry is used to control submersible vehicles and to channel collected information back to the surface. Acoustic waves are used to count migrating fish in rivers.

Wednesday, 11th April

ANNUAL GENERAL MEETING at 7.15 p.m.

Followed by

Thyristor Gadgets for Home Entertainment

By R. G. Dancy (*International Rectifiers*)
The Polytechnic, Wolverhampton

South Midland Section

Tuesday, 17th April

New Radio Receiver Development

By Professor W. Gosling (*University College of Swansea*)

To be followed by

ANNUAL GENERAL MEETING

GCHQ, Oakley, Cheltenham, 7 p.m.

South Western Section

Monday, 7th May

ANNUAL GENERAL MEETING

Royal Hotel, Bristol, 7 p.m.

Southern Section

Wednesday, 21st March

JOINT MEETING WITH IEE

Application of Control to Artificial Limbs

By Professor J. M. Nightingale (*University of Southampton*)

Lanchester Theatre, University of Southampton, 6.30 p.m. (Tea in Senior Common Room, 5.45 p.m.)

Thursday, 29th March

ANNUAL GENERAL MEETING at 6.30 p.m.

Followed by

Aspects of Stereo Broadcasting

By J. H. Brookes (*BBC*)

Farnborough Technical College, 7 p.m.

Stereophony is an illusion, and, at the point of programme origination, something of an art. Artistically it is limited by the technology, and by the need to produce signals acceptable to listeners to monophonic, as well as stereophonic, reproduction equipment. How we hear and interpret sound, and how stereophonic material is originated in studio, is the main subject of this talk and demonstrations, though brief reference is also made to problems of distribution and transmission.

Wednesday, 4th April

Southampton Road Traffic Control System

By K. Newton (*GEC-Elliott Traffic Automation*) and J. Porter (*Southampton Corporation*)
Lanchester Theatre, University of Southampton, 6.30 p.m. (Tea in Senior Common Room from 5.45 p.m.)

Saturday, 12th May

Afternoon visit to Atomic Energy Establishment, Winfrith, 2–5 p.m.
Lecture and Tour. Apply for tickets to Honorary Secretary, Southern Section.

South Wales Section

Monday, 19th March

JOINT MEETING WITH IEE

Modern Measurement Techniques in Control Engineering

By Dr. J. D. Lamb and Dr. P. A. Payne (*UWIST*)
UWIST, Cardiff, 6.30 p.m. (Tea 5.30–6.30 p.m.)

Various methods of dynamic measurement widely used will be reviewed and the bases of their mechanization unified through an analysis of the cross-correlation process. Methods for predicting the noise rejection properties will be discussed, and an important difference between the serial and parallel modes of time domain measurement will be presented. The noise rejection capabilities of a range of available equipments will be demonstrated.

Thursday, 12th April

JOINT MEETING WITH IEE

Some Recent Research on Electrical Contacts

By Dr. A. Fairweather (*PO Research Station Dollis Hill*)
Department of Applied Science, University College, Swansea, 6.30 p.m.

Northern Ireland Section

Monday, 13th March

(PLEASE NOTE CHANGE OF DATE FROM 27th MARCH)

The Eric Megaw Memorial Award

Four student papers will be presented.
Ashby Institute, Stranmillis Road, Belfast, 6.30 p.m.

Tuesday, 10th April

Outside Broadcasting

By B. J. Slamin (*BBC, Northern Ireland*)
Ballymena Technical College, 7.30 p.m.