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The Engineer's Pay, Progress and Prospects

PROGRESS made by the British engineering profession over the past eight years can be assessed from over 20 pages of tables and graphs in the latest survey prepared by the CEI.* The wealth of information this contains will surely be of interest and value in many quarters—not least to engineers themselves who have generally co-operated in sufficient numbers—27 000—to enable the statisticians and their computers to produce reliable figures in the commendably brief period of just over three months. Well over a thousand Corporate Members and Graduates of the IERE had completed and returned questionnaires by the end of September when the data reduction and analysis operation started.

Natural curiosity and self-interest directs attention first to the data on incomes which are recorded for the year up to April 1973. Basic salary has continued to increase, but when the figures are adjusted to a common basis—using the Retail Price Index—it is noted that over the five years from 1966 to 1971 there was only a very gradual improvement in the value of lower incomes and, in fact, a slight decline in higher incomes. On the same basis the 1973 figures indicate that over the past two years the value of incomes has risen faster than the cost of living. Over the seven years 1966 to 1973 the overall increase in salaries is hardly excessive in the context of the general income pattern in the country, the median salary, for all engineers, irrespective of age, having risen in that period from £1949 to £3300 (by £600 since 1971).

Arising from analysis of the incomes of non-graduates, first-degree graduates and higher-degree graduates, it is clear that all graduates have an income advantage over non-graduates, increasing appreciably with age after about 30 and to a lesser extent there is a further income advantage in possessing a higher degree. These differentials persist to retirement even though there is some narrowing of the gap after about 50. Happily, overall unemployment at 0·52% has halved over the past two years and is down to 0·4% in the electronic apparatus manufacture area—the lowest level, shared with vehicle manufacture. Of all fields of work, more engineers from construction are unemployed (0·8%) and those from education have been least affected (0·2%).

The age distribution pattern continues to reflect a steady recruitment into the profession and there has been a further increase in the percentage—to 78%—of engineers falling into the 25 to 50 age bracket. The percentage of engineers under 30 has fallen, however, and the under-30 intake shows a very high percentage of University graduates which may be expected to continue as the new entry requirements to the profession take full effect. Already nearly 50% of all engineers now have a university or CNAA first degree.

Data on fields of work and employers of engineers show that while more engineers are in the manufacturing industries, there are small falls (8·0% to 6·8%) for 1971 to 1973 for electronic apparatus manufacture, and in broadcasting and telecommunications (2·6% to 2·2%) but a small rise in the Armed Forces (2·4% to 2·8%). General technical administration, research and development, and design are the types of work in which more engineers are now employed, though these figures have fallen slightly, counterbalanced by small increases in construction and installation, consultancy and, especially, non-engineering occupations. Predictably, the average ages of engineers in the general

* 'The 1973 Survey of Professional Engineers', obtainable from the Council of Engineering Institutions, 2 Little Smith Street, London SW1P 3DL, price £4·50.

technical administration, consultancy, commercial and teaching areas are the highest at 43. The mean incomes in these areas are also the highest, teaching being £3900, and the others £4000 to £4500. Self-employed engineers, who may be principals or partners in firms have an average age of 46 and mean incomes of about £5700.

In every field there are now more engineers in management positions than in 1971 although the fact that in several areas there was a fall from 1968 to 1971 suggests that there may have been a change in the way that engineers themselves interpret their responsibilities. Only in the fields of instrumentation and control, R&D, design, and teaching are less than 50% of engineers in managerial positions and this is probably in accord with the statistic that 47% of engineers under 40 are in such positions compared with 74% of those aged over 40. Business studies or management courses were attended full time by nearly 16% of all engineers and 9% part-time; technical updating or retraining is mainly full time and undertaken mostly by younger engineers, while foreign languages are nearly always studied part-time and by older engineers.

This year's survey includes a subjectively assessed section on job satisfaction, in which the questions covered related to satisfaction with salary, responsibility, work experience, training and career prospects. It is not easy to draw hard and fast conclusions from this analysis but it seems that engineers appear to be satisfied with their work experience but members of the profession are not generally so happy with their career prospects after reaching the age of 35–40 years.

As a result of the four Surveys so far conducted, the engineering profession is beginning to know something about itself, to establish yardsticks for comparing its different disciplines one with another. It would be interesting to be able to study comparable facts about other professions because just as there are common factors within engineering so too should there be common factors throughout the professions. Data such as are provided in this latest survey enable all individuals and career advisers to assess in what direction the individual is best able to determine both job satisfaction and possible renumeration. Future surveys on the engineering profession might well include comparison with the rewards to be obtained from other fields of employment. This would not necessarily show engineering in an unfavourable light; job satisfaction is not necessarily associated with renumeration. The majority of engineers are general practitioners but who would be content with the general medical practitioners lot on constant call or the repetitive work of many accountants or solicitors?

F.W.S.

PREVIOUS SURVEYS OF THE PROFESSION

'The Survey of Professional Engineers 1966', Ministry of Technology and the Council of Engineering Institutions (HMSO, London 1967).

'The Survey of Professional Engineers 1968', Ministry of Technology and the Council of Engineering Institutions, (HMSO, London 1970). (See *The Radio and Electronic Engineer*, **40,** p.217, November 1970, and **41,** p.561, May 1971).

'The 1971 Survey of Professional Engineers', The Council of Engineering Institutions (CEI, London 1971). (See *The Radio and Electronic Engineer*, **42,** p.59, January 1972).

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The computation of the best windward and running courses for sailing yachts

J. ELLIOT*

SUMMARY

The best course for a yachtsman to steer, for any course to windward, can be shown to be that which resolves the maximum speed in the direction of the true wind for any course lying between the reciprocal sailing vectors producing such a maximum. Outside these vectors it is nearly always quicker to steer a direct course, the exception being on the running points of sailing in light wind conditions.

Unfortunately measurements of true wind bearing and speed are not possible on a moving yacht and measurements of only the apparent wind parameters, namely, apparent wind velocity and apparent wind angle β less the leeway angle λ, are possible. The other practical velocity measurement is the yacht's speed through the water. The parameters available for the computation of the best course are therefore two velocities and one inaccurate angle—the inaccuracy being due to leeway.

Previously postulated solutions to the problem have confessed lack of feasibility due to the leeway error and due to the cost of the computation, for instance circuits employing 105 transistors consuming 5 W of power.[1]

This paper explains measurement methods which are free of errors due to leeway and describes an electronic computer which will enable the best sailing vector to be found and indicated on both the close-hauled and running points of sailing. The additional power required to compute the best course from the input of established instruments measuring apparent wind velocity and sailed velocity is 0·12 W.

* *EMI Electronics Ltd., Wells, Somerset.*

1 Introduction

1.1 *The Problem in Theory*

It must be recognized by most laymen that a sailing yacht cannot be expected to sail directly into the eye of the wind and the necessity to tack a course for destinations to windward is likewise generally appreciated. In the event, a yacht will begin sailing some 30 degrees off the wind—making a certain amount of fuss about it. As the yacht is brought further off the wind, its speed through the water will increase until the wind is broadside-on. As the yacht is turned further to bring the wind over the quarter and finally over the stern, the speed through the water will will tend to fall off—appreciably in light airs—but less significantly in stronger winds. The performance of a yacht may be readily appreciated pictorially by plotting the yacht's speed in knots against direction, thus producing a diagram in polar co-ordinates. (See Fig. 1.)

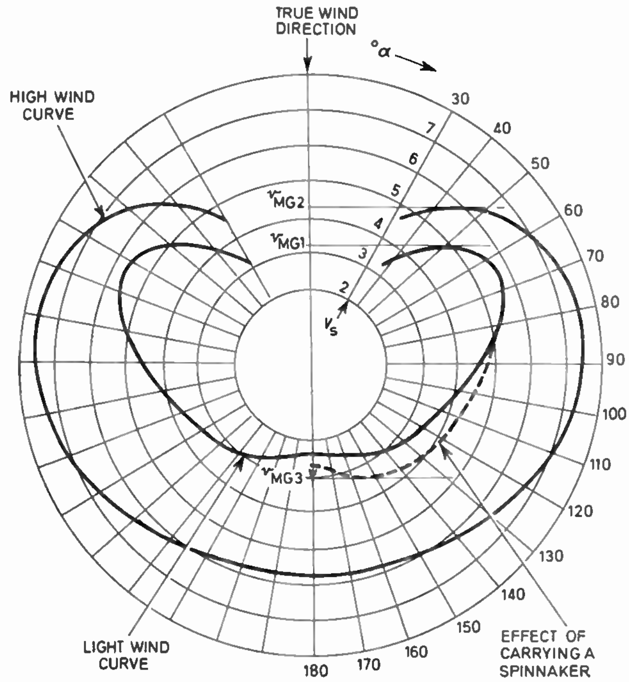

Fig. 1. Polar diagram of sailing yacht performance.

If in the example the desired destination lies within the arc $\pm 30°$ from the wind direction it will be necessary to tack the required course but the question arises as to which? Is it better to hug the 30° vector at just over 3 knots in a light wind or will it pay off to sail at 40° where we can do 4 knots? or even at 5 knots at 50°? Do we sail at the same wind angle on each tack or is there a combination of tacking angles which will minimize the time taken to make a desired destination?

A straightforward mathematical exercise cannot readily be undertaken due to the unwieldy terms with which one must cope. A retreat to the use of a general-purpose computer suggested an iterative program of the examination of elapsed time for trial combinations of courses using an equation which approximates a practical windward section of the polar diagram. The exercise confirmed that there was only one pair of tacking angles (one each side of the wind) for the minimum elapsed time to reach any destination lying within the angles. The actual angle is that at which a resolution of the yacht's velocity along the wind vector produces a maximum. That is at 42° in Fig. 1 for the light wind conditions and nearer 45° in the heavier weather. The resolved velocity along the wind vector is, conventionally, the velocity made good and the two velocities have been labelled v_{MG1} and v_{MG2} respectively for the two conditions just mentioned.

A similar situation arises on the running point of sailing (wind aft) where an inflexion will be seen in the polar curve especially when carrying a spinnaker. It pays off therefore sometimes to tack downwind, v_{MG3} in Fig. 1.

Superficially the problem would appear to be simply the determination of the maximum value of v_{MG} which may be had from

$$v_{MG} = v_S \cos \alpha$$

where α is the angle between the wind vector and the yacht's track and v_S is the sailed velocity.

The yacht's velocity v_S is readily and accurately measured by an electrical ship's log but the wind direction cannot be measured due to the forward motion of the yacht.

Having now stated the problem in its simplest terms and having no immediate practical solution, it will be necessary to familiarize the reader with the sailing parameters and the conventional signs allocated thereto. These are illustrated and listed in Fig. 2.

Of these parameters only the yacht's speed along its track (v_S), the apparent wind velocity (v_A) and the yacht's impression (γ) of the apparent wind angle (β) are capable of accurate measurement.

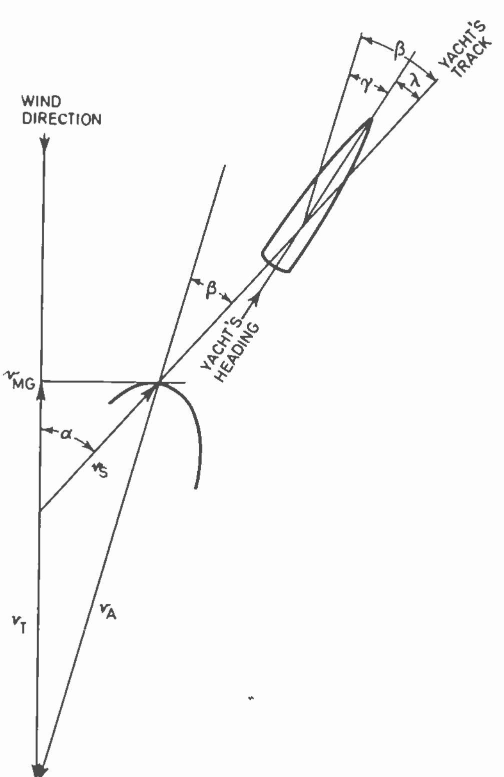

Fig. 2. The sailing parameters.

To establish a case for an electronic computation of v_{MG} it will be instructive to consider the alternative.

1.2 *The Problem in Practice*

The alternative to an available computer is to construct a family of polar curves for all sailable weather conditions. This involves tabulating the values of the measurable parameters of: the yacht's velocity, the apparent wind velocity and the yacht's heading at frequent intervals of azimuth. The leeway angle should also be taken and, if available, the apparent wind angle should also be recorded to correlate the results.

It will not always be practical to ask the helmsman to hold the yacht to a prescribed heading but rather slowly to bring the yacht round on or off the wind and to record values while the instruments are steady—since the wind will not be entirely constant in direction and strength in the long-term period. When dealing with results, only those measurements which will indicate a common true-wind strength can be used to produce each polar diagram. To obtain sufficient scatter points, through which a fair curve can be drawn, will require many runs covering quite a few miles and may well take half a day. During this time there will not only be short-term changes but also long-term weather changes may well occur.[2] The exercise must be repeated on occasions of differing wind strengths until a library of curves is accumulated for future reference.

Clearly the exercise is tedious and the nett result is a reference to which prevalent conditions may be related. In addition to this there is a hazard of the results being open to individual interpretation and be fraught with potential and real errors, e.g. heeling and leeway.[3]

To quote an informed opinion, McKinnon[4] has said, 'Yachtsmen as a whole have learnt the hard way a great deal which could have been predicted on the basis of such polar curves, had they been available, and if they can be measured or computed from other performance measurements, the methods evolved by yachtsmen are capable of further refinement'.

Proposals for such a refinement are the essence of this paper.

2 Practical Solutions

2.1 *Application and Ergonomics*

The alternative to the compilation of data as described above is to compute v_{MG} from the electrical analogues of the measurable parameters which are instant samples of the yacht's behaviour under the twin influences of the helmsman and the environment.

For this purpose the relevant formulae are developed in Appendix 1 from which the following three solutions are offered:

$$v_{MG} = \frac{v_A^2 - v_S^2}{2v_T} - \frac{v_T}{2} \qquad (1)$$

$$v_{MG} = \frac{v_A v_S \cos \beta - v_S^2}{v_T} \qquad (2)$$

$$v_{MG} = \frac{v_A v_S \cos \beta - v_S^2}{(v_A^2 + v_S^2 - 2v_A v_S \cos \beta)^{\frac{1}{2}}} \qquad (3)$$

The last two solutions involve an analogue of the apparent wind angle β but Fig. 2 shows that $\beta = \gamma+\lambda$ and λ cannot be measured.

However, the first solution does not use the wind angle measurement hence the maximum value of v_{MG} is therefore clear of leeway errors. Having discovered this, it is implicit that an on-board comparison of v_{MG} in instruments using equations (1) and (2) above, would enable λ to be defined. The value of knowing λ will be appreciated from the fact that the navigational compass readings are also in error by λ. In other words both the compass and wind vane indicate measurements taken with respect to the yacht's fore and aft line and not with respect to the yacht's track. To verify this, resort to empirical solutions was made and to this end Appendix 2 is included to show that all the methods of computing $v_{MG(max)}$ are all free ot positional errors due to leeway after all—a conclusion not known to be previously recognized.

Equations (1) and (2) contain an unmeasurable parameter (v_T) which, when neglected, yields in each case a v_{MG} which is in error but since v_{MG} has no practical value v_T may be ignored and the maximum value of 'v_{MG}' sought—a qualitative in lieu of the quantitative solution.

Furthermore since the concept of v_{MG} is based upon a polar diagram, each drawn for specific values of v_T, it follows that v_{MG} is meaningful only as long as v_T is known to be steady. Therefore wherever v_T appears in the equation it may be ignored or even made use of, for example, ergonomic objections to a display of qualitative v_{MG} can be overcome by indicating v_{MG}/v_T as will be evident later in this paper.

The dependence of a polar diagram upon a specific value of v_T raises a question of computer viability when, as is too often evident, the wind velocity is not at all constant.

In practice the procedure for determining $v_{MG(max)}$ is to 'bracket' the expected bearing. When close hauled, the maximum value of v_{MG} is not too far removed from the angle α at which the yacht stalls. The initial yaw should therefore be away from the true wind direction, then returning steadily to overshoot the $v_{MG(max)}$ indication, finally settling down at the maximum value. The bracketing angle would therefore be between $\pm 3°$ to $5°$. Assuming the turning rate of a yacht to be around 5 degrees per second, it should take 3 to 4 seconds to find $v_{MG(max)}$. A further bracket should not be necessary until a change in the parameters is sensed. From the personal comfort point of view it will probably be undesirable to update the course more frequently than at 5-minute intervals during which time 0·5 nautical miles will have been covered at 6 knots. The ability to practice this procedure implies steady-state conditions, that is, when the reaction time of the yacht and its crew is very short by comparison with the natural changes in the winds' mood.

The plausibility of this argument may be assessed by fitting the situation into the spectrum of the wind. Such a spectrum is illustrated by Watts[5] in which he shows, at one end of the spectrum, stable winds taking from 3 to 24 hours for the wind velocity to increase from 15 to 25 knots. At the opposite end of the spectrum gusty conditions obtain, the gusts arriving in cells at fairly regular intervals as follows:

short cells lasting 1 to $2\frac{1}{2}$ minutes at $v_T = 40$ kn mean,

medium cells lasting 4 to 9 minutes at $v_T = 10$ kn mean.

Between these extremes long gusts occur, separated by intervals of $\frac{1}{2}$ to 2 hours at a v_T of 10 kn and $\frac{1}{2}$ to 1 hour at a v_T of 20 kn. Taken in the context of a 6 min updating interval, steady-state conditions prevail in all but the medium and short gust conditions.

Although the reaction time of the yacht and crew is still relatively short the behaviour of the v_{MG} computer should be examined for usefulness during the gust-lull sequence.

In the Northern Hemisphere a typical gusting wind is accompanied by a kick-back followed by a veer in direction. Thereafter the wind will back as its velocity steadily diminishes to approximate its original value. During this latter period it may well be possible to bracket the heading and seek $v_{MG(max)}$ but during the turbulent period, lasting but a few seconds, the v_{MG} indicator will behave in the following manner.

When the wind gusts, the anemometer and wind vane which have very little mass, will respond rapidly, inserting new values into the computer. Representative values are shown in Fig. 3 and the values at the various phases are given in Table 1.

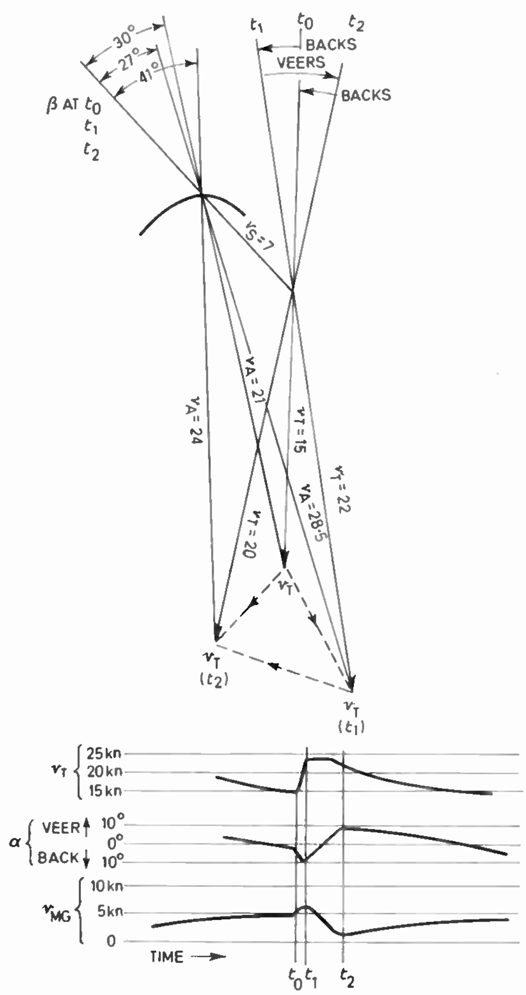

Fig. 3. The polar diagram in transient conditions.

Table 1

Phase	v_T	v_A	v_S	β	Indicated v_{MG}
t_0 (pre-gust state)	15	21	7	30°	4·9
t_1 (wind backing 10°)	22	28·5	7	27°	6
t_2 (wind veering 10°)	20	24	7	41°	3·55

Also shown in Table 1 are the time sequence for the true wind velocity, relative wind bearing and v_{MG}. As can be seen the yacht is assumed to have maintained its velocity, which is likely due to its inertia. With respect to the original true wind bearing, the v_{MG} has remained constant, but the v_{MG} meter will have indicated an oscillation—and this is a transient lasting for a few seconds.

In the example, the polar diagram has been drawn as if on a starboard tack (which is prudent under the circumstances since the initial backing of the wind will not cause the yacht to be caught in irons) and the subsequent veer will tend towards the reaching point of sailing. On the port tack the yacht may well be headed off by the major, veering, wind shift.

What of instrument-sailing under these conditions? Well, all is not lost, since the yacht, in high winds, will most likely be encountering winds in excess of the velocity required to produce the maximum speed for the yacht. Consider then, equation (11) in Appendix 1:

$$v_{MG} = \frac{v_A^2 - v_S^2}{2v_T} - \frac{v_T}{2}$$

from which it is clear that when v_A is considerably greater than v_S then the latter may be ignored. In other words v_S cannot substantially alter the point at which v_{MG} maximizes; then $v_{A(max)}$ and $v_{MG(max)}$ will have almost identical peaks. If any difference is perceptible, v_A will maximize slightly further off the wind thereby erring on the side of safety in the prevalent conditions. Thus the v_A-meter takes over the duties of a v_{MG}-meter in the heavier weather. Again, this is a fact not known to have been previously recognized.

It has been argued that v_{MG}-meters should not be used while racing on the grounds that their use removes some skill and puts those without to a disadvantage. However the foregoing has indicated that the helmsman will have to learn to use his computer and to discriminate between the indications of natural changes in the winds, and those brought about by his own actions. For example, under steady-state conditions, v_A, β and v_{MG} will respond only to alterations in the helm and under transient conditions the parameters will change on a nominally constant

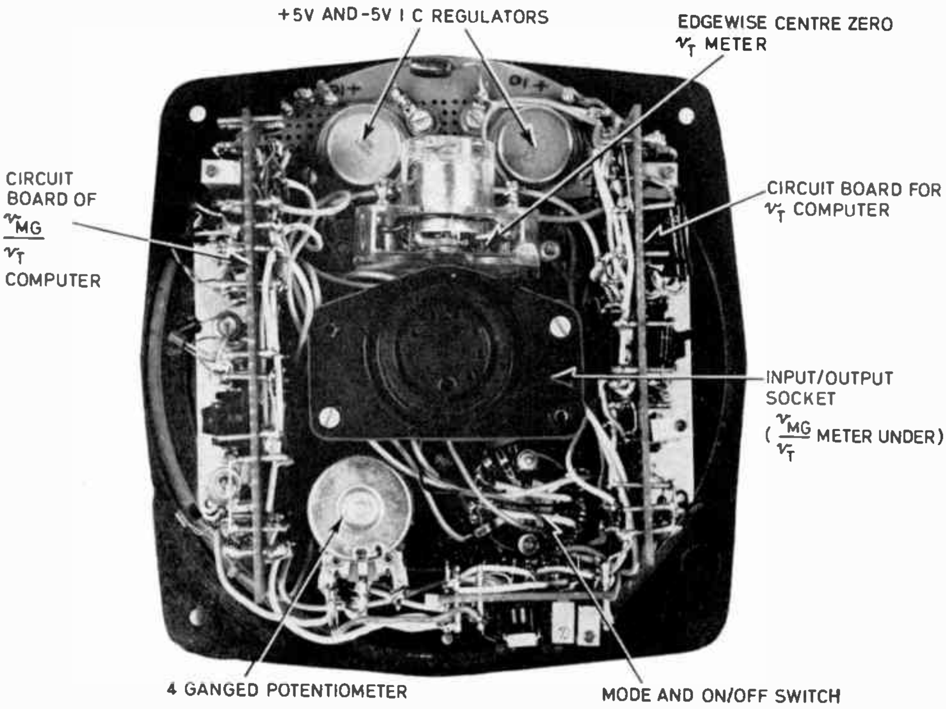

Fig. 4. (a) Front view of the v_{MG}/v_T computer.

(b) Rear view of the circuitry of the computer.

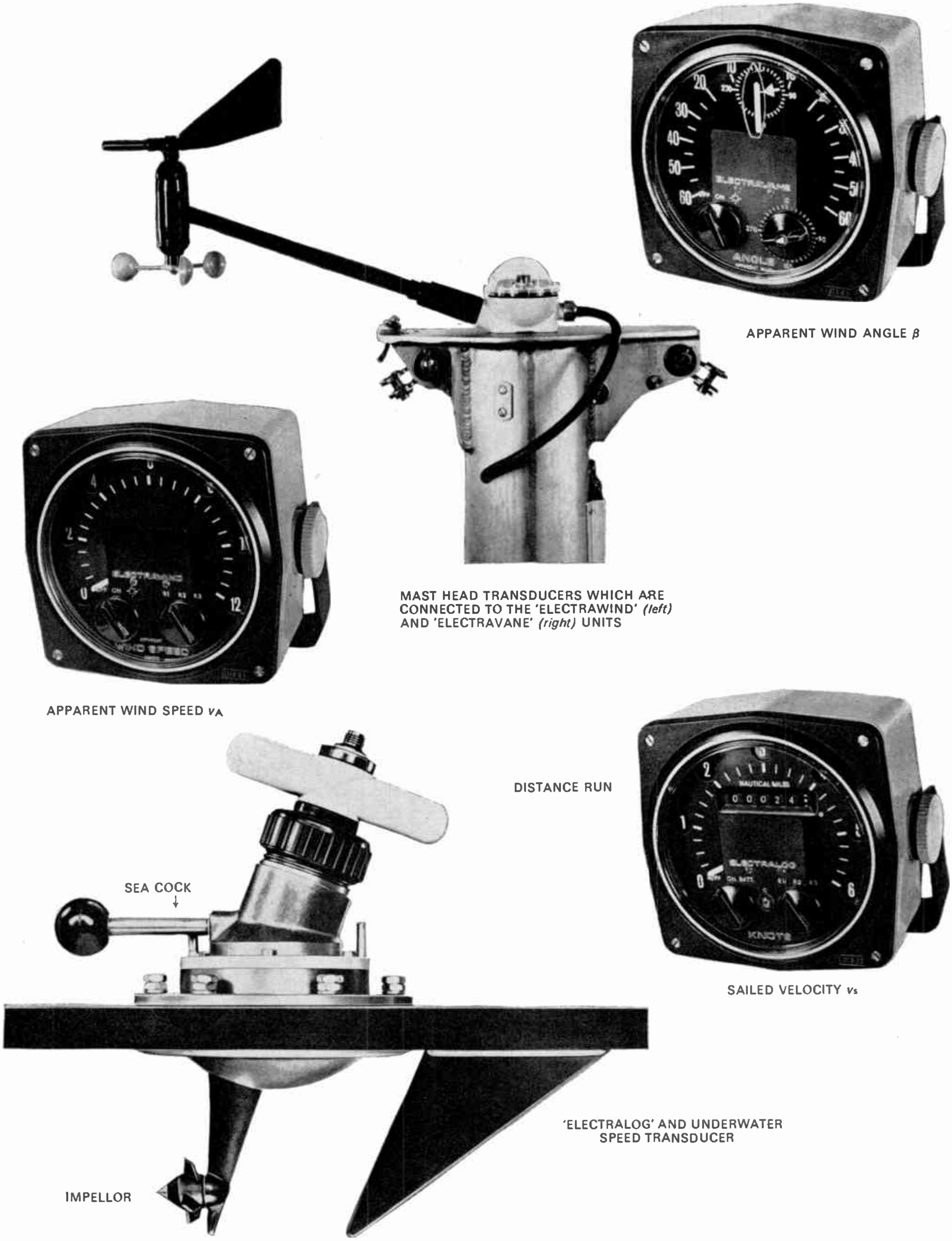

Fig. 4 (c) The input transducers associated with the v_{MG}/v_T computer.

Photographs by courtesy of EMI Marine.

course. It could be argued therefore that the introduction of a v_{MG} computer introduces new skills.

The validity of seeking $v_{MG(max)}$ can be ascertained by introducing a further indicator, namely that of v_T, for which all the necessary parameters for its computation are already available, as can be seen from the derivation of the formulae (equation (5)) covered in Appendix 1. The opportunity to seek v_{MG} will then be seen to exist, especially if the two indicators are combined as shown in Fig. 4 (a). To compute v_T reinvokes the angle β which we have striven to avoid to achieve greater accuracy. However if equation (8) is reconsidered, it may be manipulated from

$$v_{MG} = \frac{v_A^2 - v_S^2}{2v_T} - \frac{v_T}{2}$$

to

$$\frac{v_{MG}}{v_T} = \frac{v_A^2 - v_S^2}{2v_T^2} - \tfrac{1}{2}$$

Therefore,

$$\frac{2v_{MG}}{v_T} + 1 = \frac{v_A^2 - v_S^2}{v_T^2} \tag{4}$$

The factor of 2 and the constant on the l.h.s., is a matter of scaling and offset, enabling v_{MG}/v_T to be indicated. In practice v_S is hardly likely to exceed $v_T/2$ and v_{MG} will have a maximum of not more than $0{\cdot}9\, v_S$, i.e. $\alpha = \sim 25°$. A v_{MG}/v_T scale of 0 to 0·5 should, therefore, cover all possible performance in all wind velocities. The apparent wind angle β must be used to determine the true wind velocity v_T bit since this is to be a constant then the quotient can be determined by a hand-set potentiometer calibrated in terms of v_T. The variable part of the equation determining v_{MG}/v_T, that is $(v_A^2 - v_S^2)$, may then perform without the dubious aid of β.

2.2 Practical Hardware

2.2.1 *Options*

Appendix 1 shows the following three options by which v_{MG} may be found:

$$v_{MG} = \frac{v_A^2 - v_S^2}{2v_T} - \frac{v_T}{2} \tag{1}$$

$$v_{MG} = \frac{v_A v_S \cos\beta - v_S^2}{v_T} \tag{2}$$

$$v_{MG} = \frac{v_A v_S \cos\beta - v_S^2}{(v_A^2 + v_S^2 - 2v_A v_S \cos\beta)^{\frac{1}{2}}} \tag{3}$$

The computation of v_{MG} from (1) and (2) above is an identical exercise, that of the difference of two products. In the first, the products are v_S^2 and v_A^2 and in the second $v_S v_A \cos\beta$ and v_S^2.

The third option is more complex in its solution but contains no unmeasurable parameters albeit β is in error by the leeway angle.

2.2.2 *Transducers*

The design for the computation of v_{MG} is dominated by standard practice transducers which produce electrical analogues of the speeds and angle (Fig. 4 (c)).

(i) The ship's velocity, v_S can be measured to a good degree of accuracy by means of an impellor on the ship's bottom. The impellor produces a pulse or a sinusoid per revolution and is usually designed to produce pulses which may be counted to define distance and integrated to indicate speed. A convenient point will be readily available in the circuit of such instruments to provide suitable signals to the v_{MG} computer.

(ii) The apparent wind velocity, v_A, is measured with almost identical equipment to that used for the measurement of v_S, the impellor being replaced by an anemometer.

(iii) The apparent wind angle, β, is represented by an electrical signal produced from a wind vane at the masthead. The actual angle measured is, of course, γ, but as can be seen from Appendix 2 no error of $v_{MG(max)}$ ensues, in assuming this value for β. A convenient but relatively expensive apparent relative wind angle indicator can be provided by the well-known synchro method of angular position transmission. From such a three-wire system, which normally enables a linear presentation of angle, the cosine of the angle may be extracted by means of a Scott type of connexion and by appropriate initial orientation of the synchro.

2.3 Practical Circuits

Options (1) and (2) can be handled by almost identical circuits which have been combined and outlined in Fig. 5. The alternative modes of measurement are indicated by the switch position.

2.3.1 *Option 1*

For method 1 the switch is in position 1 when TR1 is fed with constant-area negative pulses from the apparent wind-speed indicator. TR1 is normally conducting and when turned off by the v_A pulses, C1 charges from $V+$ via R1 and R2. The steady state voltage on C1 depends on the pulse rate or wind speed. The v_A pulses also turn off TR2 which with R3 and R4 enables C2 to acquire a charge. The voltage across C2 is therefore proportional to v_A^2 and applied to the inverting input of operational amplifier IC1.

The v_S parameter is squared by a similar circuit comprising TR3, R6, R7 and TR4, R9 and C4. The voltage proportional to v_S^2 is applied to the non-inverting input of IC1. The difference signal will be indicated on the meter M1. To accommodate the wide variation in v_A the gain of the system is set by the tandem variable resistors RV1 and RV2 incrementally calibrated in terms of v_T.

2.3.2 *Option 2*

To compute by method 2, the switch would be in position 2 which invokes the apparent wind angle. Where the data are by means of a three-line synchro transmission system, the electrical analogue of the cosine of the angle may be obtained by means of a Scott type of connexion, the resistive network and diode bridge connected to the S_1, S_2 and S_3 windings of the synchro output in Fig. 5.

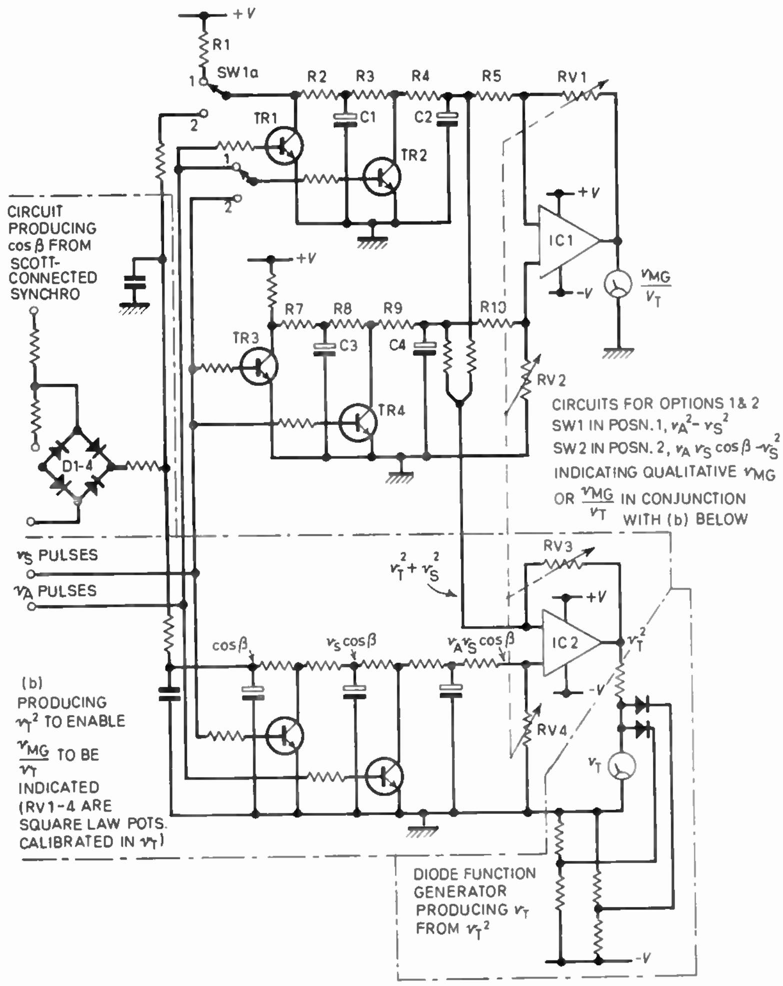

Fig. 5.
Circuit diagram for options 1 and 2.

The direct voltage produced by rectifying the synchro carrier, which will have an envelope proportional to the angle cosine, is now chopped by the v_A constant-area pulses which will have a repetition rate proportional to the apparent wind velocity.

After smoothing, a d.c. analogue of $v_A \cos\beta$ is then chopped by the constant area v_S pulses producing, after further smoothing, the $v_A\, v_S \cos\beta$ function. The difference between this and v^2 is performed by IC1 as before.

2.3.3 *Option 3*

To exercise equation (5) of Appendix 1 by means of a step-by-step analogue process would require no mean quantity of arithmetic circuits but there is an alternative worthy of consideration, and that is to use a diode generator of the required functions.

Schlang and Jacobs[6] have demonstrated the philosophy and design procedure for determining the transfer functions of two parameters. A function generator with three inputs is not as readily apparent and may be better understood by a three-dimensional representation, to which an attempt will be made to fit flat planes to curved surfaces. This may be compared to the more common practice of fitting straight to curved lines. Since we are equipped to think in only three planes the equation:

$$v_{MG} = \frac{v_A v_S \cos\beta - v_S^2}{(v_A^2 + v_S^2 - 2v_A v \;\cos\beta)^{\frac{1}{2}}}$$

which contains four terms, can be written

$$\frac{v_{MG}}{v_S} = \left(\frac{v_A \cos\beta}{v_S} - 1\right)\left(\frac{v_A^2}{v_S^2} - \frac{2v_A \cos\beta}{v_S} + 1\right)^{-\frac{1}{2}}$$

thereby reducing the terms to three in number.

Putting $\dfrac{v_{MG}}{v_S} = y, \qquad \dfrac{v_A \cos\beta}{v_S} = b$

and $\dfrac{v_A}{v_S} = x$

we get $y = (b-1)(x^2 - 2b + 1)^{-\frac{1}{2}}$

In practice y will have a maximum value of 1, which means that v_S can never exceed v_{MG}.

v_S will hardly ever exceed 10 knots and v_A 40 knots. v_S is also unlikely to be greater than $v_A/2$ and therefore:

$$a < b < 20 \text{ and } b < x < 20$$

Contours may now be produced for values of b from 1 to 20, a representation of which is partially depicted in Fig. 6 in which b/x and y/b are almost perfectly linear.

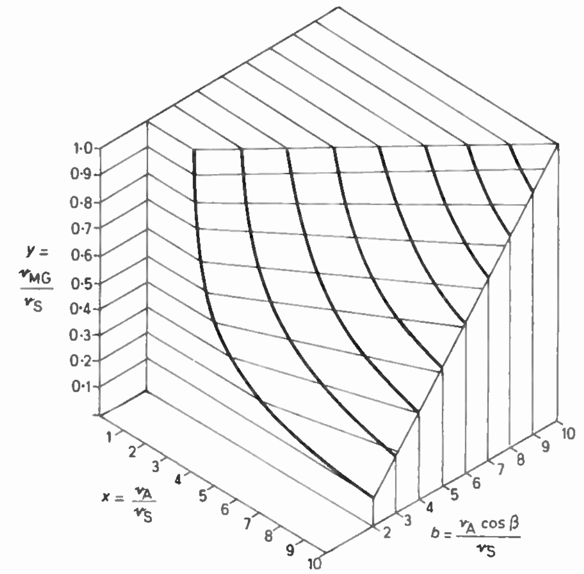

Fig. 6. A three-dimensional representation of

$$y = \frac{x\cos\beta - 1}{(x^2 - 2x\cos\beta + 1)^{\frac{1}{2}}}$$

The y-axis which is v_{MG}/v_S will be also recognized as $\cos\alpha$. From typical polar diagrams it will be seen that the maximum value of v_{MG} will occur within true wind angles of 35° to 50°, the corresponding cosines of which are 0·82 and 0·64.

There is no shortcut equation by which the planes and break points may be determined, the choice being purely intuitive, apart from defining the boundaries of greatest required accuracy, that is, $y = 0{\cdot}82$ and $y = 0{\cdot}64$ in the windward sector.

Now, within these boundaries, the slope of y to b is steeper at low values of x and b and flatter at the higher values: a feature which might be viewed with some misgiving.

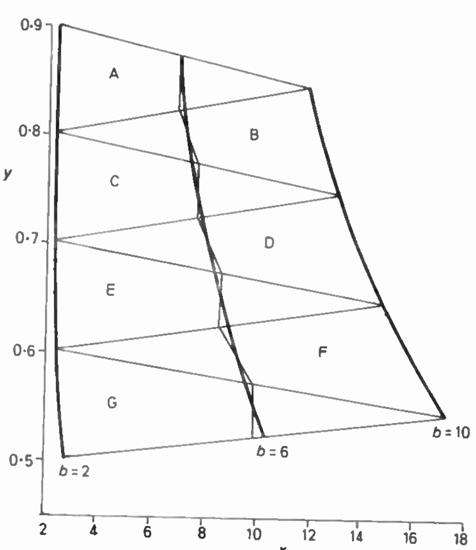

Fig. 7. Diode function generator. Division of playing area into seven planes (close-hauled region).

However, for a first approximation, assume that the area within the aforementioned limits is divided into seven planes, (as in Fig. 7) where plane A has three salient points with the coordinates of:

$$y = 0{\cdot}9 \qquad x = 2{\cdot}06 \qquad b = 2$$
$$y = 0{\cdot}8 \qquad x = 2{\cdot}14 \qquad b = 2$$
$$y = 0{\cdot}85 \qquad x = 11{\cdot}4 \qquad b = 10$$

To define the planes the coordinates must be extrapolated to the axes which is best done by solving the equations simultaneously producing

$$Qy + Px + Rb = 1$$

For plane A, $Q = 1{\cdot}76$, $P = 2{\cdot}2$ and $R = -2{\cdot}56$

i.e. $1{\cdot}76y + 2{\cdot}2x - 2{\cdot}56b = 1$

or

$$1{\cdot}76\frac{v_{MG}}{v_S} + 2{\cdot}2\frac{v_A}{v_S} - 2{\cdot}56\frac{v_A\cos\beta}{v_S} = 1$$

$$v_{MG} = \frac{v_S - 2{\cdot}2v_A + 2{\cdot}56v_A\cos\beta}{1{\cdot}76}$$

$$= 0{\cdot}57v_S - 1{\cdot}25v_A + 1{\cdot}45v_A\cos\beta$$

and other planes have been calculated in this fashion and appear in Table 2.

Table 2: Coordinates of the selected planes (close hauled region).

Plane	y	x	b	Coefficients y	x	b	v_A	$v_A\cos\beta$	v_S
A	0·8	2·14	2	1·76	2·2	−2·55	1·25	−1·45	0·57
	0·9	2·06	2						
	0·85	11·4	10						
B	0·85	11·4	10	1·29	0·09	−0·115	0·07	−0·09	0·77
	0·75	12·8	10						
	0·8	2·14	2						
C	0·7	2·4	2	3·48	3·48	−4·61	1	−1·21	0·216
	0·8	2·14	2						
	0·75	12·8	10						
D	0·65	14·5	10	1·5	0·09	−0·13	0·06	−0·086	0·66
	0·75	12·8	10						
	0·7	2·24	2						
E	0·6	2·4	2	5·08	3·17	−4·83	0·62	−0·95	0·2
	0·7	2·24	2						
	0·65	14·5	10						
F	0·55	16·9	10	1·78	0·07	−0·12	0·039	−0·07	0·56
	0·65	14·5	10						
	0·6	2·4	2						
G	0·5	2·64	2	9·8	4·08	−7·34	0·41	−0·75	0·25
	0·6	2·4	2						
	0·55	16·9	10						

Notice that although the planes were defined by v_{MG}/v_S, v_A/v_S and $(v_A\cos\beta)/v_S$, the required inputs to the DFG planes are v_S, v_A and $v_A\cos\beta$ of which only the latter requires computing before being offered to the DFG array.

A problem in which the sums of the parameters and their coefficients causes the plot to be exercised in plane A can be solved simply by making use of a circuit shown in Fig. 8(a); where the sum of the factored parameters

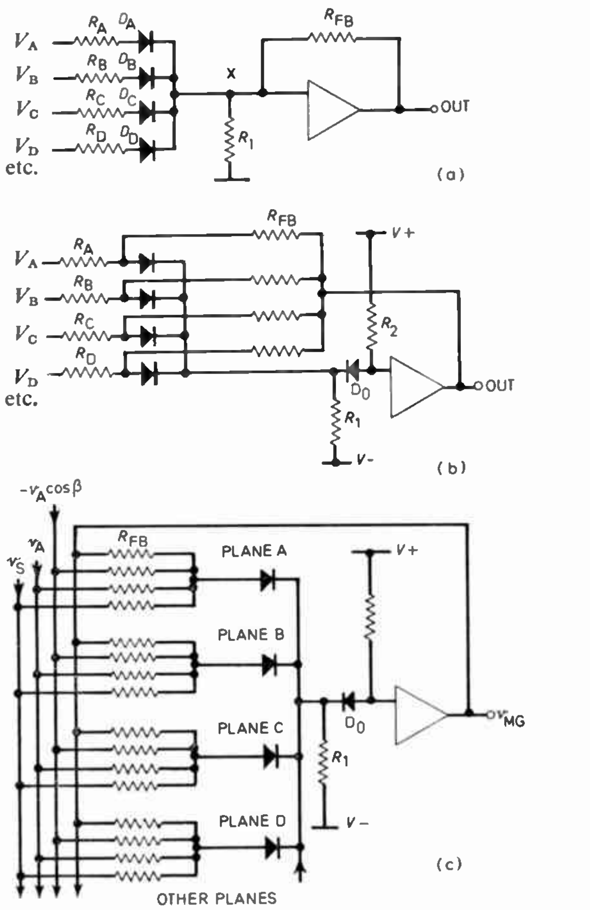

Fig. 8. Circuit of diode function generator to produce the analogue computation of

$$v_{MG} = \frac{v_A v_S \cos\beta - v_S^2}{(v_A^2 + v_S^2 - 2 v_A v_S \cos\beta)^{\frac{1}{2}}}$$

causes the plot to be properly housed in another plane then that plane must be switched in.

In the earlier example[6] of a simple shunt diode function generator, the selection of the relevant segment was made by the signal itself where its magnitude exceeded the breakpoint voltages of the lower segments.

However from Fig. 7 it will be seen that the corresponding break points are no longer constant throughout the length and breadth of the plane, i.e. the plane is tapered. Somehow the signals themselves must be made to invoke the correct plane and this may be done by causing a circuit to respond to that plane containing the maximum total value. For example if the parameters of the sailing example of Fig. 12 are taken where v_{MG} has a maximum we find that:

$$v_A = 18{\cdot}4,\ v_S = 8 \text{ and } v_A \cos\beta = 16{\cdot}744$$

which belongs in plane C. Then in this and the flanking planes we have:

$$\text{plane B } v_{MG} = 0{\cdot}07\, v_A + 0{\cdot}77\, v_S - 0{\cdot}09\, v_A \cos\beta = 5{\cdot}6$$

$$\text{C } v_{MG} = v_A + 0{\cdot}216\, v_S - 1{\cdot}21\, v_A \cos\beta = 6{\cdot}08$$

$$\text{D } v_{MG} = 0{\cdot}06\, v_A + 0{\cdot}66\, v_S - 0{\cdot}086\, v_A \cos\beta = 5{\cdot}2$$

An amplitude-selective circuit would therefore correctly take signals from plane C and give the answer to within 1%. With hindsight of course it could be said that if an adjacent plane should incorrectly establish supremacy then it has been incorrectly defined in the first place.

To abandon a philosophy of fixed breakpoint voltages in favour of a maximum-sensitive circuit raises two circuit design problems:

(i) how to make the circuit discriminate between the values offered for selection,

(ii) how to compensate for the diode biasing now that the breakpoint voltage has been lost.

For the purpose of argument let it be assumed that the voltages from the three input parameters are lumped together and fed through separate diodes to a summing junction, for example, in plane A.

$V_A = P_A v_A + Q_A v_S - R_A v_A \cos\beta$ which is fed through R_A and D_A etc. to R1. Then in Fig. 8(a) up to the point X there are a number of conductive branches which will put the complex series resistors all in parallel making the calculation of appropriate values an impossible and useless task. Similarly, it would be so if the network was offered to a feedback amplifier with the feedback resistor taken to point X.

If however separate feedback resistors are taken to the junctions of the lumped plane resistors and their corresponding diodes and if the gains R_{FB}/R_A, R_{FB}/R_B etc. are made identical, then the virtual earth points are all at the same potential as shown in Fig. 8(b).

If the input voltage to the R_A plane exceeds the input to each of the other branches then the virtual earth will be the pivot point between the output and the input to R_A. The relative pivot points of the output and the other inputs will be more negative, and assuming perfect diodes, they will be cut off. The circuit will respond to the dominant input and the demarcation between the planes should be clean. There is another philosophy whereby the resistor ratios are manipulated such that the demarcation point is smeared into a zone producing, in effect, a further plane but this is a refinement not indulged in this discussion.

The remaining problem of the diode offset voltage may be overcome by adding another diode D_0 and the resistors R1 and R2 which only add a few microamperes of bleed current to keep the diodes alive.

The basic circuit would finally appear as in Fig. 8(c). For the purpose of a practical illustration, the sailing example of Fig. 11 has been taken and the values calculated from the relevant plane coefficient and these are compared with the true values, i.e. $v_S \cos\alpha$, in Table 3.

The errors for the example do not exceed 2% in v_{MG}, but more importantly, the peak value occurs at the correct angle α. The worst errors in v_{MG} will occur for sailing parameters producing a track in the vicinity of the $b = 6$ contour (Fig. 7). Obviously if these errors are thought to be significant in α more planes can be taken.

As the circuit diagram implies, there are very few components—the tedium of design being rewarded by a simplicity of execution. The many and diverse resistors could be manufactured *en bloc* using the monitored blast etching process.

Table 3: Calculated values of sailing example of Fig. 11

v_A	v_S	β	$\cos\beta$		$\cos\alpha$ (y)	Operating plane	True v_{MG}	Approx v_{MG}	% v_{MG} error
17	6	20	0·94	30	0·866	A	5·196	5·3119	+2%
18	7·3	22	0·93	35	0·82	A	5·986	5·9017	−1·5%
18·4	8	24	0·91	40	0·766	C	6·128	6·0798	−1%
18·4	8·4	26·5	0·89	45	0·707	D	5·88	5·913	+0·5%
18·25	8·6	29·5	0·87	50	0·64	E	5·5	5·4	−1·8%
18	8·75	31·5	0·85	55	0·57	G	4·99	4·8584	−1%
17·8	9	34	0·83	60	0·5	G	4·5	4·5732	+1·5%

Solutions of other problems by mathematical model are possible such as true wind angle,[8] and the three-dimensional diagram of the v_{MG}/v_T nomogram also has possibilities. The v_{MG} model is to be preferred because of its direct application to the construction of polar diagrams.

3. Conclusions

Up to the time of writing the use of v_{MG} computers has been experimental, the major drawbacks being their size and power requirement. The foregoing philosophy has indicated lines of thought which have produced simple electronics making for economically practical and useable instruments which ought to find widespread appeal. As practiced use spreads, one could expect to see developments of design, use and philosophy. For example, the options were developed in the reverse order to this presentation. The option 1 solution has been readily developed into a combined indicator of v_{MG}/v_T and v_T as all the necessary circuits have already been produced, i.e. since

$$v_T^2 = v_A^2 + v_S^2 - 2v_A v_S \cos\beta,$$

the first two terms are used in the option 1 solution and $v_A v_S \cos\beta$ is used in option 2.

The investigation has shown that the leeway angle may be ignored and the errors due to β may be avoided. It has indicated that instrument sailing is possible in most weather conditions. It could be concluded that skill has not been eliminated especially when success in racing is also due to tactics and anticipation—particularly in respect of the wind shifts in relation to the tidal streams.

The use of the computer during working-up trials would be invaluable and it is the only means of confounding or confirming the predicted windward performance of a new yacht—a yardstick of design. The cost of the computer should be small bearing in mind that the inputs are from instruments which, most likely, have already been fitted as standard offshore equipment.

Even if the v_{MG}-meter continues to be banned by the Royal Ocean Racing Club while racing, the owner building specifically to the new racing rules (seemingly always) invariably takes delivery of his new yacht and, within a short space of time, races it. It is unlikely then that the owner will be prepared, or even have the time, to take the necessary range of polar diagrams before he will be expecting peak performance from his yacht costing many thousands of pounds sterling.

The final choice of instrument configuration would depend partly on its permissibility but all other things being equal, the option 1 solution is the author's preference due to its ability to perform equally well when running and to the slight additional cost of indicating v_T.

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5 Appendix 1: Derivation of Formulae

5.1 *Upwind Destination*

The parameters available to determine the equation to v_{MG} are two velocities and one inaccurate angle. The solution of the problem is, in essence, the solutions of the two triangles involved, namely,

(i) the triangle formed by the vectors v_{MG}, v_S and the angle α;

(ii) the triangle formed by the vectors v_A, v_S, v_T and the angle β.

In the first there are only two measurable parameters since α cannot be measured due to the forward motion of the yacht; but we cannot determine v_{MG} from only two parameters.

In the second there is also an unmeasurable parameter —the true wind velocity—again due to the yacht's velocity. The information in the triangle will, however, permit the calculation of the true wind velocity. Thus, putting $v_A = a$, $v_S = c$, $v_T = b$ and leaving β as it is, we have from Fig. 9, using the trigonometrical cosine rule

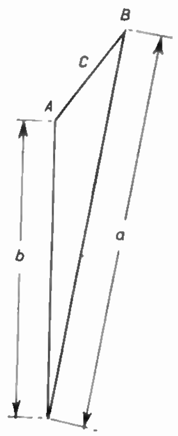

Fig. 9. Vector diagram of measured parameters.

i.e. $b^2 = c^2 + a^2 - 2ac \cos \beta$,

$$v_T^2 = v_S^2 + v_A^2 - 2v_A v_S \cos \beta \qquad (5)$$

This equation is useful because, if equipped with the relevant instruments, v_T can be found for each of the logged data acquired to produce polar diagrams. The exercise of this equation alone emphasizes the tedium of the process.

Clearly, to compute v_{MG}, both triangles must be considered jointly since neither is capable of the required answer alone.

The relevant pair of triangles are those shown by solid lines in Fig. 10, where once again the trigonometrical signs have been used. Therefore by putting

$$v_{MG} = b, \qquad v_S = c, \qquad v_T = c'$$

and leaving v_A and β as they are—since they will not be featured in the manipulation which follows—we have set out the known parameters.

To aid the solution a ghost triangle is invoked by subtending c to make b' and from this new vector, striking a perpendicular to meet the intersection of v_A and c'. There is now a third triangle of sides a', b' and c' containing a right angle at C′.

The first triangle is formed by the sides a, b and c with a right angle also at C. Since the angles at A are the same in each triangle and as each contains a right angle, then the triangles are similar and it can be said that

$$\frac{a}{a'} : \frac{b}{b'} : \frac{c}{c'}.$$

Now by substitution

$$\frac{v_{MG}}{b'} = \frac{v_S}{v_T}$$

$$v_{MG} = \frac{b' v_S}{v_T} \qquad (6)$$

A third right-angled triangle exists and is formed by the sides:

$(c+b')$, the base

a', the perpendicular,

v_A, the hypotenuse.

The angle β is contained within the sides $(c+b')$ and v_A. Then from the trigonometry

$$b' + c = v_A \cos \beta$$

since $c = v_S$ then

$$b' = v_A \cos \beta - v_S \qquad (7)$$

By substituting for b in equation (6) we get

$$v_{MG} = \frac{v_A v_S \cos \beta - v_S^2}{v_T} \qquad (8)$$

The velocity, v_T, has already been defined in equation (5)—it is the root of the right-hand side—and so

$$v_{MG} = \frac{v_A v_S \cos \beta - v_S^2}{(v_S^2 - v_A^2 - 2v_A v_S \cos \beta)^{\frac{1}{2}}} \qquad (9)$$

It would be desirable if the errors due to heeling and leeway could be avoided altogether. The source of each of these errors is the same, i.e. the apparent wind angle indicator. As already pointed out in Section 1.1, the equations were developed as if β and not γ was being measured, i.e. $\beta = \gamma + \lambda$.

At the moment there is no ready means of measuring the leeway angle and to correct γ to allow for the heeling error would require an electrical sensor of the angle of heel—a further complication involving a further trigonometrical equation.[3]

It can be shown however that the error due to the leeway angle does not significantly alter the angle α at which v_{MG} maximizes.

Having made the assumption that the true wind velocity may be regarded as a constant can we, in fact, simplify further and avoid β altogether?

Reverting to equation (1) we can, by transposition, rearrange

$$v_T^2 = v_S^2 + v_A^2 - 2v_A v_S \cos \beta$$

to

$$v_A v_S \cos \beta = \frac{v_S^2 + v_A^2 - v_T^2}{2} \qquad (10)$$

We can now dispose of β by substituting in equation (10)

$$\frac{v_S^2 + v_A^2 - v_T^2}{2}$$

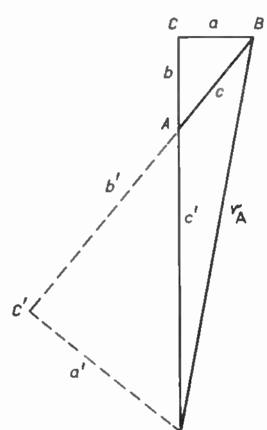

Fig. 10. Vector diagram of developed equations.

for $v_A v_S \cos\beta$ in equation (8)

$$v_{MG} = \frac{v_S^2 + v_A^2 - v_T^2}{2} - v_S^2/v_T$$

$$= \frac{v_A^2 - v_S^2 - v_T^2}{2} \Big/ v_T$$

$$= \frac{v_A^2 - v_S^2}{2v_T} - \frac{v_T}{2} \qquad (11)$$

Once again we have an equation in which we can consider the true wind velocity as being constant in the short-term period.

5.2 *Downwind Destination*

In the introduction emphasis was laid on the close-hauled case simply because the yachtsman has no option but to tack a course to his destination. It was also explained that the helmsman is faced with the enigma of decision regarding a slow direct course or a faster indirect course. A similar condition obtains in the vicinity of a dead run in light wind conditions where the polar curve may have an inflexion especially when carrying a spinnaker. Once again the course which results in a maximum of the yacht's velocity along the true wind vector will pay off, and once again we must resort to the arithmetic to see why this is so.

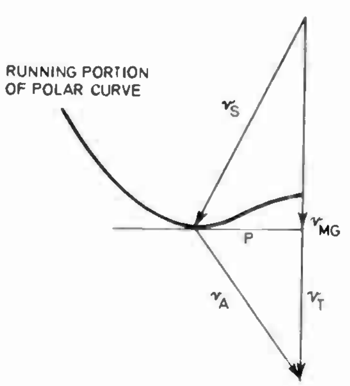

Fig. 11. Vector diagram of downwind case.

The portion of the polar curve when the wind is aft is shown in Fig. 11 where it will be seen that all the velocities are with the wind. There is now no difficulty of producing ghost triangles since all the necessary parameters aid a simple solution using only those of v_A and v_S as the following argument shows.

In the diagram the perpendicular P, making a tangent with the polar curve, forms a right-angled triangle with v_S and v_{MG} providing the other two sides.

The perpendicular also forms part of a second right-angled triangle with v_A and $(v_T - v_{MG})$. We have now two triangles which may be solved, using Pythagoras's theorem, when:

$$v_{MG}^2 = v_S^2 - P^2$$

and

$$(v_T - v_{MG})^2 = v_A^2 - P^2$$

$$(v_T - v_{MG})^2 - v_A^2 = v_{MG}^2 - v_S^2$$

$$v_T^2 - 2v_T v_{MG} + v_{MG}^2 - v_A^2 = v_{MG}^2 - v_S^2$$

$$-2v_T v_{MG} = v_A^2 - v_S^2 - v_T^2$$

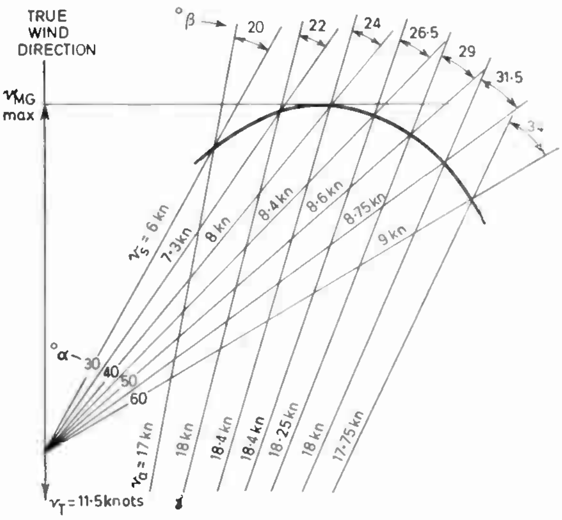

Fig. 12. Numerical example of polar curve.

hence

$$v_{MG} = \frac{v_S^2 - v_A^2}{2v_T} + \frac{v_T}{2} \qquad (12)$$

The reader will now recognize this as equation (11) but having the arithmetical signs reversed.

6 Appendix 2. Numerical Examples including the Effect of Leeway

It will be instructive to now consider a numerical example to illustrate the foregoing. This is shown in Fig. 12 from which it will be seen that the yacht starts sailing at an angle α of 30° at a v_S of 6 knots when the true wind speed v_T = 11·5 knots. The instruments on board would indicate an apparent wind angle, β (assuming no leeway) of 20° and an apparent wind speed, v_A of 17 knots. The yacht's speed is still increasing beyond α = 60° but notice that the maximum v_{MG} occurs at α = 40° when the yacht is doing 8 knots. The velocity made good to windward is over half a knot faster at 40° than it is at 50°—almost 10% better.

The relevant parameters of the polar diagram of Fig. 13 are shown in the table together with the v_{MG} values for

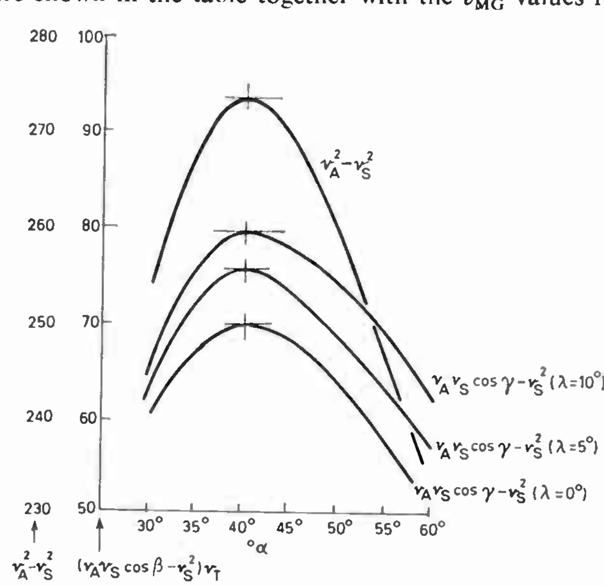

Fig. 13. Graphs showing Table 4 results.

Table 4

α	β	$\cos\beta$	γ	$\cos\gamma$ $\lambda = 10°$	γ	$\cos\gamma$ $\lambda = 5°$	v_A	v_B	$v_{MG} = v_A^2 - v_B^2$	$v_T v_{MG} = v_A v_B \cos\beta - v_B^2$	$v_{MG} = v_A v_B \cos\gamma - v_B^2$ $\lambda = 10°$	$v_{MG} = v_A v_B \cos\gamma - v_B^2$ $\lambda = 5°$
30	20	0·94	10	0·98	15	0·97	17	6	253	60	64	63
35	22	0·93	12	0·98	17	0·96	18	7·3	270	67	73	71
40	24	0·91	14	0·97	19	0·95	18·4	8	274	70	80	75
45	26·5	0·89	16·5	0·96	21·5	0·93	18·4	8·4	268	67	78	73
50	29·5	0·87	19·5	0·94	24·5	0·91	18·25	8·6	257	63	74	69
55	31·5	0·85	21·5	0·93	26·5	0·89	18	8·75	248	57	70	63
60	34	0·83	24	0·91	29	0·87	17·75	9	234	51	64	58

NOTE: (i) The scale of β is more compressed than the scale of α.

(ii) The maximum value of v_{MG} (equation 4) occurs as expected from the polar diagram at $\alpha = 40°$ and has a v_{MG} of 274.

$$\text{The true } v_{MG} = \frac{v_A^2 - v_B^2}{2v_T} - \frac{v_T}{2} \qquad \text{At } 40°\ v_{MG} = \frac{274}{2\times 11{\cdot}5} - \frac{11{\cdot}5}{2} = 6{\cdot}1 \text{ knots.}$$

(iii) Using equation (2), $v_{MG} = 70$ and the true $v_{MG} = \dfrac{v_A\, v_B \cos\beta - v_B^2}{v_T} = \dfrac{70}{11{\cdot}5} = 6{\cdot}1$ knots

(iv) Had we been able to determine α accurately, then true v_{MG} is also $v_B \cos\alpha = 8\times 0{\cdot}766 = 6{\cdot}128$.
The curves for the four tabulated values of v_{MG} are shown in Fig. 13.

equations (6) and (8) but neglecting the factors due to the true wind velocity. Since we ought to consider the ramification of the leeway angle, two columns of figures are also included for the angles of $\lambda = 5°$ and $\lambda = 10°$. Reference will be made to this later.

6.1 *The Effect of Leeway*

Throughout this paper, the arguments have omitted the leeway angle λ. Recognizing that an onboard measurement of β would be therefore in error, it had always been considered to be a restriction in the measurement of true v_{MG}. Since equation (11) determines v_{MG} without recourse to an angle measurement then it is

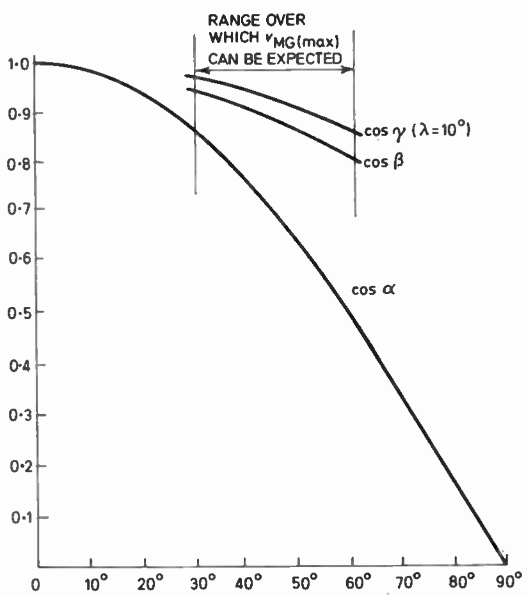

Fig. 14. Comparison of cos α, cos β and cos γ curves.

implicit that from γ, which is the yacht's impression of β, the leeway angle could be measured. The desirability of knowing the leeway will be recognized from the fact that both the magnetic compass and the windvane are using the ship's head as the reference. Allowances must be made, when reckoning the ship's position, for the leeway angle. If then we had both methods of determining v_{MG}, one using equation (2) and the other equation (8), then maximum values of v_{MG} would be expected at different bearings—the difference being the leeway angle. In practice the helmsman would be bracketing the v_{MG} maximum course using the equation (8) indicator and then an adjustment to the measured angle γ could be made by adding the analogue of the angle λ until both v_{MG} indicators rose and fell together. This is especially pertinent to windvane instruments using synchro transmission when a synchro differential transmitter may be interposed between the vane and the indicator. However examination of the curves for β, and γ with 5° and 10° added leeway all show peaks at the same angle α which has already proved to be the best course. This is a surprising result since, knowing from the outset that β would be in error by λ, one would have expected that not only would the true value of v_{MG} be in error but the angle at which the peak occurs would also be expected to be in error.

A clue as to the reason for the apparent absence of error may be gleaned from Fig. 14 where the relevant cosines of α and λ have been plotted for the two leeway angles considered. The curve for cos γ shows no significant change of slope when compared with that of cos β, the only effect being to elevate the v_{MG} curve as the leeway angle is increased.

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Before and immediately after the war, during which he served as a Radar Artificer in the REME, Mr. Elliot was with Denco (Claton) Ltd., a company which he helped to found in 1937; from 1946–1949 he was its Chief of Test. For the next four years he worked at the Admiralty Signal and Radar Establishment.

Mr. Elliot's study of computers for sailing yachts, on which the present paper is based, arises from a long standing interest in sailing and began as a private venture, subsequently supported by the company.

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The design of a precision video delay line

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SUMMARY

The paper reviews briefly the very stringent conditions which must be fulfilled by video delay lines connected in a television programme chain. The most suitable method of construction is found to be a cascade of delay sections of the same image impedance, and three types, one believed to be novel, are examined in detail with respect to their suitability for long delays inserted in main signal-handling paths. It is concluded that the advantage lies with low-pass sections combining pairs of complex conjugate m values in an appropriate configuration. These offer a markedly improved performance compared with conventional methods of construction, together with a substantial reduction in the number of inductors for a given delay over the video band. Guide lines for the design of video delay networks are laid down, and methods of compensation for component dissipation are discussed in some detail. The performance of a prototype 1 μs delay line built on these principles has fully confirmed the theoretical findings.

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1 Introduction

Video delay lines are employed in television practice for the equalization of the electrical lengths of signal paths in order to ensure that all video signals arrive at a common destination, usually a programme mixer, with precisely the correct time relationship. For example, in programme origination areas the individual picture sources, video tape recorders, cameras, telecine machines, and so on, may be located at considerable distances apart. The path differences at a mixer may then amount to as much as several microseconds.

One way of overcoming this difficulty is to adjust the timings of the synchronizing signals fed to the picture sources, and indeed this has the great advantage of not distorting the video signal. On the other hand, the need to route the output from a picture source over a number of different paths according to circumstances renders this practice operationally rather inconvenient, and signal delays are preferred provided the distortion they introduce can be made sufficiently small.

Until comparatively recently, video delays almost universally took the form of a suitable length of coaxial cable of the type for interconnexion purposes, since its group delay characteristic is remarkably flat and such cable is relatively inexpensive. On the other hand, its very low value of delay per unit length, slightly less than 5 ns/m, creates space problems which, with delays up to the 5 μs or so at present required by the BBC, become quite insoluble. Additionally, since the loss at the top of the video band for a 5 μs delay will be around 20 dB, fairly extensive equalization and amplification will be required. Nevertheless, for very short delays the use of flexible coaxial cable has the merits of cheapness, and ready availability of materials, combined with low distortion.

An improvement in the space factor is possible by the use of specialized delay cables, although not quite as much as one might suppose since a minimum radius of curvature must be observed when they are coiled, and since in the highest-delay types no electromagnetic screening is provided, turns must be adequately spaced to avoid crosstalk effects. Other practical difficulties are experienced, which will not be entered into here.

Within the last few years lumped constant delay lines in miniaturized, encapsulated form have appeared on the market, and have already replaced coaxial cable as a delay element in many areas in view of their excellent performance and small size. However, increasing difficulty is experienced in meeting the very strict specification for lines used in main signal path as the delay is raised above, say 1 μs. This is attributable on the one hand to the fact that it is not possible to relax performance requirements proportionally with an increase in delay, and on the other hand to the inevitable difficulties associated with the use of a large number of delay sections in tandem.

It was accordingly considered that an investigation into the possibility of employing alternative forms of delay section might prove profitable, with lines of around 5 μs in electrical length particularly in mind. This has revealed, as will be demonstrated below, that a reduction by a factor up to four in the number of sections required for a given delay is practicable, depending upon the particular design, with at the same time a significantly improved performance compared with units at present available.

2 Performance

2.1 *Tolerances*

When considering the allocation of performance tolerances one must take a global view of the entire television signal path which, as far as the responsibility of the broadcasting authority is concerned, extends from the picture source up to the transmitter. The figures used in practice by the BBC and the IBA are available,[1] but when one attempts to allocate a logical proportion of these to a single item of equipment such as a delay line, the resulting limits are found to be impracticably small. One is consequently forced to ask instead what the best is one may expect to achieve with such a delay line, bearing in mind always that if this optimum seems inadequate for the purpose, it will become necessary to seek some alternative solution of the problem. By the same token, one can only concede a relaxation of tolerances with longer delays to the extent that the degradation in performance resulting from the increased electrical length seems unavoidable. However unsatisfactory this may seem, it is after all only another instance of the universal engineering compromise between the theoretically desirable, the practically achievable, and the expense.

2.2 *Linear Distortions*

Fortunately, non-linear types of video signal distortion can be ruled out completely in this instance, and one is left with luminance–chrominance gain and delay inequalities and the sine-squared pulse K-rating. Detailed information on these and methods of measurement can be found elsewhere.[2,3] Briefly, the former are the differences between the gain (or loss) and the delay, at low video frequencies and in the sub-carrier region, of a colour video signal, whereas the latter is an assessment of the distortion experienced by a standardized pulse waveform in traversing the network, expressed in a manner which can be related to the subjective impairment of the corresponding picture.

The provisional tolerances allowed by the BBC at the moment for a 1 μs delay line are $\pm 0{\cdot}05$ dB gain inequality and ± 5 ns delay inequality, which are evidently approaching the limits of practicable measurement. For the K-rating the limit is $\frac{1}{2}\%$, which is again a low figure. Not surprisingly, these figures have hitherto proved difficult to meet with delay lines up to 1 or 2 μs in length, and their attainment with longer delays poses formidable problems with the methods of construction used hitherto.

2.3 *Return Loss*

The return loss of a network is defined as $20 \log_{10}|\rho|^{-1}$, where ρ is the reflexion coefficient of one end of the line against its nominal termination, the other end being terminated in the nominal impedance. There are accordingly two values for each delay line. Full information on methods of measurement is given in Refs. 2 and 3.

The extent to which return loss is important depends entirely upon circumstances. If the delay line is always to be used between resistive terminations it will not be necessary to measure return loss specifically, since any reflective effects will be taken into account implicitly in the overall measurement of performance. On the other hand, if delay lines are to be used to terminate other equipment, or if a long delay is to be made up from a number of shorter delays, then some limit must be placed on the return loss in order to ensure that the overall response will not be degraded by the consequent reflections to an intolerable extent.

In general, of course, a good return loss is a desirable feature since the utilization of the delay line need not then be restricted in any way, but unfortunately experience has proved that high values are difficult to achieve in a consistent manner in practice. As an example, the present BBC specification for a 1 μs delay line calls for a 29 dB return loss up to 100 kHz, and 24 dB thence up to 5·5 MHz (one end only), although this figure is largely dictated by what can be obtained. If it were possible, one would like to insist upon a minimum of 30 dB up to 5 MHz in each direction, after which the return loss could be allowed to deteriorate. These figures, of course, refer to 625-line colour television systems.

2.4 *Effects of Dissipation*

Ideally, the networks employed as delay sections should have zero insertion loss when correctly terminated, but in practice this is never the case, principally as a result of dissipation in the inductors. The capacitors also contribute to the dissipation loss, but almost invariably to a much smaller extent.

The reason why this loss is relatively important with delay lines is readily explained by Mayer's theorem,[4] which states that an approximate value for the increase in loss due to dissipation is given by

$$\Delta A \simeq 8{\cdot}7\, \omega D \frac{\mathrm{d}B}{\mathrm{d}\omega} \text{ decibels} \tag{1}$$

where D is the mean of the dissipation factors of the inductors and capacitors, assumed uniform throughout the network, and $\mathrm{d}B/\mathrm{d}\omega$ is the group delay at the angular frequency ω. Thus the dissipation loss is directly proportional to the magnitude of the group delay, which is naturally always relatively large in a delay network. The accuracy of this expression is surprisingly good for values of D less than, say, 0·02. Incidentally, eqn. (1) also demonstrates that if D is a constant, which is often roughly the case, then the insertion loss increases linearly with frequency.

In a typical example, assume that the Q-factors of the inductors in a 1 μs delay line are 100, and that the Q-factors of the capacitors are very high in comparison. Then the loss at the top of the video band is found to be about 1·5 dB. When it is recalled that the maximum allowable difference in insertion loss between low video frequencies and the sub-carrier region is $\pm 0{\cdot}05$ dB, it becomes evident that precise compensation is required. Indeed, the observance of the tolerance on gain inequality is entirely dependent upon the equalization, which must be precisely tailored to each individual delay line in all except very short lines. This point will be considered further below.

3 Delay Line Design

3.1 *Individual Networks*

One possibility which springs to mind when considering the design of delay networks is the use of a single linear-phase low-pass filter, designed to have the correct value of delay.

Such networks have certain virtues; in particular, since they can be designed to meet a specific requirement they can be very economical in terms of numbers of components. They may be predistorted if desired in order to compensate for dissipation effects,[5] and can also provide band limiting.[6,7] Finally, their synthesis presents fewer problems than formerly now that computers can take over the considerable amount of calculation involved.

On the other hand, they do not lend themselves particularly well to quantity production since each delay line is an individual filter with a variety of element values and resonant frequencies, which implies care in assembly and a relatively complex, and hence expensive, alignment procedure. Furthermore, problems arise where networks have to be cascaded without the use of intermediate buffer amplifiers. Although it is possible to design them to have the same image impedance, which permits interconnexion without mismatch, reflexion effects at the terminations may nevertheless modify the overall response quite significantly. If this difficulty can be circumvented, it is possible to envisage the construction of delay lines of any given length by the use of a series of such networks in a binary sequence of delay values.

3.2 *All-pass Networks*

Another superficially attractive procedure is the construction of a delay line from a cascade of constant-resistance all-pass delay networks, since this appears to offer a complete solution to the problem of obtaining good return losses. Information is available in the literature for the design of either maximally flat or equi-ripple delays.[8,9]

Unhappily, this expectation is not realized in practice. Even comparatively small amounts of dissipation in the components have the effect of spoiling to a significant degree the constant-resistance property of the networks, and the resulting return loss characteristic is likely to be no better than with suitably-designed low-pass filter sections. Procedures are known for compensating for dissipation[10] but their use could not be envisaged under production conditions.

Chains of all-pass networks also suffer from the disadvantage of requiring a large range of component values with a variety of resonant frequencies, which not only renders their manufacture uneconomical, but which also gives rise to a loss curve of a form difficult or even impossible to equalize when, as is normally the case, the dissipation is not uniform between the various resonant circuits. Each individual all-pass section has a delay/frequency curve with a single peak of delay, which by Mayer's theorem then gives rise to a loss curve with a single peak. If the dissipation is non-uniform, the resultant undulatory curve is very difficult to deal with. The degradation of the section impedances due to the dissipation also gives rise to internal reflexions.

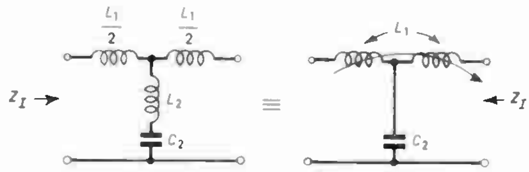
Yet a further drawback is the generation of anticipatory transients resulting from the abrupt drop in delay beyond the designed band of frequencies. Many of the waveforms generated in studio areas have a spectrum considerably exceeding the nominal video bandwidth, and if the signal energy is sufficiently great above, say, 6 MHz, the corresponding signal components will appear as 'rings' in advance of each transition. Any distortion of this kind is greatly disliked by operational staff. Such effects are minimized when low-pass configurations are employed first because the rate of fall of delay beyond the band is rarely as severe as with all-pass networks, and secondly because the irrelevant high frequency energy is in any case attenuated.

3.3 *m-derived Sections*

A much superior and more flexible procedure from the point of view of manufacture is the use of a cascade of identical low-pass filter sections with the same image impedance. The number of individual component values is reduced to a minimum and the alignment is much simplified, with a corresponding reduction in costs. The dissipation should be the same in each section, with the result that the overall insertion loss curve will be simple in form and easily equalized.

The earliest low-pass delay lines employed Zobel image parameter prototype sections,[11] whose group delay is flat for a very small fraction of the pass-band near zero frequency, but it was soon realized that m-derived sections with m greater than unity offer a higher delay per section over a greater fractional bandwidth. Such sections, which have an imaginary frequency of infinite loss, require one negative component, which can be realized conveniently in the form of the mutual inductance of a transformer with windings in series-aiding. A version of this which took the form of a continuous solenoidal winding tapped at regular intervals was devised by Lalande and Gloëss[12] in 1938. In various modified forms, principally aimed at reducing the deleterious effect of the higher-order mutual couplings, this line has been in use until comparatively recently. However, the somewhat inconvenient physical form, and the difficulty of minimizing manufacturing irregularities led finally to its replacement by a chain of discrete sections of the configuration shown in Fig. 1. All the commercially available lumped-constant delay lines with which the writer is acquainted utilize this method of construction.

The greatest difficulty from the manufacturing viewpoint lies in the transformer, which must be constructed with specified and accurate values of both self and mutual inductance. This is by no means impossible, but is nevertheless quite difficult when close tolerances have to be met. A classic device which renders the adjustments of the self and mutual inductances largely independent of one another is to wind the transformer for a coupling coefficient which is higher than is actually required. The



$L_1 = 2mL_0$; $-L_2 = \dfrac{m^2-1}{m}L_0$; $C_2 = mC_0$; $k = 1-m^2$.

$L_0 = \dfrac{R_0}{\omega_c}$; $C_0 = \dfrac{1}{\omega_c R_0}$

image impedance $\quad Z_I = R_0\sqrt{1-x^2}$

group delay $\quad T_g = T_0 \times \dfrac{1}{\sqrt{1-x^2}[1+(m^2-1)x^2]}$

where $T_0 = \dfrac{m}{\pi f_c}$ and $x = f/f_c$

Fig. 1. m-type delay sections.

correct mutual inductance is then obtained by inserting a suitable inductor in the common leg of the transformer, thus reducing the negative mutual inductance to the desired value. Since inductors are relatively costly, however, this expedient is not often employed in commercial units.

3.3.1 Dimensioning of m-derived sections

The group delay of an m-derived low-pass filter section with image terminations is given by

$$T_g = \frac{dB}{d\omega} = \frac{1}{\pi f_c} \cdot \frac{m}{\sqrt{1-x^2}[1+(m^2-1)x^2]},$$

which is more conveniently written as

$$\frac{T_g}{T_0} = \frac{1}{\sqrt{1-x^2}[1+(m^2-1)x^2]} \tag{2}$$

where $T_0 = m/\pi f_c$ is the zero-frequency delay, and $x = f/f_c$. T_0, as was shown by Moss,[12] corresponds to the shift in time of the centroid of a pulse transmitted through the network, and hence may be taken as defining the nominal delay.

It is usually assumed that the optimum dimensioning of delay sections is obtained when the parameters are chosen so as to yield a maximally flat curve of T_g/T_0, achieved by equating the coefficients of like powers of the frequency variable in the numerator and denominator. This is not possible in principle in equation (2), but a close approximation can be found for m by expanding the term under the square root sign in series form, giving

$$\text{denominator} = 1+(m^2-3/2)x^2+(3/8-m^2)x^4+\ldots \tag{3}$$

Assuming the 4th and higher powers of x can be neglected, the maximally flat curve is evidently given by

$$m^2 = 1{\cdot}5. \tag{4}$$

Unfortunately, the order of this curve is so low that the range of over which T_g/T_0 is substantially flat is severely limited, as is borne out in Fig. 2(a), which leads to a rather large number of sections per microsecond for a video delay line. An alternative is to recognize that the prime requirement is for the delays at zero frequency and in the region of sub-carrier frequency to be equal, whereas a small deviation in between these two frequencies is of lesser importance.

One accordingly equates T_g/T_0 to unity at $x = x_1$, where x_1 is the normalized frequency judged to correspond to colour sub-carrier frequency. The value of m is then obtained immediately. The magnitude of x_1 will always be made as large as is consistent with a tolerable error at lower frequencies. A suitable value for high-grade video lines might be, for example, $x_1 = 0{\cdot}3$, the curve for which is shown in Fig. 2(b).

3.3.2 Return loss

The image impedance of the m-derived low-pass section is

$$Z_I = R_0\sqrt{1-x^2} \tag{5}$$

so that a good return loss can only be expected at the lower values of x. However, in the present instance the utilization of the pass-band is low enough for this not to be found restrictive, for example, at $x = 0{\cdot}3$ the worst return loss is still slightly over 30 dB, assuming ideal components. With more efficient types of delay section which employ a greater fraction of the pass-band and have a prototype image impedance a good return loss requires the addition at each end of the line of a matching half-section with $m = 0{\cdot}684$, which will provide an almost perfect termination over the utilizable bandwidth. In principle, the same result could be achieved by designing the sections with a more complex image impedance, but the number of components required is thereby increased to an uneconomical extent.

The ability of the $m > 1$ section to dispense with matching end sections offers the valuable facility of being able to construct any required value of delay, within the delay of a single section, simply by adding the correct number of sections in series. No account need then be taken of the non-linear delay characteristics of the terminating sections, which otherwise might require some change in the line parameters. This flexibility is of considerable importance in some practical applications.

3.4 *The M, r Delay Section*

The drawback of the m-derived section lies in its poor utilization of the pass-band, resulting in a rather large number of sections when a long delay is required. This can be attributed to the fact that only one variable parameter, that is the value of m, is available for shaping

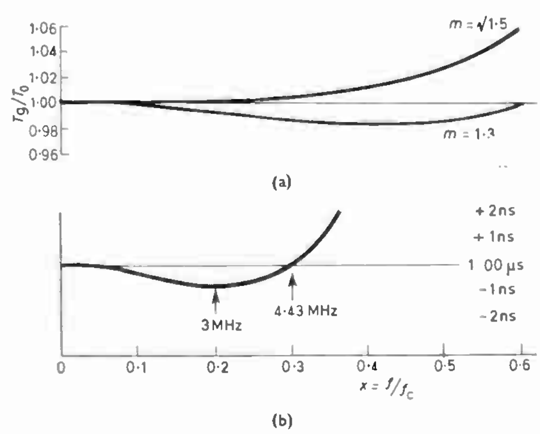


Fig. 2. The group delay of m-derived delay sections.

the curve. As a result, one can only effect what is virtually a tilt of the curve with respect to the $T_g/T_0 = 1$ line, as is clear from Fig. 2(a).

An improvement in pass-band utilization of about 2 : 1 can be achieved at the expense of one additional capacitor per section by means of the 4th-order network shown in Fig. 3. This has two independently variable parameters, M, which evidently plays a role rather similar to the m of the m-derived section, and r, which controls the magnitude of the coefficient of coupling of the transformer.

The configuration of this network superficially resembles that of a 4th-order all-pass section, but its use as an image-parameter type delay section is believed to be novel. Individual sections can, of course, be cascaded at will provided they are designed for the same impedance, and the loss is ideally zero over the pass-band when an image termination is used. This latter is appreciably flatter than the mid-series impedance of the $m > 1$ section, as can be seen from Fig. 4 where the relative impedance is plotted for the typical value $M^2r = 0{\cdot}3$.

An important practical advantage is conferred by the presence of the capacitor C_1 across the windings of the transformer L_1 (Fig. 3). This latter inevitably possesses stray capacitance across the windings which modifies the behaviour of the section at the higher frequencies, although its effect is mitigated by the low fractional bandwidth employed. The presence of C_1 across the transformer windings in the M,r section enables this stray capacitance to be absorbed.

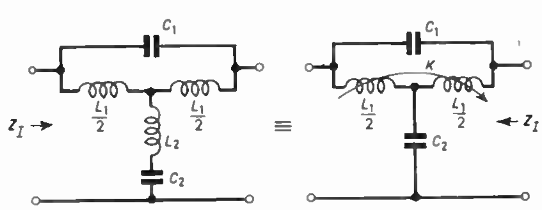
3.4.1 Dimensioning of the *M, r* section

The group delay for convenience may be written as

$$\frac{T_g}{T_0} = \frac{1-Bx^4}{[1+(A-B-1)x^2+Bx^4]\sqrt{1-(B+1)x^2+Bx^4}} \qquad (6)$$

where $A = M^2$ and $B = M^2r$. $T_0 = M/\pi f_c$.

If the procedure of equating T_g/T_0 to unity at a normalized frequency x_1 is again adopted, an infinite



$$L_1 = 2ML_0;\ L_2 = -\frac{1-r}{4}L_1 = -\frac{(1-r)M}{2}L_0;$$

$$C_2 = 2MC_0;\ k = \frac{1-r}{1+r}\ \ L_0 = \frac{R_0}{\omega_c};\ C_0 = \frac{1}{\omega_0 R_0}$$

$$Z_I = R_0\sqrt{\frac{1-M^2rx^2}{1-x^2}}$$

$$\text{group delay}\quad T_g = T_0 \times \frac{1-M^2rx^2}{[1+(M^2-M^2r-1)x^2+M^2rx^4] \times \sqrt{1-(1+M^2r)x^2+M^2rx^4}}$$

where $T_0 = \dfrac{M}{\pi f_c}$ and $x = f/f_c$.

Fig. 3. The M,r type of delay section.

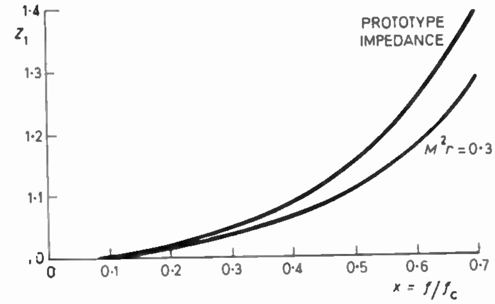


Fig. 4. Variation of impedance with frequency of M,r section.

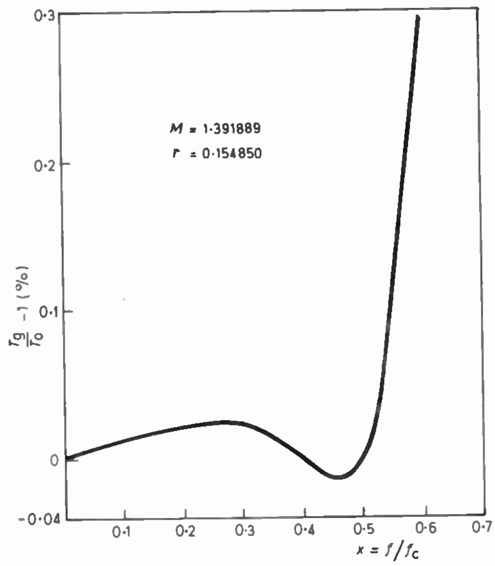


Fig. 5. Variation of delay with normalized frequency for M,r section with image terminations.

number of pairs of values of A and B is available for each value of x_1, so that some initial exploratory work becomes necessary. However, by means of a simple optimization program it is readily found that the flattest curve of T_g/T_0 against x is always given with B equal to, or close to, 0·30, and that x_1 may be made 0·5 or even greater, according to the amount of deviation from flatness which is allowed. The value of A corresponding to $x_1 = 0{\cdot}5$, which seems suitable for the present purpose, is then found to be 1·93736, that is $M = 1{\cdot}39189$, $r = 0{\cdot}15485$. The resulting curve of T_g/T_0 against x for image terminations is shown in Fig. 5; it has a maximum error of only 0·04% up to $x = 0{\cdot}5$, and increases by only 0·29% at $x = 0{\cdot}6$, which confirms the choice of parameter values. It is evident that for many purposes an even higher value of x_1 would be acceptable.

This ability to utilize a very high proportion of the pass-band is not an unmixed blessing, since appreciably more severe demands are made upon the terminal matching of a chain of such sections if the extremely good potential performance is to be realized in practice, as has already been pointed out. An inspection of the image impedance of sections with $M^2r = 0{\cdot}30$ reveals that an image-parameter prototype impedance such as can be

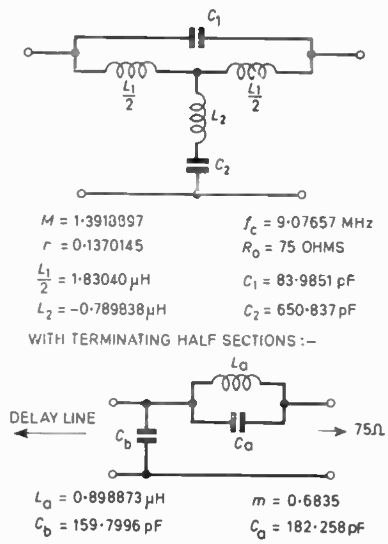


Fig. 6. Modified M,r section for use with m-derived terminating half-sections.

provided by an m-derived half-section with $m = 0{\cdot}684$ is a vastly better match than a resistor (Fig. 4). An even better match, indeed so close that it was not really practicable to show it, is furnished by the mid-series impedance of an m-derived section with $m = 0{\cdot}920$. This can be arranged by the use of two half-sections at each end of the delay line, one with $m = 0{\cdot}920$ and $m = 0{\cdot}684$, but of course it will become more difficult to accommodate their combined group delay characteristic in the overall response. Nevertheless, this is perfectly practicable with long delays.

As an example, the calculation of a 1 μs delay line with $m = 0{\cdot}684$ terminating half-sections will be carried out, using as initial values $M = 1{\cdot}39189$, $r = 0{\cdot}15485$; these may need to be modified subsequently to accommodate the group delay characteristic of the terminations.

To start with, the reference point $x_1 = 0{\cdot}5$ is assigned to 4·433 MHz, that is to colour sub-carrier frequency. The cut-off frequency then becomes 8·860 MHz, making T_0 for a single section, quite fortuitously, equal to precisely 50 ns. A total of 20 sections is thus required for a 1 μs delay.

The addition of the terminating sections increases the total T_0 to 1024·56 ns, and the delay at $x = 0{\cdot}5$ to 1032·71 ns. No change in the number of sections is required, and the zero frequency delay can be restored to 1 μs merely by raising the cut-off frequency to 9·07657 MHz. The delay at $x = 0{\cdot}5$ then becomes 1007·96 ns, that is the sections must now be modified so as to provide a value of T_g/T_0 at $x = 0{\cdot}5$ of 0·9921.

The most convenient method of setting about this is to notice that T_0 is a function of M only (Fig. 3), so that if M is kept constant and M^2r is varied, the shape of the delay curve will change without any modification of the zero-frequency delay. The new value of M^2r can only be found by approximate methods; the simple device employed in this instance was the calculation of T_g/T_0 at $x = 0{\cdot}5$ for three values of M^2r around the expected value, after which the final figure was quickly arrived at by standard interpolation techniques. The result is $M^2r = 0{\cdot}26545$, compared with $M^2r = 0{\cdot}30$ initially.

The element values of a section for an impedance of 75 ohms are given in Fig. 6, and the calculated performance of a cascade of 20 such sections, including the terminating half-sections, is shown in Fig. 7. The insertion loss was omitted since it proved to be extremely small over the entire range. The overall performance is obviously of a very high order. There is little doubt that the maximum deviation of group delay over the video band, small though it is, could be improved still further by a closer approximation to the optimum values of M and r, but it was decided that the viability of the M,r section had already been sufficiently well justified. In any case, the flatness of the band would be improved with longer delays.

3.5 *Complex Conjugate m-derived Sections*

It is possible to imagine an m-derived low-pass filter section with a complex m value, although it would not be physically realizable. However, it can readily be shown that the lattice network corresponding to the series connexion of a pair of sections with complex conjugate m values has all its element values real, as is demonstrated by Fig. 8, and must therefore be realizable in unbalanced form with the aid of a transformer. Since each such network possesses a pair of independently variable parameters, m_1 and m_2 say, the possibility again exists of effecting a wide degree of linearization of the group delay characteristic.

The network was first encountered by the writer in a paper by Golyshko and Silvinskaya,[14] who proposed a method of dimensioning the sections so as to obtain an equi-ripple group delay response over a chosen fractional bandwidth. A number of lines with delays up to 1 μs were successfully constructed on this principle for use within the BBC.[15] The maximum bandwidth was 3 MHz, corresponding to the 405-line system in use at that period.

These delay lines yielded very good results by the then existing standards, even though the equi-ripple response provided was by no means ideal as regards transient distortion.[16] However, considerable difficulty during manufacture was experienced in the construction and

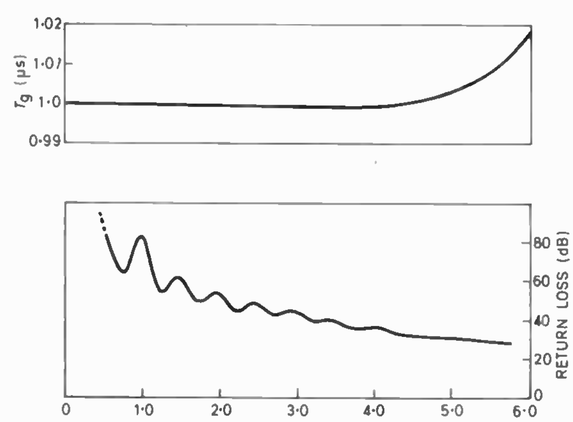
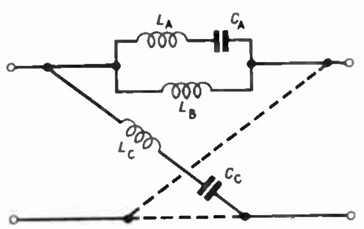


Fig. 7. 1 — μs M,r delay line with m-derived terminations.



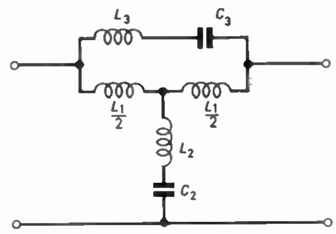
$$L_A = \frac{2m_1}{s-1} L_0;\ C_A = \frac{s-1}{2m_1} C_0;\ L_B = 2m_1 L_0$$

$$L_C = \frac{s}{2m_1} L_0;\ \ C_C = 2m_1 C_0$$

The individual sections have 'm' values $m_1 \pm jm_1$,

$$s = 1 + m_1^2 + m_2^2,\ L_0 = \frac{R_0}{\omega_C},\ \text{and } C_0 = \frac{1}{\omega_c R_0}.$$

Fig. 8. The combination of two low-pass sections with complex conjugate m values.



$$L_1 = 4m_1 L_0;\ L_2 = -\frac{4m^2 - s}{4m_1} L_0;\ C_2 = 4m_1 C_0;$$

$$L_3 = \frac{4m_1}{s-1} L_0;\ C_3 = \frac{s-1}{4m_1} C_0.$$

$$s = 1 + m_1^2 + m_2^2,\ L_0 = \frac{R_0}{\omega_c},\ C_0 = \frac{1}{\omega_c R_0}.$$

Fig. 9. Unbalanced form of lattice of Fig. 8 after Golyshko and Silvinskaya.

alignment of the networks in the unbalanced configuration originally proposed (Fig. 9), the principal causes being the self-capacitance of the transformer windings and the stray capacitance to earth of the junction of L_3 and C_3. Since the need for long delays was not particularly pressing at that time, no further work was carried out.

A later re-appraisal of this network revealed that an alternative unbalanced development from the lattice of Fig. 8 has considerable practical advantages, and in fact is superior in all respects to that originally proposed. In particular, as can be seen from Fig. 10, a capacitance is available to absorb each of the stray capacitances of the section, unlike the version proposed by Golyshko and Silvinskaya.[14] It will be noticed that in principle the series-arm capacitor C_1 can assume a negative value, but fortunately with the optimum m values for flat delay, the capacitance is small but positive. However, if a high value is preferred to a low, the π-configuration of capacitors formed by C_1 and C_3 may be replaced by its equivalent star configuration.

3.5.1 Dimensioning of complex conjugate m-derived sections

The relative group delay for image terminations is given by

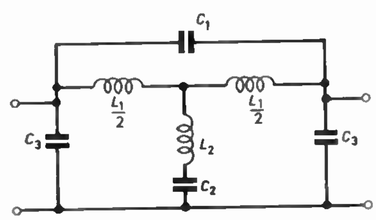
$$\frac{T_g}{T_0} = \frac{1+(s-2)x^2}{[1+2(2m_1^2-s)x^2+(s^2-4m_1^2)x^4]\sqrt{1-x^2}} \qquad (7)$$

where $T_0 = 2m_1/\pi f_c$, $s = 1+m_1^2+m_2^2$, and $x = f/f_c$. The individual m values are $m_1 \pm jm_2$, but the expressions are simplified if m_2 does not appear explicitly.

A brief investigation quickly revealed that flat group delay characteristics require m_1 to be in the neighbourhood of 1·33, which means that T_0 is now slightly more than twice the corresponding value for a section with $m > 1$. An alternative value $m_1 = 0{\cdot}48$ gives a non-physical unbalanced form.

The procedure used for the design of a 1 μs line will again be briefly indicated. The value of x_1, corresponding to colour sub-carrier frequency will be allocated, slightly conservatively, the initial value of 0·45. From this the cut-off frequency must be about 10 MHz, and 12 sections will be required. T_0 then becomes 83·333 ns, and f_c is now 10·160 MHz.

In order to improve the matching a pair of terminating half sections with $m = 0{\cdot}684$ may be added. It then becomes necessary to adjust the cut-off frequency to 10·378 MHz in order to restore the zero-frequency delay to 1 μs. Furthermore, the value of T_g/T_0 at subcarrier frequency must now be made 0·9948 instead of unity. When the known quantities $x_1 = 0{\cdot}45$ and $m_1 = 1{\cdot}33$ are inserted in the resulting equation, a quadratic is obtained



$$L_1 = 4m_1 L_0;\ C_1 = \tfrac{1}{2}\left(\frac{s}{2m_1} - \frac{2m_1}{s}\right) C_0;$$

$$L_2 = -\left[m_1 - \frac{s^2}{4m_1(s-1)}\right] L_0;\ C_2 = \frac{4m_1(s-1)}{s} C_0;$$

$$C_3 = \frac{2m_1}{s} C_0;\ L_0 = \frac{R_0}{\omega_c};\ C_0 = \frac{1}{\omega_c R_0}$$

If L_1 and L_2 are provided by a transformer with coupling coefficient k, then

$$k = \frac{4m^2(s-1) - s^2}{4m_1^2(s-1) + s^2} \text{ (windings in series aiding).}$$

The group delay

$$T_g = T_0 \times \frac{1+(s-2)x^2}{[1+2(2m_1^2-s)x^2+(s^2-4m_1^2)x^4]\sqrt{1-x^2}}$$

where $T_0 = \dfrac{2m_1}{\pi f_c}$ and $x = f/f_c$.

The image impedance $\quad Z_I = \dfrac{R_0}{\sqrt{1-x^2}}.$

Fig. 10. Pair of complex conjugate m sections: preferred unbalanced configuration.

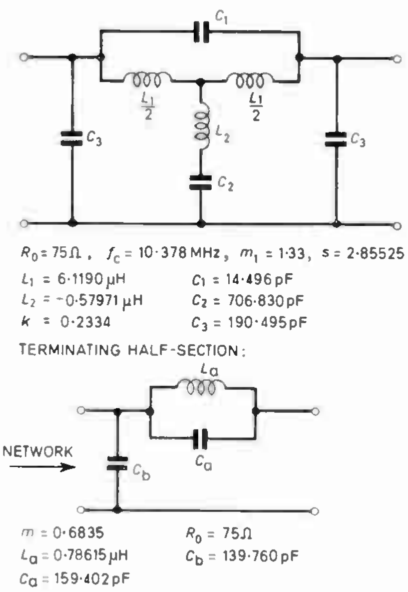


Fig. 11. Component values for a 1000 ns delay line employing complex conjugate m sections.

from which s is found to be 2·85525. The corresponding element values are shown in Fig. 11.

The computed delay characteristic of a cascade of 12 sections, including terminating half-sections, is given in Fig. 12. The deviations from the target of 1000 ns are less than 1 ns up to 5 MHz, where the delay increases to 1001 ns, finally reaching 1004 ns at 5·5 MHz. The return loss was not plotted since over this range it nowhere dropped below 74 dB. These results are extremely satisfactory.

As a matter of interest, the performance of 12 sections of this type without terminating half-sections was also calculated, the cut-off frequency being restored to the original value of 10·160 MHz in order to make the zero-frequency response 1 µs once again. The mid-frequency response, as is clear from Fig. 13, exhibits the droop deliberately introduced in order to compensate for the delay of the terminating half-sections, but at the higher frequencies this is counteracted by reflective effects from the increasing mismatch. The return loss is also better than might have been anticipated.

4 Practical Verification

4.1 *Choice of Section*

A very useful comparative figure of merit for a delay section is the expression $f_c T_0 x_2$, where x_2 is the highest normalized frequency up to which the delay is to be held flat. In this instance it may be replaced by x_1 since there is a fixed relationship between colour sub-carrier frequency and the upper limit of the video band. Then for the $m > 1$ section the figure of merit is in round figures 0·1, for the M, r section it is 0·25, and for the complex conjugate m section it reaches 0·4.

It is clear that in terms of numbers of inductors the complex conjugate m section possesses a very marked superiority over the other two types investigated. As in other respects it seemed to be eminently suitable for the purpose in mind, it was decided to standardize upon this type.

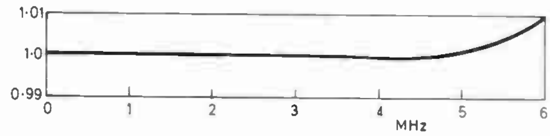


Fig. 12. Computed response of 12 sections $m = 1·33, s = 2·85525$ with $m = 0·6835$ terminating half-sections. The return loss is nowhere worse than 74 dB.

4.2 *Construction of Model*

A prototype 1 µs delay line was constructed from 12 sections with the element values of Fig. 12, together with $m = 0·684$ matching end half-sections.

A preliminary investigation into the sensitivity of the complex conjugate m section to small changes in the component values had revealed that the self-inductance of L_1 is the most sensitive parameter since it simultaneously affects both the low and high frequency delays to a marked extent. It was accordingly wound on a miniature adjustable ferrite core so that it could be resonated *in situ* against an accurately known capacitor, a procedure which experience has shown to be very reliable. A Q-factor of 140 over a large part of the designed frequency band was achieved. Furthermore, to ensure flexibility in testing the delay line, the precise adjustment of the mutual inductance of L_1 was achieved by means of a bifilar winding, the high negative mutual inductance of which was reduced to the correct amount by another inductor in series with the centre tap.

The advantage of this procedure is to some extent offset by the rather large self-capacitance of a bifilar winding even though, as was shown by Starr,[17] the effective value is only one-third of that measured between the windings at low frequency. By this means, the effective capacitance was found to be 4 pF, leaving 10·5 pF to be added externally. The sensitivity calculation predicted that variations up to $\pm 10\%$ in C_1, that is ± 1 pF, would be just acceptable. Nevertheless, for this preliminary investigation at all events, it was decided to employ a miniature variable capacitor in that position. An

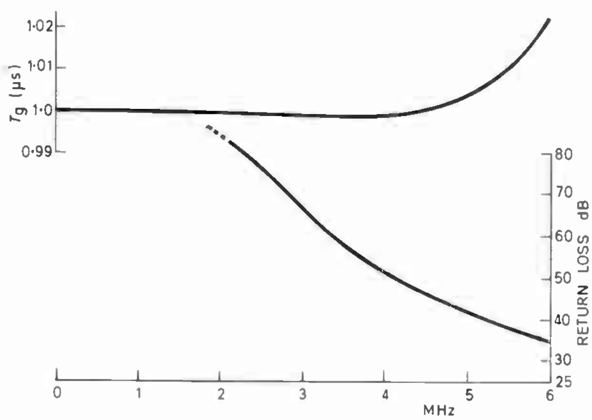


Fig. 13. Response of 12 sections $m_1 = 1·33$, $s = 2·85525$, $f_c = 10·160$ MHz, with resistive terminations.

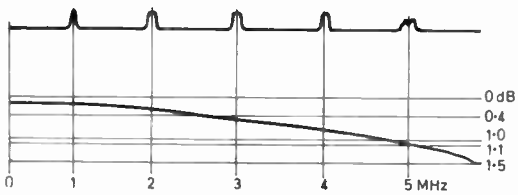


Fig. 14. Insertion loss of 1 μs delay line.

alternative procedure, as has already been mentioned above, would have been the conversion of the π-configuration of capacitors into the equivalent star-network, resulting in two capacitors with roughly the same value as C_3 and a coupling capacitor around 2·8 nF, which is not unmanageable. The remaining capacitors in the network were all purchased within $\pm 1\%$ of the specified value, no attempt at selection being made.

4.3 *Testing of Delay Line*

4.3.1 Steady-state tests

When the network had been aligned, both T_0 and the delay at colour sub-carrier frequency were found to be 1001 ns. The extremely small delay distortion elsewhere is best judged from the waveform tests given below, which in any case are always a more dependable guide to the performance of equipment handling video signals than any steady state measurement. The insertion loss and return loss characteristics were measured by means of a sweep method; the results were automatically recorded by an X–Y plotter.

The insertion loss thus obtained is shown in Fig. 14. The change in loss up to 5 MHz is 0·9 dB, which agrees with the value expected by application of Mayer's theorem. The return loss characteristic in the one direction (Fig. 15) is better than 35 dB up to 5 MHz, and only slightly worse in the other (Fig. 16), the close similarity between the two curves demonstrating the high degree of symmetry achieved in the construction. The reason for the rapid deterioration of the return loss above 5 MHz is not clear, but in any event it is not of great significance for the intended purpose.

4.3.2 Waveform tests

In the writer's opinion the most searching and convincing measurements are those carried out by the use of the standard television test waveforms. Full details on these and their methods of use will be found elsewhere.[2, 3]

The measurements of principal interest in the present context are the $2T$ sine-squared pulse, and the $10T$ composite chrominance pulse. The former is a pulse of accurately standardized shape whose spectrum is substantially confined within a bandwidth of 5 MHz, thereby enabling it to ignore effects above the video band which are irrelevant as far as the television signal is concerned. Its sensitivity to deviations in loss and group delay over the lower half of its spectral range is very high, and it may therefore be considered as a test of the luminance region of a colour television signal in particular.

The composite chrominance pulse complements this by confining its effect to the chrominance region only. It takes the form of another sine-squared pulse, but this time of 1 MHz spectral range, which is linearly added to itself after modulation by colour sub-carrier to a depth of 100%. When the amplitudes and timing are precisely adjusted, the lower lobe of the modulated wave just disappears, leaving a flat base-line. This is a condition of very exact balance, which reveals even very small errors in relative amplitude and delay by characteristic deformations of the base-line, symmetrical for gain and skew-symmetrical for delay. Under optimum conditions inequalities as low as 0·05 dB and 2 ns can be measured using specially designed apparatus,[2] and even smaller values can be detected by a comparison of input and output.

The two sine-squared pulses are complemented for general video measurements by corresponding bar waveforms, one important function of which is to show up more clearly certain long-duration effects. Some mid-band distortions are also revealed by typical deformations of the luminance bar transitions.

Photographs of the composite chrominance pulse at the output and input respectively of the delay line are given in Figs. 17(a) and (b). The upward bowing of the

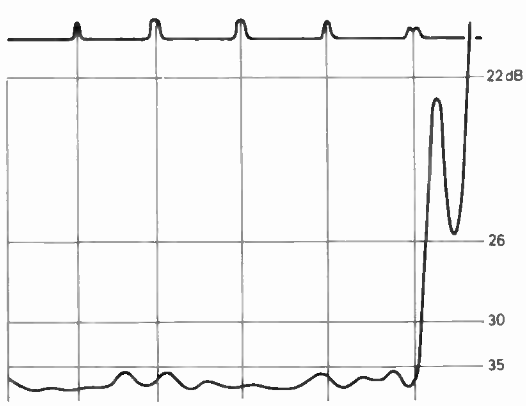


Fig. 15. Return loss of 1 μs delay line A → B.

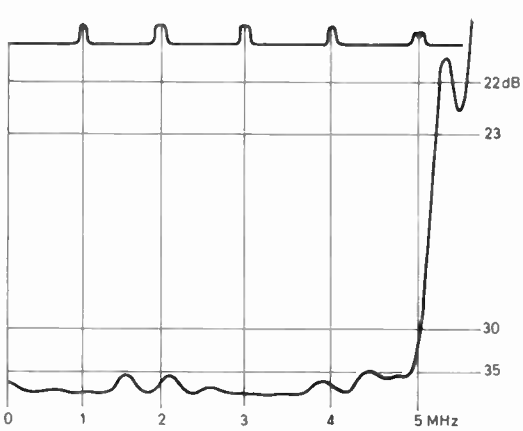
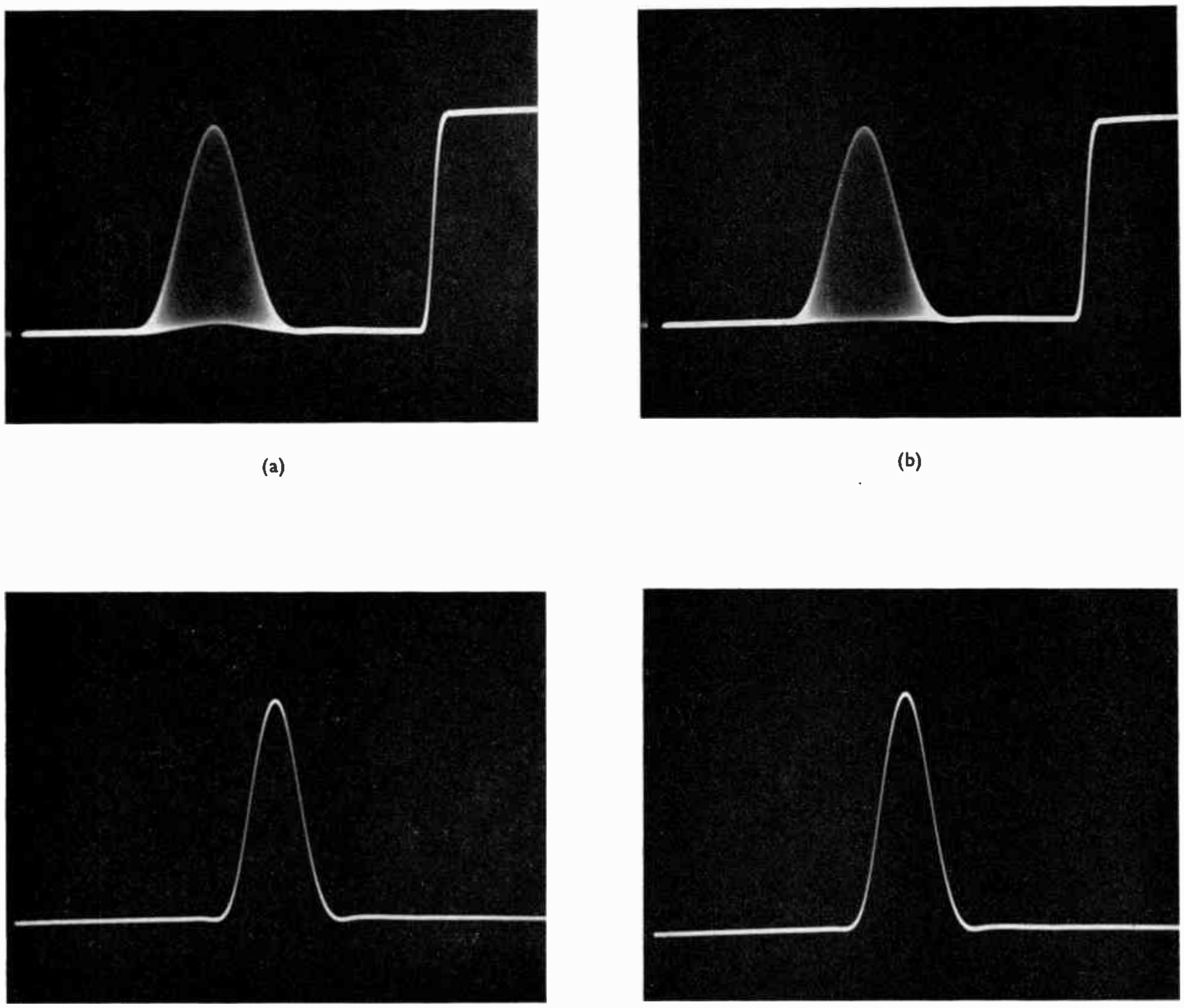


Fig. 16. Return loss of 1 μs delay line B → A.



(a) (b)

(c) (d)

Fig. 17. Waveform response of the delay line shown in Fig. 11 (unequalized).

baseline in Fig. 17(a) is characteristic of an increased loss at colour sub-carrier frequency, known here to amount to 0·9 dB. For reasons given immediately above, the leading edge of the luminance bar has been included in each instance. The corresponding output and input sine-squared pulses are shown in Figs. 17(c) and (d).

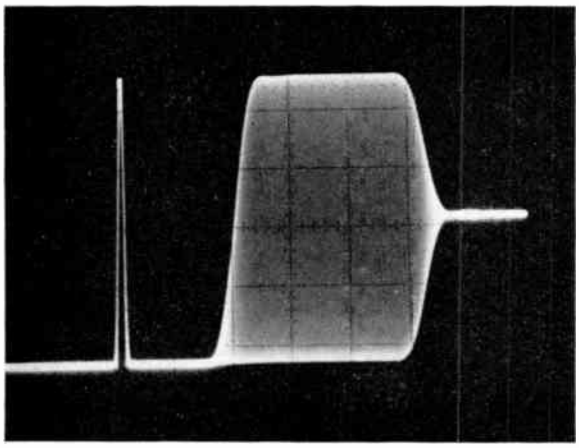
For the sake of completeness, the composite chrominance bar at the output of the delay line has been included in Fig. 18(a), even though its contribution in this instance is not very significant. Of much greater interest is the highly magnified sine-squared pulse of Fig. 18(b) showing the train of small echoes resulting from terminal and internal mismatches. Their amplitude is slightly less than 1 mV peak-to-peak for a test signal amplitude of 1 V peak-to-peak, i.e. an echo attenuation of somewhat more than 60 dB, which is highly satisfactory.

4.3.3 Equalization

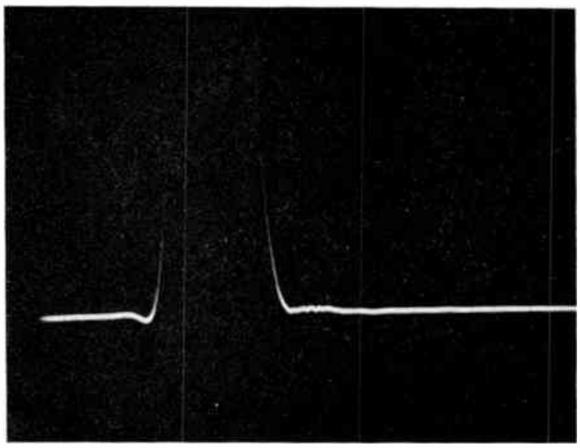
Before any final waveform testing can be undertaken, the delay line must be compensated for the insertion loss across the band resulting from dissipation; in this instance the error was 0·9 dB at 4·4 MHz. The standard method is the use of a constant-resistance equalizer-section in series with the line, the series arm of which only rarely needs to be more complicated than a capacitor and resistor in parallel. The delay of the equalizer can usually be neglected.

In practice by far the quickest and the most satisfactory means of finding the optimum component values of the equalizer is the following. The delay line is arranged for normal testing by means of sine-squared pulse and bar waveforms, then two attenuator pads, say each of 10 dB minimum, are connected in series with it. A parallel combination of a variable resistor and a variable capacitor is next inserted in series between the two attenuator pads, and both are adjusted until the input and output composite chrominance pulses and sine-squared pulses are as closely alike as possible. These final values, but halved in the case of the resistor and doubled in the case of the capacitor, form the series arm of the constant-resistance equalizer.

This process is very rapidly carried out, and has the

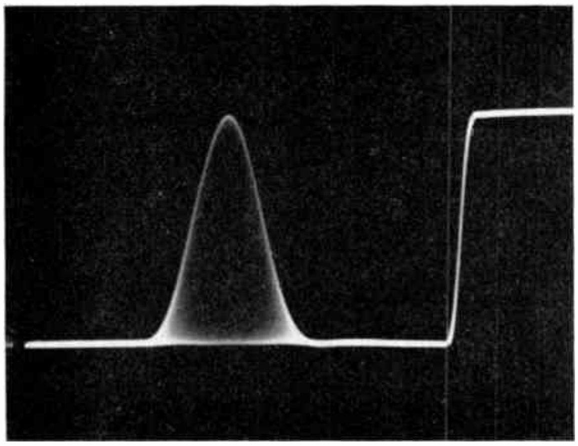


(a) Chrominance bar at output line.

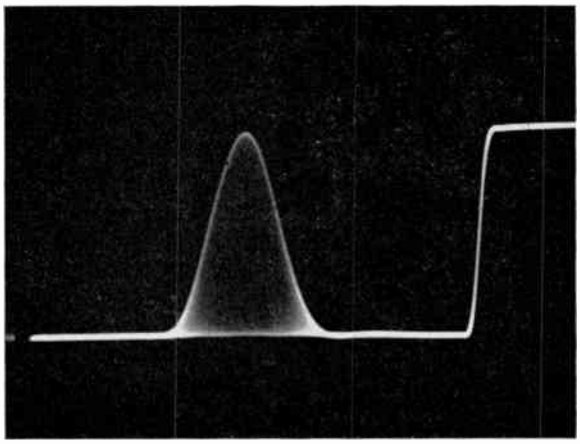


(b) Magnified $2T$ pulse showing reflexions.

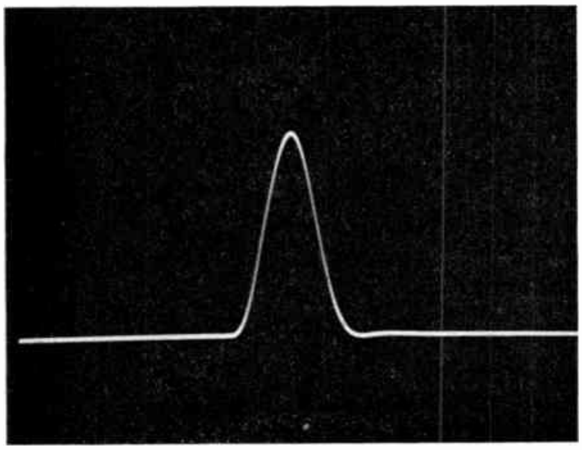
Fig. 18. Waveform response of the delay line of Fig. 11 (unequalized).



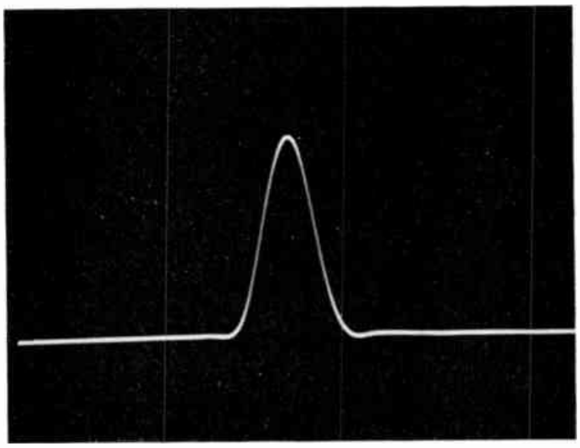
(a)



(b)



(c)



(d)

Fig. 19. Waveform response of delay line of Fig. 11 (equalized).
(a) and (c) are output waveforms; (b) and (d) are input waveforms.

advantage, apart from not requiring any calculation, that a 'best-fit' is always obtained. It is of great importance to achieve this optimum by a comparison of the input and output waveforms, since the inherent distortion of even very high-grade oscilloscopes is frequently found to be large enough to affect the measurement to a significant degree. This should be taken more as an indication of the sensitivity of sine-squared pulse and bar testing, than as a criticism of commercial oscilloscopes.

The result of this process is given in Figs. 19(a) and (c) where it will be found extremely difficult to distinguish the output from the input waveform in each instance. This must be counted as a remarkably fine performance for a 1 μs delay line, and fully justifies the method of design.

4.3.4 Resistance compensation

An alternative, internal compensation for the effects of dissipation may be effected by means of a resistor R_c connected from the centre-tap of each transformer L_1 to ground. This is shown in Fig. 20(a), where the transformer is assumed to have unity coupling coefficient, and the inductor L_1 of Figs. 10 and 11 then becomes equal to $L_2' - \frac{1}{4}L_1$. The dissipation is supposed to be capable of representation by a resistor R_s in shunt with L_1.

A lattice equivalent, derived immediately by means of Bartlett's Bisection Theorem, is given in Fig. 20(b). Each reactance arm has a resistor in parallel, and if these are made equal, i.e. $\frac{1}{2}R_s = 2R_c$, they may be extracted from the lattice and placed across the terminals of the network. The final unbalanced equivalent, Fig. 20(c), is now an ideal, dissipationless network, except for the resistors

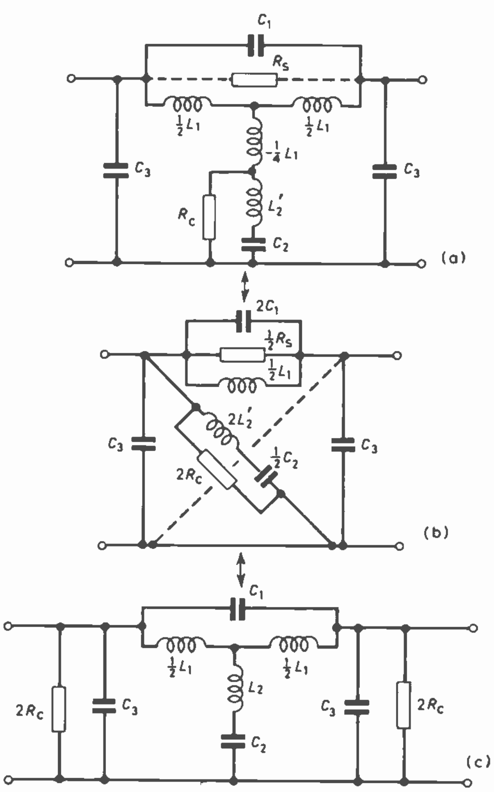


Fig. 20. Derivation of resistance compensation method.

across input and output. Since these are normally high compared with the image impedance one would not expect them to have any serious detrimental effect on the network behaviour so long as the frequency is high enough for them to be considered as distributed. On the other hand, as zero frequency is approached all resistors will tend to appear lumped across the input and output terminations of the delay line, and one would expect the return loss to decrease fairly sharply.

The experimental delay line was compensated in this manner. Assuming a slightly pessimistic Q of 120 for L_1, to account for slight dissipation elsewhere, the equivalent shunt resistance at colour sub-carrier frequency becomes, in round figures, 20 kΩ, making R_c equal to 5 kΩ. The nearest standard value of 5·1 kΩ was accordingly connected between the centre tap and earth on each of the 12 sections forming the line. In principle, some error should have been introduced from the fact that the coupling coefficient of L_1 was around 0·9 instead of unity, but experience elsewhere with this type of resistance compensation has always shown that in practice it is remarkably tolerant in this respect, to the extent that some degree of compensation is available even with R_c connected across C_2 only.

The efficacy of this form of compensation as far as the waveform response is concerned is clearly demonstrated by a comparison of Figs. 21 and 19, the latter corresponding to equalization by means of a constant resistance network. The return loss (Fig. 22) is degraded, as expected, by the expected sharp rise towards zero frequency, and also by a rise at the higher frequencies, but the network would nevertheless still be perfectly acceptable for a wide range of purposes. However, if a better return loss is essential it can still be obtained at the expense of some extra capacitors and resistors. To achieve this, the delay line must be manufactured from individual sections, that is the pairs of adjacent shunt capacitors C_3 are not compounded into single capacitors of value $2C_3$. Then the successive sections are connected through small resistors of such a value that they form an attenuator pad in conjunction with the pair of effective shunt resistors $2R_c$ (Fig. 20(c)).

The tolerance of the method of resistance compensation is so great that the equalization of the entire delay line may even be effected by means of a single resistor R_c in one section only; which is used is not very material. Figure 23 demonstrates clearly that as far as the waveform response is concerned, a single resistor of 330 Ω in one of the end sections is quite as effective as resistors of 5·1 kΩ distributed throughout the line, or indeed as a constant resistance equalizer. On the other hand, as is to be expected, the return loss (Fig. 24) is considerably degraded. Nevertheless, the line would still be perfectly acceptable when used between resistive terminations.

It might be pointed out in passing that the amount of control which R_c can exert on the luminance-chrominance gain ratio is very considerable, and it is possible to envisage the use of this as a simple method when a delay line is present of adjusting that parameter. Likewise, a range of control of the luminance-chrominance delay inequality is available without a change in amplitude or

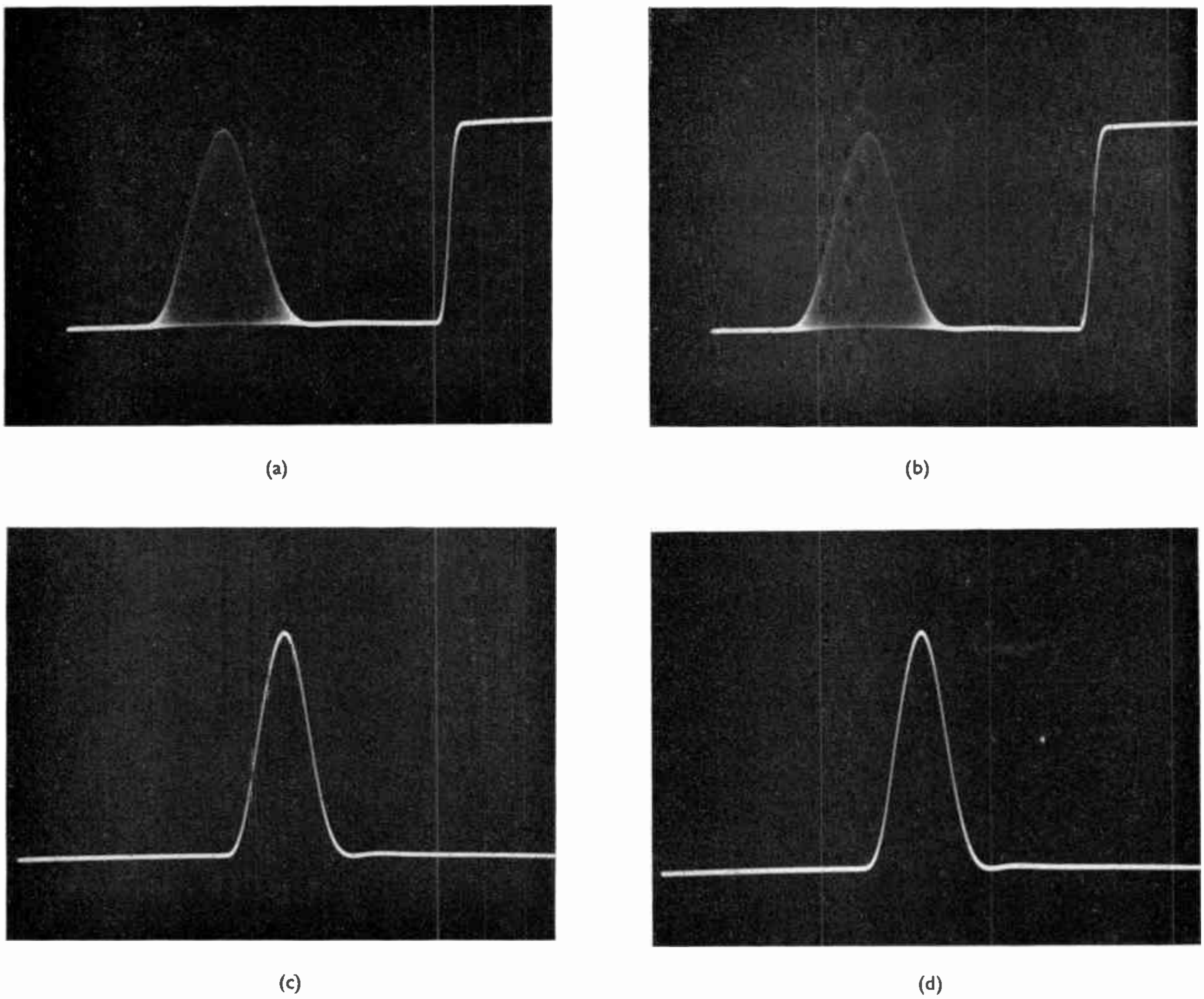


Fig. 21. Waveform response of delay line compensated by a resistor in each section. (a) and (c) are output waveforms; (b) and (d) are input waveforms.

zero-frequency delay by adjustment of C_1 or L_2. In both instances, remote or automatic control is feasible by the use of voltage-controlled components.

Finally, it should not be overlooked that the type of resistance compensation described above is also applicable to other types of network employing a transformer, for example the low-pass sections with $m > 1$ and M,r sections.

5 Conclusions

A critical examination of three types of delay sections has shown that two of them are markedly superior to the type hitherto widely employed, for the building of long delay lines for video applications to the very stringent BBC specifications. The M,r section is novel and has a number of specific advantages, but on balance the complex conjugate m type in the proposed new configuration is judged to be superior for this type of application. Its use makes possible the manufacture of delay lines for the video band with a significantly improved performance and a considerable reduction in the number of inductors compared with present commercial designs. The construction of a 1 μs delay line has confirmed that a performance close to the theoretical can be obtained with-

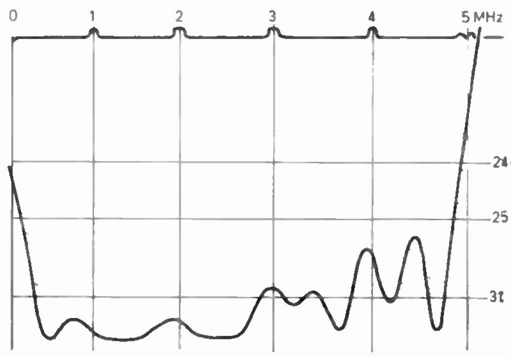


Fig. 22. Return loss characteristic corresponding to Fig. 21 compensation by means of 5·1 kΩ resistor in each section.

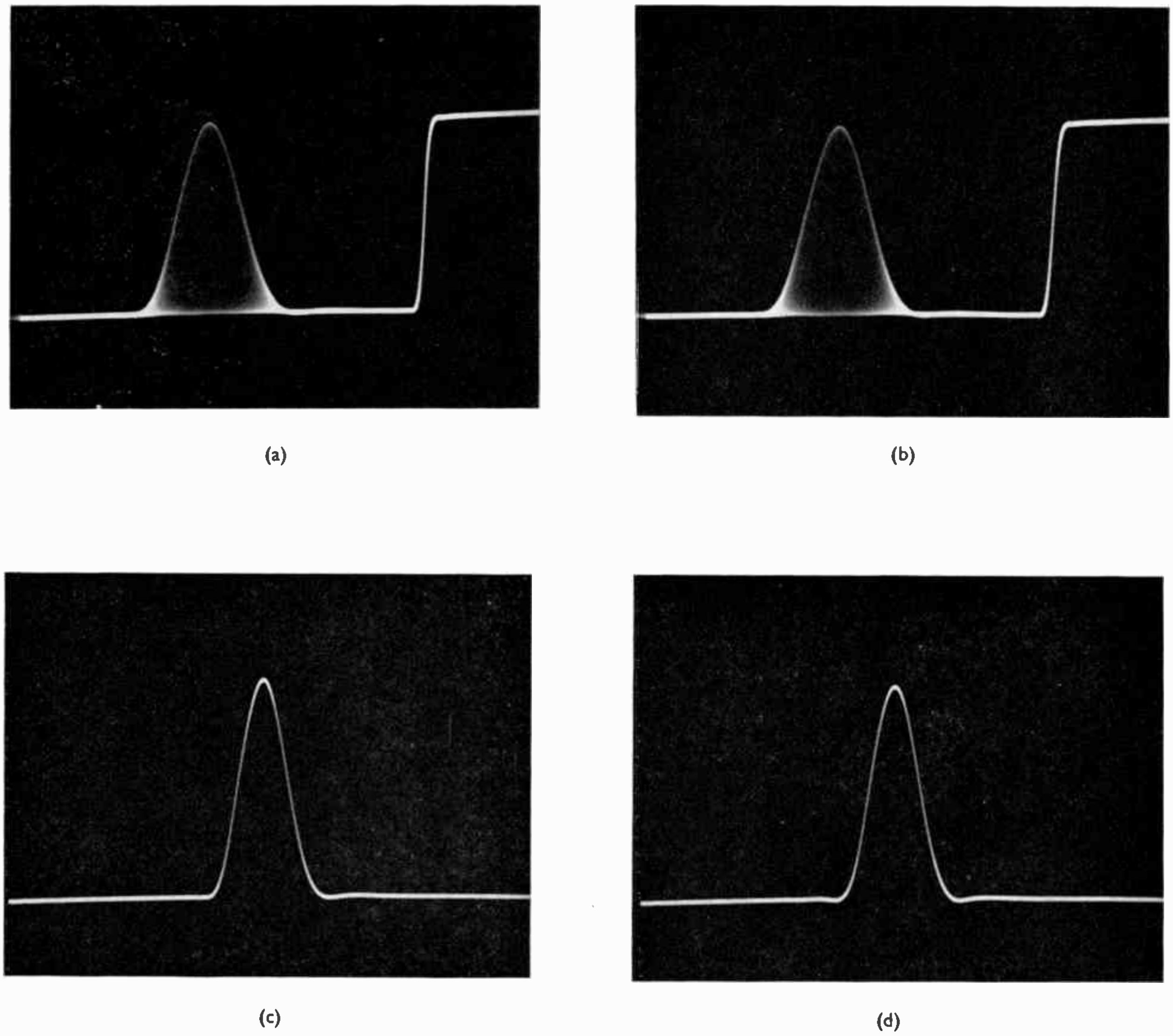


Fig. 23. Waveform response of delay line compensated by a single resistor in one end-section.
(a) and (c) are output waveforms; (b) and (d) are input waveforms.

out undue difficulty. It is believed this represents a decided advance in lumped-constant delay line techniques.

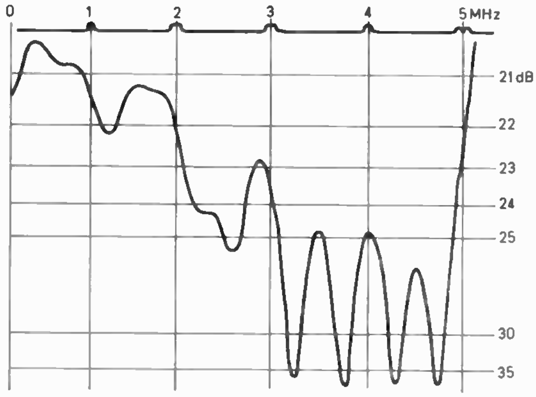


Fig. 24. Return loss characteristic corresponding to Fig. 23. Single 330 Ω resistance compensation.

6 Acknowledgments

The author's thanks are due to Mr. D. C. Broughton for his assistance with the computation of the overall delay line responses, to Mr. B. P. George for carrying out the experimental work, and to the Director of Engineering of the BBC for permission to publish this paper.

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STANDARD FREQUENCY TRANSMISSIONS— October 1973

(Communication from the National Physical Laboratory)

Oct. 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)		Oct. 1973	Deviation from nominal frequency in parts in 10^{10} 24-hour mean centred on 0300 UT)			Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	0	0	−0·1	687	598·3	**17**	−0·1	−0·1	0	691	606·1
2	0	0	0	687	601·5	**18**	0	0	0	691	606·4
3	+0·1	+0·1	0	686	601·7	**19**	0	0	0	691	606·8
4	−0·1	−0·1	+0·1	687	602·3	**20**	−0·1	0	−0·1	692	607·0
5	0	−0·1	0	687	603·1	**21**	0	0	0	692	607·1
6	−0·1	−0·1	−0·1	688	603·9	**22**	0	0	0	692	607·0
7	−0·1	0	−0·1	689	604·3	**23**	0	0	0	692	607·0
8	−0·1	−0·1	+0·1	690	605·2	**24**	0	0	+0·1	692	606·9
9	−0·2	−0·1	+0·1	692	605·9	**25**	0	0	+0·1	692	606·8
10	+0·1	0	0	691	605·7	**26**	+0·1	+0·1	+0·1	691	606·0
11	0	+0·1	0	691	604·8	**27**	0	0	0	691	605·8
12	+0·1	+0·1	0	690	604·1	**28**	0	0	+0·1	691	605·6
13	+0·1	0	0	689	604·1	**29**	0	+0·1	+0·1	691	604·8
14	0	−0·1	−0·1	689	604.8	**30**	0	+0·1	+0·1	691	603·4
15	−0·2	−0·1	−0·1	691	605·3	**31**	+0·1	0	0	690	603·4
16	+0·1	0	−0·1	690	606·5						

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

* Relative to UTC Scale; (UTC_{NPL} − Station) = + 500 at 1500 UT 31st December 1968.

† Relative to AT Scale; (AT_{NPL} − Station) = + 468·6 at 1500 UT 31st December 1968.

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Static voltages on the guy insulators of m.f. and l.f. broadcast tower antennas

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SUMMARY

A new approximate method for calculating the static voltages on the guy insulators of m.f. and l.f. broadcast tower antennas is given. The method is based on an integral-equation technique, where the integral equation is solved by the so-called point matching method; the unknown charge densities are approximated by polynomials. The polynomial expansion provides very significant advantages, primarily because it leads to the integrals in closed form, and because it requires a very low order of polynomial. A numerical example is given to illustrate the method.

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1 Introduction

The overcrowding of medium and low frequency bands and resulting interference in the European broadcasting area often force broadcasting organizations to increase the powers of their transmitters and powers of the order of 1000 kW are not infrequent both in Europe and in neighbouring countries. In this connexion the problems concerning the voltages and insulating material of broadcast tower antennas as well as the methods of calculating these voltages are becoming again of growing importance.

The first thorough study of radio-frequency voltages appearing on the insulators of guy ropes was made in 1939 in the classical paper by Brown[1]. Though the voltages resulting from the electrostatic field during thunderstorms have not been theoretically treated in Brown's paper, he pointed to these voltages as a cause liable to produce sparks across the guy insulators. Once the static voltage has broken the gap, the r.f. voltage maintains the arc even if it were much less than the voltage required to start the arc. In order to overcome these difficulties Brown proposed the use of high-resistance leaks across the guy insulators which could clear up all troubles due to the static charge. However, according to widespread experience, a direct or near stroke of lightning may destroy the leak resistors entirely. At any rate, a quantitative knowledge of the static voltages and the methods for their determination deserve to be investigated.

In a recent paper Bruger[2] has given an interesting method for calculating both the r.f. and static voltages on the guy insulators. As stressed by the author himself, in this method emphasis was laid on the clarity and easy applicability, instead of on extreme precision.

The purpose of the present paper is to develop a more accurate, but conceptually simple method for calculating the static voltages on the guy insulators. Essentially, the method is based on an integral equation technique, where the integral equation is approximately solved by the so-called point-matching method.[3] The unknown sub-integral functions, representing in the present case the charge densities per unit length of conductors, can be approximated by functional series of convenient functions with unknown coefficients. In this paper, however, a polynomial approximation is adopted. Besides the conceptual simplicity, the polynomial expansion offers very significant advantages in the necessary computational work on the electronic computer. Apart from the fact that the true distribution function can be well approximated by a low-order polynomial, the polynomial approximation leads to integrals which, by means of the recurrent formulas, can be reduced to elementary integrals.

2 Static Field in the Vicinity of the Antenna Tower

Most of the m.f. and l.f. transmitting antennas are built in the form of guyed steel towers of uniform cross-section. For the purpose of analysis they can be approximated by cylinders[4] of equivalent radius a. The supporting guy wire-ropes are broken up at several points by insulators. In addition to the r.f. voltages induced in the guy sections, before and during thunderstorms, the insulators are exposed to high quasi-static voltages produced by very strong atmospheric electric fields. For the sake of simplicity we assume that this field is static and, in the absence of the antenna tower, is homogeneous having a vertical field strength E_0.

As the first step in determining the static voltages on the guy insulators, the resultant field in the vicinity of the tower has to be calculated. In this calculation, the presence of the guys could be ignored, because induced charges on them are too small to affect appreciably the charge distribution along the tower. The validity of this assumption will be checked numerically in Section 4. Since the base of the antenna tower is earthed through the static drain coil, the tower is at the static potential of the ground, which is taken as a reference. In order to avoid dealing with an infiite ground plane, the method of images will be used. So, the ground plane is replaced by the electrical image of the cylinder representing the antenna tower. Both the antenna cylinder and its image, with height h and radius a, are shown in Fig. 1. The z-axis of cylindrical co-ordinates r, ϕ, z, coincides with the axis of the cylinders.

Let the unknown charge density per unit length of the antenna cylinder be denoted by $q(z)$. According to the method of images the following symmetry condition must be fulfilled:

$$q(-z) = -q(z). \tag{1}$$

If we denote

$$\phi_0 = -E_0 z \tag{2}$$

the scalar potential of the unperturbed atmospheric electric field and by ϕ_i the scalar potential created by the induced charges on the antenna, the initial equation expressing the boundary condition for potential has the form

$$\phi_0 + \phi_i = 0\Big|_{r=a,\,-h\leqslant z\leqslant h}. \tag{3}$$

Since $h \gg a$, in calculating the potential ϕ_i at a point $P(r \geqslant a, z)$ it can be assumed that the induced charges are localized on the axis of the cylindrical conductor. So, we have

$$\phi_i = \frac{1}{4\pi\varepsilon_0}\int_{-h}^{h} \frac{q(z')\,dz'}{\sqrt{r^2+(z-z')^2}}. \tag{4}$$

Introducing equations (4) and (2) into equation (3), and taking into account the symmetry condition (1), we obtain the fundamental integral equation in the form

$$\int_0^h \frac{q(z')\,dz'}{\sqrt{a^2+(z-z')^2}} - \int_0^h \frac{q(z')\,dz'}{\sqrt{a^2+(z+z')^2}} = 4\pi\varepsilon_0 E_0 z. \tag{5}$$

This equation is automatically satisfied for $z = 0$.

Solving equation (5), the unknown charge distribution can be determined and, consequently, the resultant potential is

$$\phi = \phi_0 + \phi_i. \tag{6}$$

Although an exact method for solving equation (5) does not exist, there are several methods for solving it approximately. A simple, but very general approximate method is the above mentioned point-matching method. According to this method the unknown charge distribu-

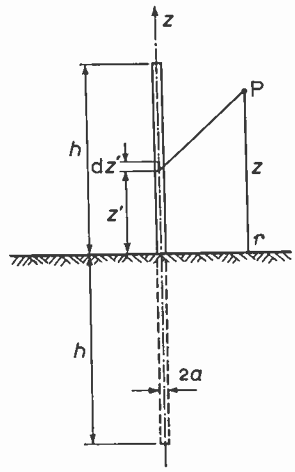

Fig. 1. Cylinder representing the antenna tower and its image.

tion can be approximated by a finite functional series with unknown coefficients. These coefficients can be determined by satisfying the integral equation at a sufficient number of points along the conductor. Though the basic functions in the form of rectangular and triangular pulses are in common use, we will choose the polynomial series. In solving equation (5) the polynomial series have an outstanding advantage because they lead to the integrals which are reducible to the elementary ones. On the other hand, polynomials of relatively low order approximate the charge distribution in a very satisfactory manner. So, using polynomials, both the number of integrals and, especially, the computation time for their evaluation are extremely favourable.

In order to make further analysis irrespective of the height h of the tower, let us first normalize all the lengths, so that

$$u = z/h; \qquad u' = z'/h; \qquad A = a/h \tag{7}$$

and, consequently

$$q(z') = dQ/dz' = q(u')/h. \tag{8}$$

With the new, normalized coordinates the integral equation (5) can be written in the form

$$\int_0^1 \frac{q(u')\,du'}{\sqrt{A^2+(u-u')^2}} - \int_0^1 \frac{q(u')\,du'}{\sqrt{A^2+(u+u')^2}} = 4\pi\varepsilon_0 E_0 h^2 u. \tag{9}$$

Let us now approximate the unknown charge density function by the polynomial

$$q(u') = \sum_{n=1}^{N} B_n u'^n \qquad u' \geqslant 0, \tag{10}$$

where N is the order of the polynomial and B_n are the coefficients to be determined. The constant term B_0 is set equal to zero because the polynomial (10) must satisfy

$$q(0) = 0, \tag{11}$$

imposed by the symmetry condition (1).

With the charge density distribution function (10), equation (9) becomes

$$\sum_{n=1}^{N} B_n J_n(u) = 4\pi\varepsilon_0 E_0 h^2 u \tag{12}$$

where

$$J_n(u) = P_n(u) - P_n(-u), \tag{13}$$

and

$$P_n(u) = \int_0^1 \frac{u'^n \, du'}{\sqrt{A^2+(u-u')^2}}. \tag{14}$$

The integrals (14), for different values of n, can be reduced to elementary integrals:

$$\begin{aligned} P_n(u) &= \operatorname{arcsinh}\{(1-u)\}A + \operatorname{arcsinh}(u/A), \quad \text{for } n = 0 \\ &= \sqrt{A^2+(1-u)^2} - \sqrt{A^2+u^2} + uP_0(u), \quad \text{for } n = 1 \\ &= \{\sqrt{A^2+(1-u)^2} + (2n-1)uP_{n-1}(u) \\ &\quad -(n-1)(A^2+u^2)P_{n-2}(u)\}/n, \quad \text{for } n \geqslant 2. \end{aligned} \tag{15}$$

In order to determine N unknown coefficients B_n of the charge density function (10), we take N convenient points u_i along the antenna cylinder and stipulate that equation (12) be satisfied at these points. So, we get a system of N linear equations in N unknowns B_n:

$$\sum_{n=1}^{N} B_n J_n(u_i) = 4\pi\varepsilon_0 E_0 h^2 u_i, \quad i = 1, 2, \ldots, N. \tag{16}$$

Although the choice of points u_i is arbitrary, it is quite natural to select them equidistantly, i.e.

$$u_i = i/(N+1), \quad i = 1, 2, \ldots, N. \tag{17}$$

The end-point $u = 1$ is excluded because the function $q(u)$ is discontinuous there; the point $u = 0$ is already taken into account by (11).

By solving system (16) we get unknown coefficients B_n, and finally determine the charge density function (10). The resultant potential ϕ in the vicinity of the antenna tower can be calculated from equation (6). In final form, we can write

$$\phi = \frac{1}{4\pi\varepsilon_0 h} \sum_{n=1}^{N} B_n\{P_n(u, R) - P_n(-u, R)\} - E_0 hu, \tag{18}$$

where $P_n(u, R)$ is obtained from $P_n(u)$ by substituting A for $R = r/h$; r and $z = uh$ are cylindrical coordinates of the point wherein the potential is to be calculated.

If necessary, the cylindrical components of the field strength can be calculated by means of the formulae given in the Appendix.

3 Static Voltages on the Guy Insulators

Essentially, the same method will be used in determining the static voltages on the insulators of guy wire-ropes.

In order to fix the notation, a sectionalized antenna guy with its electrical image is shown in Fig. 2. There are M guy sections having lengths d_m, $m = 1, 2, \ldots, M$, and a radius b. The first guy section is directly anchored to the ground, without an insulator. Let a new coordinate system, x–y–z, be introduced and positioned in such a way that the z-axis coincides with the axis of the guy, and x-axis lies in the ground plane. The distances along the image of the guy are measured by the coordinate t, whose origin coincides with the origin of the x–y–z system.

The charge density distribution function $q(z)$ along the guy sections, as well as the voltages V_m on the guy insulators are determined by solving numerically a system of integral equations. These equations are derived by making the following simplifying suppositions:

(i) The charge distribution on the antenna tower is not disturbed by the presence of the guys. (This will be fully justified later by comparing the calculated charge densities per unit length on the tower and on the guys.)

(ii) Mutual effects between the guys are negligible and can be ignored.

(iii) The induced charges are localized on the axes of the guy sections.

Accordingly, the potential at a point $P(x, y, z)$ in the presence of only one guy can be written as the sum

$$\phi = \phi_0 + \phi_i + \phi_g + \phi_{gi} \tag{19}$$

where

ϕ_0 is the scalar potential of the unperturbed atmospheric electric field, given by (2) (note the change of the coordinates);

ϕ_i is the potential created by induced charges on the antenna tower and its image, the guys being neglected;

ϕ_g is the potential due to the induced charge on the guy;

ϕ_{gi} is the potential due to the charge on the guy-image.

According to the notation in Fig. 2, the two latter potentials can be written as follows:

$$\phi_g = \frac{1}{4\pi\varepsilon_0} \int_0^d \frac{q_g(z')\, dz'}{\sqrt{x^2+y^2+(z-z')^2}} \tag{20}$$

$$\phi_{gi} = \frac{1}{4\pi\varepsilon_0} \int_0^d \frac{q_{gi}(t)\, dt}{\sqrt{x^2+y^2+z^2+t^2-2t \times (z\cos 2\theta + x\sin 2\theta)}} \tag{21}$$

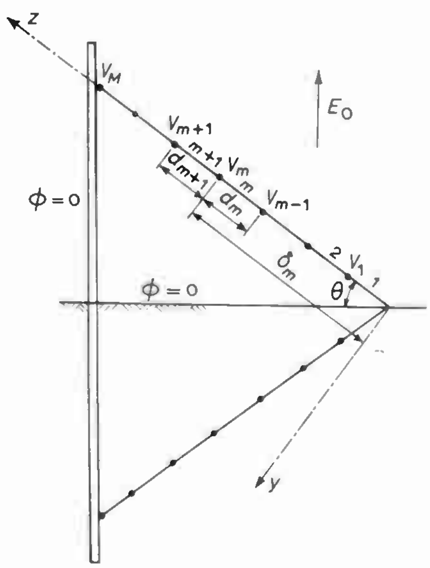

Fig. 2. A sectionalized antenna guy with its electrical image.

where

$$d = \sum_{m=1}^{M} d_m \tag{22}$$

is the total length of the guy, $q_g(z')$ and $q_{gi}(t)$ are the charge densities on the guy and its image, respectively. For $t = z'$

$$q_{gi}(t) = -q_g(z'). \tag{23}$$

At a point $P(x = 0, y = b, z)$ on the surface of the guy, equation (19) can be put in the form:

$$\phi = \phi_0(P)+\phi_i(P) + \frac{1}{4\pi\varepsilon_0}\int_0^d \frac{q_g(z')\,dz'}{\sqrt{b^2+(z-z')^2}} - \frac{1}{4\pi\varepsilon_0}\int_0^d \frac{q_g(z')\,dz'}{\sqrt{b^2+z^2+z'^2-2zz'\cos 2\theta}}. \tag{24}$$

Since the total charge on a section of the guy must be equal to zero (except for the section $m = 1$, which is directly connected to the ground), the function $q_g(z')$ is discontinuous and as a rule changes the algebraic sign at points where the guy is broken by insulators. For that reason the function $q_g(z')$ should be replaced by M functions $q_m(z')$, each valid for a separate section.

According to the boundary condition for potential, the potential ϕ defined by equation (24) must have a particular constant value along each section:

$$\phi = \begin{cases} \Phi_1 & \text{for} \quad 0 < z < \delta_1 = d_1 \\ \Phi_2 & \text{for} \quad \delta_1 < z < \delta_2 \\ \vdots & \\ \Phi_m & \text{for} \quad \delta_{m-1} < z < \delta_m \\ \vdots & \\ \Phi_M & \text{for} \quad \delta_{M-1} < z < \delta_M \end{cases} \tag{25}$$

where

$$\delta_m = \sum_{i=1}^{m} d_i. \tag{26}$$

The $M-1$ unknown potentials, from Φ_2 through Φ_M, are to be determined. Once the potentials Φ_m have been known, the voltages V_m on the insulators can be calculated by using the formula

$$V_m = \Phi_m - \Phi_{m+1}. \tag{27}$$

Since the potential of the tower is equal to zero, $V_m = \Phi_M$.

In order to calculate the unknown potentials Φ_m, let us approximate the charge distribution functions $q_m(z')$ by polynomials of the N-th order:

$$q_m(z') = \sum_{n=0}^{N} A_{mn}\left[\frac{z'-\delta_m+d_m}{d_m}\right]^n, \quad m = 1,2,\ldots M. \tag{28}$$

Since $q_1(0) = 0$, the constant term on the first section ($m = 1$) is

$$A_{10} = 0. \tag{29}$$

In addition to the distributed charge $q_m(z)$, there are the lumped charges Q_m and $-Q_{m-1}$ on the ends of each section. These charges are localized on the metal holders of the insulators and have the values

$$\begin{aligned} Q_m &= CV_m = C(\Phi_m - \Phi_{m+1}) \\ Q_{m-1} &= CV_{m-1} = C(\Phi_{m-1} - \Phi_m) \end{aligned} \tag{30}$$

where C is the insulator self-capacitance in farads.

Since the total charge on each section ($m \neq 1$) must be equal to zero, the charges Q_m, $-Q_{m-1}$ and $q_m(z')$ satisfy the condition

$$\int_{\delta_m-d_m}^{\delta_m} q_m(z')\,dz' + Q_m - Q_{m-1} = 0, \tag{31}$$

$$\text{i.e.} \sum_{n=0}^{N} A_{mn}d_m/(n+1) + C(2\Phi_m - \Phi_{m+1} - \Phi_{m-1}) = 0, \quad m = 2,3,\ldots,M. \tag{32}$$

With equations (28) and (25), equation (24) gives a system of M integral equations:

$$\Phi_m = \phi_0(z)+\phi_i(z) + \frac{1}{4\pi\varepsilon_0} \times \sum_{m=1}^{M}\sum_{n=0}^{N}\left\{\int_{\delta_m-d_m}^{\delta_m} \frac{A_{mn}\{(z'-\delta_m+d_m)/d_m\}^n dz'}{\sqrt{b^2+(z-z')^2}} - \int_{\delta_m-d_m}^{\delta_m} \frac{A_{mn}\{(z'-\delta_m+d_m)/d_m\}^n dz'}{\sqrt{b^2+z^2+z'^2-2zz'\cos 2\theta}}\right\}$$

$$\delta_m - d_m < z < \delta_m, \qquad m = 1,2,\ldots,M. \tag{33}$$

If we introduce the normalized co-ordinates, defined by

$$u' = (z'-\delta_m+d_m)/d_m \qquad u = z/d, \tag{34}$$

the equation (33) can be put in a simpler form:

$$\Phi_m = \phi_0(ud)+\phi_i(ud) + \frac{1}{4\pi\varepsilon_0 d} \times \sum_{m=1}^{M} d_m \sum_{n=0}^{N} A_{mn}\left[\int_0^1 \frac{u'^n\,du'}{\sqrt{\rho_1}} - \int_0^1 \frac{u'^n\,du'}{\sqrt{\rho_2}}\right]$$

$$(\delta_m - d_m)/d < u < \delta_m/d, \quad m = 1,2,\ldots,M, \tag{35}$$

where

$$\rho_1 = (b/d)^2 + \{u - u'd_m/d - (\delta_m - d_m)/d\}^2 \tag{36}$$

$$\rho_2 = (b/d)^2 + u^2 + \{(u'd_m+\delta_m-d_m)/d\}^2 - 2u\{(u'd_m+\delta_m-d_m)/d\}^2 \cos 2\theta. \tag{37}$$

The integrals of the type

$$J_n = \int_0^1 \frac{u'^n\,du'}{\sqrt{\rho}}, \qquad \rho = \alpha + \beta u' + \gamma u'^2, \tag{38}$$

appearing in equation (35), can be calculated by the following formulae:

$$J_n = \begin{cases} \dfrac{1}{\sqrt{\gamma}} \ln \dfrac{2\sqrt{\gamma\rho(1)}+2\gamma+\beta}{2\sqrt{\gamma\rho(0)}+\beta} & \text{for } n = 0 \\ \dfrac{\sqrt{\rho(1)}-\sqrt{\rho(0)}-0{\cdot}5\,\beta J_0}{\gamma} & \text{for } n = 1 \\ \dfrac{\sqrt{\rho(1)}-0{\cdot}5(2n-1)\beta J_{n-1}-(n-1)\alpha J_{n-2}}{n\gamma} & \text{for } n \geqslant 2 \end{cases} \tag{39}$$

where

$$\gamma > 0 \qquad \rho(0) = \alpha \qquad \rho(1) = \alpha+\beta+\gamma.$$

The following unknown constants will now be determined:

$M-1$ potentials Φ_m on sections $m = 2, 3, \ldots, M$ ($\Phi_1 = 0$);

$M(N+1)$ coefficients A_{mn}.

Since the coefficients A_{mn} are subject to M conditions defined by equations (29) and (32), the remaining

Table 1
The ratio E_z/E_r for different values of N

z/h	$N = 2$	$N = 3$	$N = 5$	$N = 10$	$N = 15$
0·1	1·368 × 10^{-4}	3·110 × 10^{-4}	3·992 × 10^{-4}	6·768 × 10^{-5}	1·636 × 10^{-6}
0·2	5·685 × 10^{-4}	3·566 × 10^{-4}	1·263 × 10^{-4}	6·956 × 10^{-6}	3·151 × 10^{-7}
0·3	3·387 × 10^{-4}	6·402 × 10^{-6}	7·996 × 10^{-5}	1·401 × 10^{-6}	1·794 × 10^{-7}
0·4	5·590 × 10^{-5}	1·542 × 10^{-4}	2·601 × 10^{-5}	4·431 × 10^{-7}	2·472 × 10^{-7}
0·5	3·571 × 10^{-4}	3·776 × 10^{-5}	7·123 × 10^{-5}	1·901 × 10^{-7}	9·197 × 10^{-8}
0·6	3·460 × 10^{-4}	2·131 × 10^{-4}	2·743 × 10^{-6}	8·866 × 10^{-8}	6·664 × 10^{-8}
0·7	3·266 × 10^{-4}	2·261 × 10^{-4}	2·142 × 10^{-4}	1·812 × 10^{-6}	8·031 × 10^{-7}
0·8	2·551 × 10^{-3}	1·006 × 10^{-3}	1·848 × 10^{-4}	1·613 × 10^{-5}	9·266 × 10^{-6}
0·9	1·044 × 10^{-2}	7·838 × 10^{-3}	5·619 × 10^{-3}	3·465 × 10^{-4}	1·943 × 10^{-4}

$M(N+1)-1$ necessary equations are obtained by satisfying equations (35) in $M(N+1)-1$ points along the sections. On the first section we select N matching points and on all other sections $N+1$ points each. These points can be chosen arbitrarily, but none at the ends of the sections. If the selected points are equidistant on a section, their co-ordinates z_i are determined by

$$z_i = id_1/(N+1) \quad i = 1, 2, \ldots, N \quad \text{for } m = 1 \tag{40}$$

$$z_i = \delta_{m-1} + id_m/(N+2) \qquad i = 1, 2, \ldots, N+1 \quad \text{for } m = 2, \ldots, M. \tag{41}$$

Dividing equations (40) and (41) by d we obtain the normalized co-ordinates u_i.

By solving $(M-1)+M(N+1)$ linear equations, the unknowns Φ_m and A_{mn} can be calculated. The voltages V_m are obtained by equation (27).

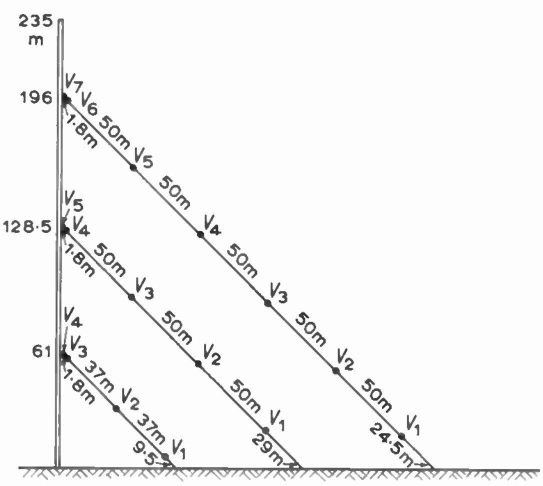

Fig. 3. Arrangement of the guy sections of the antenna tower of the main m.f. transmitter of Radio Beograd.

4 Numerical Example and Conclusion

In order to illustrate the theory and to draw some practical conclusions, the charge densities and static guy-insulator voltages will be calculated for the antenna of the main m.f. transmitter of Radio Beograd. This is a guyed steel tower with triangular cross-section, having a height of 235 m and an equivalent radius a of 0·7 m. The tower is guyed by 3 sets of 3 guy wires attached to the tower at three different heights and inclined at an angle of 45° with respect to the ground plane. In horizontal projection the guy-wires are laid out 120° apart. The radius of wires b is 14 mm. The arrangement of the guy sections in the vertical plane as well as the necessary notation are shown in Fig. 3. The insulator capacitance is about 50 pF.

Applying the method described in Section 2, the potential in the tower vicinity was first calculated. For checking the convergence and accuracy of results with respect to the order of the polynomial used, the fulfilment of the boundary condition on the antenna cylinder surface was verified. The ratio of the tangential field component, E_z, and radial component, E_r, seems to be very suitable criterion for that. That is why the ratio E_z/E_r was calculated for different values of N and it is presented in Table 1. It is surprising that even a polynomial of the order as low as $N = 2$ fulfils the boundary condition very well. The equipotential lines, relevant to an unperturbed field strength $E_0 = 100$ V/m and $N = 5$, are shown in Fig. 4.

The charge density per unit length of the tower for $E_0 = 100$ V/m and for different values of N is presented in Table 2.

Table 2
The charge density $q(z)$ along the tower in 10^{-9} C/m
($E_0 = 100$ V/m, $a/h = 0{\cdot}002978$)

z/h	0·0	0·1	0·2	0·3	0·4	0·5	0·6	0·7	0·8	0·9	1·0
$N = 2$	0	23	46	71	96	122	149	176	204	234	263
$N = 5$	0	23	47	71	95	120	147	175	209	251	307
$N = 20$	0	23	47	71	95	120	146	174	207	250	407

The results for $z/h = 1$ should be disregarded because the top of the tower is a point of discontinuity.

The voltages on guy insulators, for $E_0 = 100$ V/m, $C = 50$ pF and for the second- and fifth-order polynomial approximation of the charge density on guy sections, are shown in Table 3. (The charge density on the tower is approximated by a polynomial of the fifth-order.)

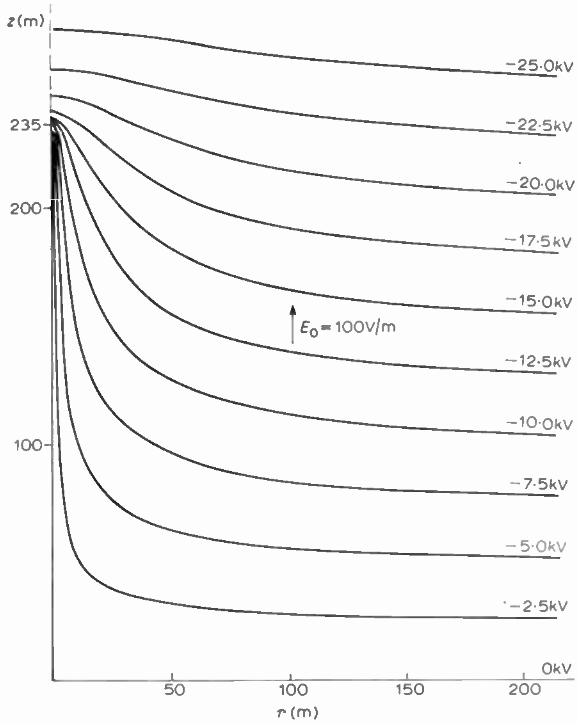

Fig. 4. Equipotential lines in the vicinity of the antenna cylinder; $h = 235$ m, $a/h = 0{\cdot}002978$ and $E_0 = 100$ V/m.

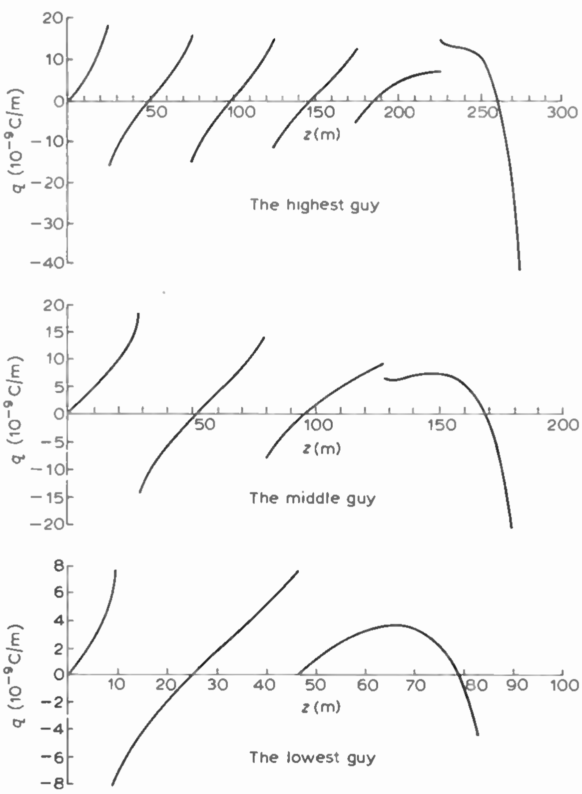

Fig. 5. Charge density per unit length along the guys from Fig. 3; $E_0 = 100$ V/mn, $C = 50$ pF and $N = 5$.

Table 3

Voltages on guy-insulators, V_m, in kV ($E_0 = 100$ V/m, $C = 50$ pF)

	V_1	V_2	V_3	V_4	V_5	V_6	V_7
			THE HIGHEST GUY				
$N = 2$	3·225	3·060	2·682	1·790	1·077	5·500	4·181
$N = 5$	3·223	3·057	2·675	1·768	1·144	5·398	4·181
			THE MIDDLE GUY				
$N = 2$	3·092	2·235	0·161	2·994	2·495		
$N = 5$	3·084	2·220	0·120	2·930	2·495		
			THE LOWEST GUY				
$N = 2$	1·271	0·636	0·768	1·139			
$N = 5$	1·263	0·619	0·744	1·139			

In order to estimate the effects of the insulator capacitance, the same voltages were calculated for $C = 0$. These values are given in Table 4.

Table 4

Voltages on guy-insulators, V_m, in kV ($E_0 = 100$ V/m, $C = 0$)

	V_1	V_2	V_3	V_4	V_5	V_6	V_7
			THE HIGHEST GUY				
$N = 2$	3·268	3·108	2·784	2·102	0·385	6·696	4·181
$N = 5$	3·259	3·106	2·780	2·087	0·443	6·608	4·181
approx.	3·272	3·122	2·811	2·184	0·139	7·343	4·181
			THE MIDDLE GUY				
$N = 2$	3·245	2·435	0·597	3·782	2·495		
$N = 5$	3·231	2·426	0·562	3·725	2·495		
approx.	3·282	2·493	0·933	4·213	2·495		
			THE LOWEST GUY				
$N = 2$	1·409	0·956	1·226	1·139			
$N = 5$	1·400	0·939	1·200	1·139			
approx.	1·445	1·121	1·427	1·139			

In addition to the values of voltages corresponding to $N = 2$ and $N = 5$, the third approximate value ('approx.') of the voltage is given in Table 4. This value is defined as the difference between the potentials in the mid-points of two adjacent sections in the physical absence of the guys. These potentials are those calculated in Section 2, i.e. $\phi_0 + \phi_i$. It is interesting to note a very good overall agreement between these approximate values and those calculated by the present method for $C = 0$, except for the minimum values of the voltages.

The curves in Fig. 5 represent the charge density distribution functions along the three guys from Fig. 3, corresponding to $E_0 = 100$ V/m, $C = 50$ pF and $N = 5$. It is to be noted that the maximum value of the charge density on the guys is at least an order of magnitude smaller than the average charge density on the tower. This justifies the assumption that the guys do not affect appreciably the charge distribution along the tower.

In the foregoing examples all the voltages and charge densities are calculated under the assumption of a relatively low field strength $E_0 = 100$ V/m, which is a representative value for calm weather conditions. However, immediately before and during thunderstorms this value can be exceeded by ten or even a hundred times. According to Müller-Hillebrand[5], maximal field strength of 3–5 kV/m is typical, and very rarely exceeds 10 kV/m. Similar data can be found in the paper by Simpson and Scrase,[6] who quote values of 5–10 kV/m.

In order to point out the significance of the static voltages, we shall compare these voltages, corresponding to a moderate static field strength $E_0 = 1$ kV/m, with the measured r.f. voltages for the same antenna. The results are shown in Table 5. The r.f. voltages correspond to an unmodulated r.f. power of 400 kW.

Table 5

Static and r.f. voltages on guy insulators, in kV ($E_0 = 1$ kV/m, $C = 50$ pF; r.f. power 400 kW)

	V_1	V_2	V_3	V_4	V_5	V_6+V_7
			THE HIGHEST GUY			
Static	32·23	30·57	26·75	17·68	11·44	95·79
r.f.	0·5	1·2	1·3	1·4	2·0	9·8
			THE MIDDLE GUY			
	V_1	V_2	V_3	V_4+V_5		
Static	30·84	22·20	1·20	54·25		
r.f.	1·8	2·0	1·9	3·3		
			THE LOWEST GUY			
	V_1	V_2	V_3+V_4			
Static	12·63	6·19	18·83			
r.f.	1·8	4·9	11·8			

The method explained in this paper can, of course, be applied to antenna towers consisting of several sections of different, but uniform cross-section.

5 Acknowledgment

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7 Appendix

The cylindrical components of the field strength in the vicinity of the tower (the guys are ignored) can be calculated by means of the following formulae:

$$E_z = E_0 + \frac{1}{4\pi\varepsilon_0 h^2}\sum_{n=1}^{N} B_n[nP_{n-1}(u,R)+nP_{n-1}(-u,R)- -\{R^2+(u-1)^2\}^{-\frac{1}{2}}-\{R^2+(u+1)^2\}^{-\frac{1}{2}}], \quad (42)$$

$$E_r = \frac{1}{4\pi\varepsilon_0 h^2}\sum_{n=1}^{N} B_n\{T_n(u,R)-T_n(-u,R)\}, \quad (43)$$

where

$$T_n(u,R) = \int_0^l \frac{u'^n\,du'}{\{R^2+(u-u')^2\}^{\frac{3}{2}}}$$

$$= \frac{1-u}{R^2\sqrt{R^2+(u-1)^2}} + \frac{u}{R^2\sqrt{R^2+u^2}} \qquad \text{for } n=0$$

$$= \frac{\sqrt{R^2+u^2}}{R^2} - \frac{(R^2+u^2-u)}{R^2\sqrt{R^2+(u+1)^2}} \qquad \text{for } n=1$$

$$= \frac{u\sqrt{R^2+u^2}}{R^2} - \frac{\{R^2-u^2+(R^2+u^2)u\}}{R^2\sqrt{R^2+(u-1)^2}} + P_0(u,R) \qquad \text{for } n=2$$

$$= \frac{[\{R^2+(u-1)^2\}^{-\frac{1}{2}}+u(2n-3)T_{n-1}(u,R)- -(n-1)(R^2+u^2)T_{n-2}(u,R)]}{n-2} \qquad \text{for } n>3 \quad (44)$$

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The theory of coupling in a tapered waveguide

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SUMMARY

A theory is given which enables the overall coupling between two modes to be calculated for a tapered waveguide, if the coupling factor between the modes, the phase constants of the modes, and the characteristic impedances of the waveguide for the two modes are known at all points along the taper. It is thus restricted to those cases where a characteristic impedance can be defined, i.e. waveguides bounded by a very good conductor and not containing very lossy materials. The theory is found to be valid even if one of the modes is cut off over part of the length of the taper. 'Taper', in the present context, includes the case of uniform dimensions and gradually varying properties of one or more of the materials contained in the guide. The overall coupling between the H_{01} and H_{02} modes of a copper-walled circular-waveguide taper of raised-cosine form is calculated as an example.

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List of Principal Symbols

$a = a(z)$	radius of a circular taper at distance z from the input end
a_0, a_l	values of a at $z = 0$, $z = l$, respectively
$b = b(z)$	dimension or other parameter which varies progressively along the taper
z	axial dimension, measured from $z = 0$ at the beginning of the taper
l	length of taper
λ_0	free-space wavelength at the working frequency
c	velocity of light in free space
f	working frequency
f_0	cut-off frequency for a given value of a
χ_n	characteristic number for the nth mode
Z_0	characteristic impedance
Z_{0n}	characteristic impedance of the nth mode
Z	impedance related to Z_0 by equation (18)
$U_n(z)$	amplitude of the nth mode at distance z from the input end of the taper
$\beta_n(z)$	phase constant of the nth mode at distance z from the input end of the taper
$\bar{\beta}_n$	normalized phase constant, $= \lambda_0 \beta_n/2\pi$
$[T]$	matrix relating output amplitudes in the pth and qth modes to the input amplitudes in these modes; the elements of $[T]$ are given as equations (5)
$\Gamma(z)$	coupling factor between the pth and qth modes at distance z from the input end of the taper
$Y(z)$	see equation (6)
Γ_0, Y_0 and Γ_l, Y_l	values of Γ and Y at $z = 0$ and $z = l$ respectively
ψ	see equation (7)
E	see equation (4)
J	see equation (8)
I	is a quantity obtained from J by replacing β_q by $-\beta_q$

1 Introduction

Tapers are in common use as a means of changing the dimensions of a waveguide. In the case of a guide whose cross-section is such that only one mode is supported, a theory exists (see Refs. 1 and 2, pp. 478–83) for calculating the overall reflexion in that mode and for designing tapers in which the overall reflexion is minimized. There appears, however, to be no comparable theory for the coupling into other modes when a multi-mode waveguide is tapered. In this paper a theory will be developed and applied to an example.

The theory has two main parts. The first is to calculate the coupling coefficient at a given point in the taper, in terms of the geometry of the taper, and the second is to use the coupling coefficient in coupled-mode theory to determine the overall coupling. The coupling coefficient can be dealt with as a development of the previous reflexion theory for a single-mode system. The coupled-mode theory is straightforward if both modes are above cut-off throughout the length of the taper, but if over part of the length of the taper one mode is cut off at frequencies of interest, some interesting features arise. These will be discussed later in connexion with a specific example.

The example to be considered is that of H_{01}- to H_{02}-mode coupling in a guide of circular cross-section whose radius, a, follows the law:

$$a = \tfrac{1}{2}\{(a_0+al)+(a_0-al)\cos(\pi z/l)\} \tag{1}$$

This gives a smooth taper between the radii a_0 at $z = 0$ and al at $z = l$, z being the axial dimension of the waveguide. It has the disadvantage of containing only one adjustable parameter—the length l—and this may be subject to constraints other than the desired coupling level. This taper is of interest in connexion with long-distance waveguide transmission systems. When it is necessary to make sharp changes in direction, the loss due to coupling into other modes can be substantially reduced by reducing the cross-sectional diameter—and hence the number of unwanted modes into which coupling can take place—in a bend of small radius of curvature. The reduction of the guide diameter is effected by means of a taper in a waveguide with smooth copper walls, and coupling may be expected to take place between the H_{01}, H_{02}, H_{03}, H_{04}, ..., modes.

2 Coupled-mode Theory of Waveguide Taper

In most practical cases, it is desired that the energy be confined to a single mode. Coupling to other modes occurs as a defect of the system, and is usually small. The power levels in the unwanted modes are then small compared with that in the wanted mode, and as far as coupling between the wanted and a particular unwanted mode are concerned, the other unwanted modes can be ignored. Thus for most practical purposes, a multimode waveguide can be considered as a two-mode system and coupling with each unwanted mode can be considered separately.

For a two-mode coupled-mode system the output amplitudes, $U_p(l)$, $U_q(l)$, at $z = l$ are obtained as functions of the input amplitudes $U_p(0)$, $U_q(0)$, at $z = 0$ by solving the coupled-mode equations

$$\left.\begin{aligned} \frac{\partial U_p}{\partial z} &= -\mathrm{j}\beta_p U_p - \Gamma^* U_q \\ \frac{\partial U_q}{\partial z} &= -\mathrm{j}\beta_q U_q + \Gamma U_p \end{aligned}\right\} \tag{2}$$

where Γ^* is the complex conjugate of Γ and the waves in modes p and q are assumed to be travelling in the same direction. If they are travelling in opposite directions, Γ^* and β_q must be replaced by $-\Gamma^*$ and $-\beta_q$, and may need redefining. The solution of equations (2) is given in Ref. 3 and on pages 472–5 of Ref. 2. It is

$$\begin{bmatrix} U_p(l) \\ U_q(l) \end{bmatrix} = E\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix}\begin{bmatrix} U_p(0) \\ U_q(0) \end{bmatrix} = E[T]\begin{bmatrix} U_p(0) \\ U_q(0) \end{bmatrix} \tag{3}$$

where

$$E = \exp\left\{\frac{-\mathrm{j}}{2}\int_0^l (\beta_p+\beta_q)\,\mathrm{d}z\right\} \tag{4}$$

$$\left.\begin{aligned} T_{11} &= (1/\psi)\{Y_0 Y_l\, \mathrm{e}^{\mathrm{j}J} + \Gamma_0 \Gamma_l^* \mathrm{e}^{-\mathrm{j}J}\} \\ T_{12} &= -(\mathrm{j}/\psi)\{\Gamma_0^* Y_l\, \mathrm{e}^{\mathrm{j}J} - \Gamma_l^* Y_0\, \mathrm{e}^{-\mathrm{j}J}\} \\ T_{21} &= (\mathrm{j}/\psi)\{\Gamma_l Y_0\, \mathrm{e}^{\mathrm{j}J} - \Gamma_0 Y_l\, \mathrm{e}^{-\mathrm{j}J}\} \\ T_{22} &= (1/\psi)\{\Gamma_l \Gamma_0^* \mathrm{e}^{\mathrm{j}J} + Y_0 Y_l\, \mathrm{e}^{-\mathrm{j}J}\} \end{aligned}\right\} \tag{5}$$

$$Y = Y(z) = \tfrac{1}{2}(\beta_q-\beta_p)+\sqrt{\Gamma\Gamma^*+(\beta_q-\beta_p)^2/4} \tag{6}$$

$$\psi = \sqrt{(Y_l^2+\Gamma_l\Gamma_l^*)(Y_0^2+\Gamma_0\Gamma_0^*)} \tag{7}$$

Γ_0, Y_0 and Γ_l, Y_l, indicate the values of Γ, Y, at $z = 0$ and at $z = l$ respectively.

$$J = \int_0^l [\Gamma\Gamma^*+(\beta_q-\beta_p)^2/4]^{\frac{1}{2}}\,\mathrm{d}z \tag{8}$$

If the input to the taper consists only of mode p, i.e. $U_q(0) = 0$, then for coupling between waves travelling in the same direction

$$U_q(l) = ET_{21}U_p(0) \tag{9}$$

For the case of a wave in mode q in the reverse direction, the output is $U_q(0)$, with $U_q(l) = 0$. Hence

$$U_q(0) = -(T_{21}/T_{22})\,U_p(0) \tag{10}$$

In equation (10), Γ^* and β_q are to be replaced by $-\Gamma^*$ and $-\beta_q$, and the value of Γ itself will need consideration.

It may be assumed that the wanted mode, p, is always above cut-off. This may not be the case with the qth mode, however. For a mode well above cut-off, β may be taken as a real quantity if the guide does not contain very lossy materials; in particular, for a metal-walled guide containing a single homogeneous isotropic low-loss non-conducting material of relative permeability μ and dielectric constant ε, the value of β appropriate to an h.s.p.† approximation can be used. Hence

$$\bar{\beta} = \sqrt{\varepsilon\mu-\chi^2c^2/4\pi^2f^2b^2} \tag{11}$$

where ε and μ are taken as real, c is the velocity of light in free space, f is the working frequency, χ is a constant for the mode concerned, and b is the varying cross-sectional dimension. Well below cut-off, too, the h.s.p. value of β or $\bar{\beta}$ may be used; this again is given by equation (11), and $\bar{\beta}$ is imaginary. Near cut-off, the value of Γ depends importantly on β_q, which must therefore be calculated accurately; the h.s.p. value is no longer good enough. If, at cut-off, β_q vanished, Γ would become infinite and the coupled-mode theory would break down. But in fact β_q is always finite in the neighbourhood of cut-off, and a finite value of Γ is obtained. Whether or not the coupled-mode theory remains valid will be considered in the next Section.

For a metal-walled circular waveguide containing only vacuum, the phase constant of an H_{0n} mode in the neighbourhood of cut-off is given approximately by

$$\beta^2 = 2\frac{\delta f}{f_0}+\frac{2}{\chi}\sqrt{\frac{\pi\varepsilon f_0}{\sigma}}(1-\mathrm{j})\left[1+\tfrac{1}{2}\frac{\delta f}{f_0}\right] \tag{12}$$

† A homogeneous waveguide is one in which the wave energy propagates in a single homogeneous medium. A simple guide is one in which all the materials have scalar properties (no ferrites, plasmas, single crystals). A perfect guide is one in which all the materials are lossless. A homogeneous, simple perfect, or h.s.p., guide is thus one consisting of a single region in which the wave energy travels, the material in this region being lossless and scalar. The rest of space is occupied by perfect conductor.

where δf is the departure of the working frequency f from the cut-off frequency f_0, being positive or negative according as $f > f_0$ or $f < f_0$, and σ is the conductivity of the metal waveguide wall. For copper, $\sigma = 5{\cdot}8\times10^7$ S/m, and if $f, f_0, \delta f$, are in GHz, equation (12) becomes

$$\beta^2 = \left\{\frac{2\delta f}{f_0} + \frac{4{\cdot}377}{\chi}\times10^{-5}\sqrt{f_0}\right\} - \mathrm{j}\frac{4{\cdot}377\times10^{-5}}{\chi}\sqrt{f_0} \quad (13)$$

In evaluating J and E, the contributions to the integrals from β_q in the neighbourhood of the value of z where a is at cut-off for the working frequency will be minute. Therefore the h.s.p. values can be used throughout. But when the working frequency is such that either end of the taper is close to the cut-off of the qth mode, β_q must be calculated more accurately in evaluating the quantities Γ_0, Y_0, or Γ_l, Y_l occurring in T_{21} and T_{12}, and in particular, for the H_{0n} modes of circular waveguide equation (13) must be used.

3 Validity of the Coupled-mode Theory

In Ref. 3 the conditions are given for the validity of the coupled-mode theory. They are

$$\left.\begin{aligned} &\left|\frac{\partial}{\partial z}(\beta_p-\beta_q)\right| \ll \left|(\beta_p-\beta_q)^2\right| \\ \text{and}\quad &\left|\frac{\partial\Gamma}{\partial z}\right| \ll \left|\Gamma(\beta_p-\beta_q)\right| \end{aligned}\right\} \quad (14)$$

Usually these conditions will be satisfied if the dimensions of the waveguide vary sufficiently slowly. But if $\beta_q \simeq \beta_p$, or if either $\beta_p \simeq 0$ or $\beta_q \simeq 0$, special consideration is necessary.

It must be noted first that β_p, β_q, or $(\beta_p-\beta_q)$ can never be exactly zero. It was pointed out in the last section that when the losses of a metal-walled guide are properly taken into account, β remains finite at all frequencies. For an h.s.p. guide, two modes may be degenerate—e.g. the H_{01} and E_{11} modes of circular guide—but when for a practical guide the losses are properly taken into account, the degeneracy disappears. Nevertheless, the zeros are replaced by very minute complex quantities, and it is not obvious, in general, whether the conditions will be satisfied or not. This question must be answered before we can say whether or not the theory of this paper is applicable to tapers with degenerate modes, or to tapers in which one of the modes becomes cut off somewhere in the length of the taper.

The case of degeneracy will not be considered further in this paper. Each case must be treated individually as it arises.

The coupling factor always contains a factor $\sqrt{\bar\beta_p\bar\beta_q}$ in the denominator, and so becomes large when either mode is near cut-off. $\mathrm{d}\Gamma/\mathrm{d}z$ depends almost entirely on $\mathrm{d}\bar\beta_q/\mathrm{d}z$, if it is the qth mode that is near cut-off, and we then obtain

$$\frac{\mathrm{d}\Gamma/\mathrm{d}z}{\Gamma} \simeq \frac{-\mathrm{d}\bar\beta_q/\mathrm{d}z}{2\bar\beta_q}$$

The second of the conditions (14) then becomes

$$\left|\frac{\mathrm{d}\bar\beta_q/\mathrm{d}z}{\bar\beta_q}\right| \ll \left|2\bar\beta_p\right| \quad (15)$$

The first condition may be written

$$\left|\frac{\mathrm{d}\bar\beta_p/\mathrm{d}z}{\bar\beta}\right| \ll \left|\bar\beta_p\right| \quad (16)$$

Clearly, as long as $\mathrm{d}b/\mathrm{d}z$ is sufficiently small these conditions will be satisfied, and the coupling theory will be valid even though the taper cuts off for one of the modes.

4 Determination of the Coupling Factor

We can calculate the coupling factor for an elementary length δz of the taper from the coupling coefficient for a step in the guide cross-section. If the coupling coefficient for a step is C, the net amplitude of unwanted mode set up at a step of magnitude δb of a dimension b is $C\,\delta b/b$, which is the coupling factor for the step. In a taper, Γ is the coupling factor per unit length and $\Gamma\delta z$ is the coupling factor for the change in the guide dimension which takes place in the elementary length δz. We can thus write

$$\Gamma\,\delta z \simeq C\frac{\delta b}{b} = C\frac{1}{b}\frac{\mathrm{d}b}{\mathrm{d}z}\delta z$$

and so

$$\Gamma \simeq Cb'/b \quad (17)$$

where the dash denotes differentiation with respect to z.

This first-order approximation to Γ can be used whenever b' is not zero, but when b' vanishes it is not correct to take Γ as zero. At any point, the value of Γ will depend on the first non-vanishing derivative of b. At the ends of a smooth taper, where $b' = 0$, the values of Γ_0 and Γ_l must be calculated to a higher order to evaluate T_{21} and T_{12} and so to determine the overall coupling.

The general method is to express Cb'/b in terms of $\bar\beta_p$, $\bar\beta_q$, Z_{0p}, and Z_{0q}, where Z_{0p} and Z_{0q} are the characteristic impedances for the pth and qth modes. By the characteristic impedance at a point in a taper is meant the characteristic impedance of a uniform waveguide of the same cross-section as the taper at that point, and for this the definition of chacteristic impedance given in Ref. 4 should be used. When b' is zero, however, Z'_{0p} and Z'_{0q} are not to be taken as zero. Instead, the following procedure is to be adopted.

It is shown in Refs. 1 and 2, (pp. 478–83) that when a waveguide is tapered the characteristic impedance which must be used in calculations of the reflexion coefficient of a single mode is not Z_0 but an effective value Z, given by

$$Z = Z_0 - \frac{\mathrm{j}}{2\beta}Z'_0 - \frac{1}{4\beta^2}\left[Z''_0 - \frac{Z_0'^2}{2Z_0} - \frac{Z'_0\beta'}{\beta}\right] + \ldots \quad (18)$$

When $Z'_0, Z''_0, \ldots Z_0^{n-1}$ all vanish, this becomes

$$Z = Z_0 + \left(\frac{-\mathrm{j}}{2\beta}\right)^n Z_0^n + \text{higher terms}$$

and then

$$Z' = \left(\frac{-\mathrm{j}}{2\beta}\right)^{n-1} Z_0^n \quad (19)$$

where $Z'_0, Z''_0, \ldots Z_0^{n-1}$ are put equal to 0 *after* differentiating. Z' is to be used instead of Z'_0 in evaluating

Cb'/b. Thus a non-zero value is always found for Γ, unless all the derivatives vanish in which case the guide is uniform.

The procedure will become clearer after the treatment of a specific example in the next Section.

5 H_{op}- to H_{oq}-mode Coupling in Tapered Circular Guide

5.1 *General Theory*

The coupling coefficient for a step in radius is (see Refs. 5, 6 and 3, p. 459)

$$C = \frac{\sqrt{(1-\bar{\beta}_p^2)(1-\bar{\beta}_q^2)}}{(\beta_q-\beta_p)\sqrt{\bar{\beta}_p\bar{\beta}_q}} \tag{20}$$

Hence

$$\Gamma = \frac{\sqrt{(1-\bar{\beta}_p^2)(1-\bar{\beta}_q^2)}}{(\beta_q-\beta_p)\sqrt{\bar{\beta}_p\bar{\beta}_q}}\,\frac{a'}{a} \tag{21}$$

if $a' \neq 0$. We note that for the H_{0n} modes

$$\frac{1-\bar{\beta}_n^2}{\beta_n}\,\frac{a'}{a} = \frac{Z'_{0n}\bar{\beta}_n^2}{-W\chi_n^{\;2}} \tag{22}$$

where $\chi_n = 3{\cdot}83171$, $7{\cdot}01559$, $10{\cdot}17347$, $13{\cdot}32369$, etc., for $n = 1, 2, 3, 4$, etc. The fundamental definition of characteristic impedance is given in Ref. 4, and this has been used in arriving at the expression (22). W is a dimensionless constant, the same for all the H_{0n} modes; it happens to be infinite, but as it will eventually cancel out no difficulty arises. Using equation (22), equation (21) can be replaced by

$$\Gamma = \frac{1}{(\beta_q-\beta_p)}\,\frac{\bar{\beta}_p\bar{\beta}_q}{W\chi_p\chi_q}\sqrt{(-Z'_{0p})(-Z'_{0q})} \tag{23}$$

If Γ is not sufficiently small to be neglected in the expression (8) for J, equation (21) or (23) can be used. Otherwise, if a' vanishes at the ends of the taper, the values of Γ given by equation (23) are zero at the ends of the taper, and this expression cannot, therefore, be used to calculate $U_q(l)$ or $U_q(0)$ from equations (9) and (10). We then replace Z'_{0p} and Z_{0q} by Z'_p and Z'_q as in equation (19), and equation (23) is replaced by

$$\Gamma = \left(\frac{-\mathrm{j}}{2\sqrt{\beta_p\beta_q}}\right)^{n-1}\frac{1}{(\beta_q-\beta_p)}\sqrt{\frac{(1-\bar{\beta}_p^2)(1-\bar{\beta}_q^2)}{\bar{\beta}_p\bar{\beta}_q}}\,\frac{\mathrm{d}^n a/\mathrm{d}z^n}{a} \tag{24}$$

This expression can be used to evaluate Γ_0 and Γ_l, and hence the elements of $[T]$.

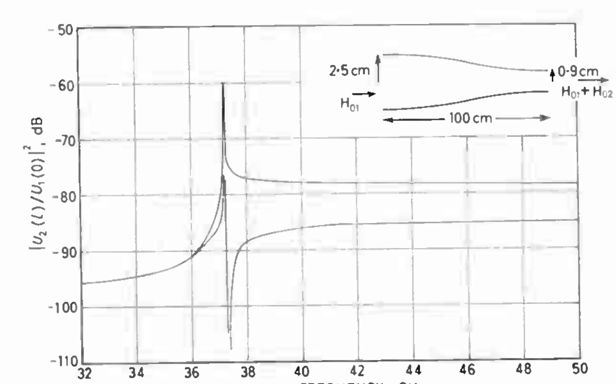

Fig. 1. Level of H_{02} mode at output (narrow end) of raised-cosine taper, for pure H_{01}-mode input at wide end.

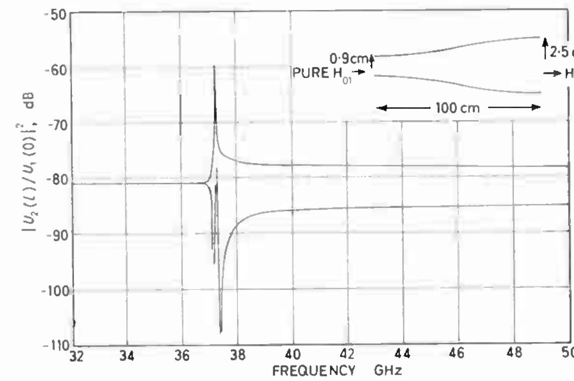

Fig. 2. Level of H_{02} mode at output (wide end) of raised-cosine taper, for pure H_{01}-mode input at narrow end.

Equation (9) now becomes

$$\left|\frac{U_q(l)}{U_p(0)}\right| = \frac{\lambda_0}{2\pi}\left|A(l)\,\mathrm{e}^{\mathrm{j}J}E - A(0)\,\mathrm{e}^{-\mathrm{j}J}E\right| \tag{25}$$

where

$$A(z) = \frac{(-\mathrm{j}/2\sqrt{\beta_q\beta_p})^{n-1}}{(\beta_p-\beta_q)^2}\sqrt{\frac{(1-\bar{\beta}_p^2)(1-\bar{\beta}_q^2)}{\bar{\beta}_p\bar{\beta}_q}}\cdot\frac{\mathrm{d}^n a/\mathrm{d}z^n}{a} \tag{26}$$

taken at $z = l$ or $z = 0$ as the case may be, n being the order of the first non-vanishing derivative of a at the end of the taper. n may have different values at the two ends.

Equation (10) becomes

$$\left|\frac{U_q(0)}{U_p(0)}\right| = \frac{\lambda_0}{2\pi}\left|B(l)\,\mathrm{e}^{2\mathrm{j}I} - B(0)\right| \tag{27}$$

where B is obtained from A, and I from J, by changing the signs of β_q and $\bar{\beta}_q$ wherever they occur.

5.2 *Raised-cosine Taper*

The raised-cosine taper has a profile of the form of equation (1), and at both ends it is the second derivative which is the first non-zero derivative, i.e. $n = 2$. Equation (25) now becomes

$$\left|\frac{U_q(l)}{U_p(0)}\right| = \frac{\chi_p\chi_q\lambda_0^4(a_0-a_l)}{64\pi^2 l^2} \times\left|\frac{E\,\mathrm{e}^{\mathrm{j}J}}{\left[\bar{\beta}_p\bar{\beta}_q(\beta_p-\beta_q)^2\right]a_l^3\Big|_{z=l}} + \frac{E\,\mathrm{e}^{-\mathrm{j}J}}{\left[\bar{\beta}_p\bar{\beta}_q(\beta_p-\beta_q)^2\right]a_0^3\Big|_{z=0}}\right| \tag{28}$$

This gives the amplitude ratio of the output H_{0q} mode, for pure H_{0p} input, when both waves travel in the same direction. If they travel in opposite directions, equation (27) gives

$$\left|\frac{U_q(0)}{U_p(0)}\right| = \frac{\chi_p\chi_q\lambda_0^4(a_l-a_0)}{64\pi^2 l^2}\times \times\left|\frac{\mathrm{e}^{2\mathrm{j}I}}{\left[\bar{\beta}_p\bar{\beta}_q(\beta_p+\beta_q)^2\right]a_l^3\Big|_{z=l}} + \frac{1}{\left[\bar{\beta}_p\bar{\beta}_q(\beta_p+\beta_q)^2\right]a_0^3\Big|_{z=0}}\right| \tag{29}$$

As an example, a raised-cosine taper of length $l = 100$ cm, with end radii 2·5 cm and 0·9 cm, is considered. Computations have been made over the range 32 GHz to 50 GHz for coupling from the H_{01} mode into the H_{02} mode, and the results are shown in Figs. 1, 2, and 3.

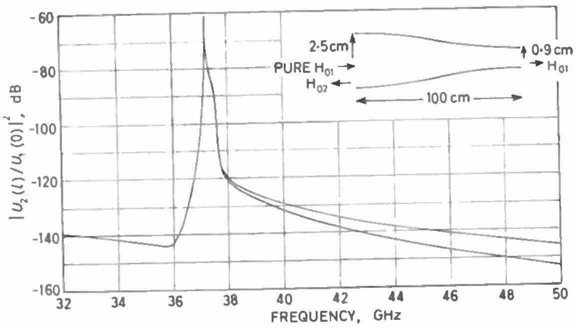

Fig. 3. Level of H_{02} mode reflected at wide end of raised-cosine taper, for pure H_{01}-mode input at wide end.

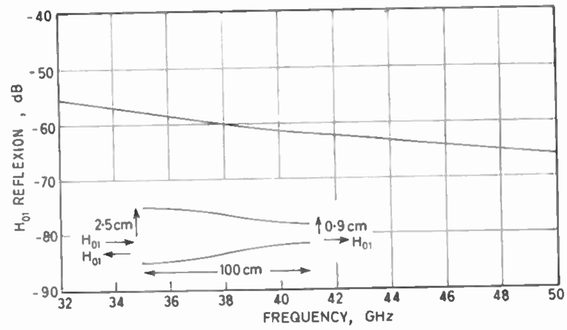

Fig. 4. H_{01}-mode reflexion from either end of raised-cosine taper.

The cut-off frequency for a radius of 0·9 cm is 37·21884 GHz and it will be seen that there are sharp peaks in the coupling curves at this frequency. Below 37·0 GHz and above 37·5 GHz, the h.s.p. values of $\bar{\beta}_1$ and $\bar{\beta}_2$ were used. Between these frequencies, the values given by equation (13) were used. The kink at 37·5 GHz in Fig. 3 is due to discrepancies between results obtained with different formulae.

Below 37·21884 GHz, the guide is cut off over a finite length, and J or I, as the case may be, develops an imaginary part giving rise to a real negative exponent. Then only one of the two terms in each of equations (28) and (29) survives, and it is this that gives the result curves in Figs. 1, 2, and 3. Above 37·21884 GHz, J and I are real and the result curves are rapid oscillations between the limits $(\lambda_0/2\pi)|A(l)\pm A(0)|$ or $(\lambda_0/2\pi)|B(l)\pm B(0)|$ as the case may be. It is these limits, forming the envelope of the true coupling curve, that are plotted in Figs. 1, 2, and 3.

Figure 1 shows the overall coupling into the H_{02} mode for a pure H_{01} mode incident at the wide end of the taper, the modes travelling in the same direction. Figure 2 shows the same thing, except that the taper is reversed, i.e. it is the narrow end that is the input end. From equation (28) it is apparent that as long as $\bar{\beta}_2$ is real, so that $Ee^{\pm jJ}$ are purely oscillating, the results are the same. However, when $\bar{\beta}_2$ becomes imaginary or complex, Ee^{-jJ} becomes minute and only the term in $1/a_l^3$ survives. The value of this differs according as a_l is taken as 2·5 or 0·9, depending on which way round the taper is. Thus Figs. 1 and 2 differ below the critical frequency of 37·21884 GHz.

Figure 3 shows the coupling into the reverse H_{02} mode for a pure H_{01} mode incident at the wide end of the taper, calculated from equation (29). As before, if the taper were reversed a difference would be expected below the critical frequency but not above it.

The sharp peaks at the critical frequency in Figs. 1, 2, and 3 can be understood as follows. If at the working frequency the cut-off occurs somewhere in the taper not near the end, the overall coupling remains small because, although the coupling factor becomes relatively large at the cut-off radius, it remains large only over a very short length of the guide. On the other hand, if the cut-off occurs at the end of the taper, the guide remains close to cut-off over a considerable length and the large coupling factor is then able to make itself felt.

Another peak in the coupling curve may be expected at a lower frequency, corresponding to the wide end being at cut-off. For the present taper this would be at 13·3988 GHz. But this peak would be much lower than those seen in Figs. 1, 2, and 3 because of the much lower general level of coupling.

It will be noticed that equations (28) and (29) contain a factor $1/l^2$ on the right-hand side. Thus the power ratio varies as $1/l^4$, as long as $\bar{\beta}_q$ and $\bar{\beta}_p$ are both real. For example, if l is halved, the curves in Figs. 1, 2, and 3 for frequencies above the critical frequency will be raised by 12 dB. Below the critical frequency it is not so simple because of the real exponents arising in the 'phase' terms.

6 Reflexion in a Single Mode

For reflexion of a mode we merely take β_q as $-\beta_p$ in equations (10), (14), (15), (23) and (24), and by $+\bar{\beta}_p$ in equations (27) and (29). The results then become identical with those of Ref. 1. As an example, for the same taper as in Section 5.2 the reflexion of the H_{01} mode has been computed over the same frequency range. The result is given in Fig. 4. Again, the power in the reverse H_{01} mode varies as $1/l^4$.

7 Discussion and Conclusion

The coupling into forward and reverse waves is given generally by equations (9) and (10) respectively. The conditions for validity of the theory are given in Section 3, and it is found that even if one of the modes cuts off at some point along the length of the taper these conditions may still be satisfied. The general method of calculating the coupling factor at a point in the taper from a known coupling coefficient for a step in the cross-section is discussed in Section 4, and in Section 5 is applied to the case of H_{0p}- to H_{0q}-mode coupling in a tapered circular h.s.p. waveguide. Equations (24) and (26) give the coupling into the forward and reverse modes respectively for this case. A raised-cosine taper with radius varying between 2·5 cm and 0·9 cm, and length 100 cm, is then considered; equations (28) and (29) give the forward and reverse coupling, and numerical values for coupling between the H_{01} and H_{02} modes are given in Figs. 1, 2, and 3 for the frequency range 32–50 GHz. The coupling into a reverse mode may be into the same mode as the forward mode, and then $\bar{\beta}_q$ is replaced by $\bar{\beta}_p$ in equation (29); results for this are plotted in Fig. 4 for the same taper as in Figs. 1, 2, and 3.

A general feature of all tapers is that if one of the modes just cuts off at either end of the taper, there will be a peak of the overall coupling at the corresponding frequency. Another general feature is that as long as both modes are above cut-off throughout the whole length l of the taper, the power output in the unwanted mode, as long as it remains small, varies as $1/l^4$. Thus if l were halved, the whole curve in Fig. 4 and those parts of the curves of Figs. 1, 2, and 3 which are above the critical frequency would be raised by 12 dB.

The theory is developed in terms of the phase constants and characteristic impedances of the modes concerned, and is applicable whenever these can be expressed as functions of the parameters of the cross-section of the guide. These parameters are geometrical dimensions—radius of a circle, length of side of a rectangle, etc.—and dimensionless properties of materials—dielectric constants, relative permeabilities. The theory is thus applicable, as discussed in Ref. 4, to waveguides whose outer bound is a perfect conductor or a very good conductor and which contains materials which may be very slightly lossy but not very lossy. Thus a dielectric-rod waveguide or a waveguide containing a semiconducting material would be excluded because in these cases it is not possible to define a characteristic impedance. The theory applies, subject to the characteristic impedance being definable, whenever one or more of the parameters of the cross-section is changing sufficiently slowly. Thus the 'taper' may not necessarily be the change of a dimension—it may be a changing dielectric constant or other property. The theory is still valid even if one of the modes is cut off for part of the length of the taper.

Postscript

Since this paper was written a paper[7] has appeared in which the same problem is solved numerically on a computer and an optimum design is arrived at. The results are similar in form to those obtained in this paper, although an exact comparison is not possible because the form of the taper was not the same. An analytical design procedure could be developed from the present theory along the same lines as for the case of reflexion in a single-mode structure in Ref. 1, but this was not done because for present purposes the coupling is negligible.

8 Acknowledgment

I wish to thank the Director of Research of the Post Office for permission to publish this paper.

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IERE News and Commentary

The 48th Annual General Meeting of the IERE

The Institution's 48th Annual General Meeting (the 12th since Incorporation by Royal Charter) was held at the London School of Hygiene and Tropical Medicine on Thursday, 25th October 1973.

The meeting was opened by the President, Mr. A. A. Dyson, at 6 p.m. when 53 corporate members had signed the Attendance Register.

The President first called upon the Secretary, Mr. G. D. Clifford, to confirm that all members had received due notice of the meeting. Mr. Clifford reported that notice of the twelfth Annual General Meeting of the Institution since its Incorporation by Royal Charter, together with the Agenda, was published on page 508 of the August 1973 issue of *The Radio and Electronic Engineer* which had been distributed to all members.

Minutes of previous Annual General Meeting

The Secretary reported that a report of the eleventh Annual General Meeting of the Institution, held on 7th December 1972, were published in the January/February 1973 issue of *The Radio and Electronic Engineer* (page 160).

The President moved that, as no adverse comment or amendment had been received this report should be approved as the official minutes. Members gave their unanimous assent.

Annual Report of Council

Mr. Dyson then referred members to the Annual Report published on pages 569 to 587 of the September 1973 issue of *The Radio and Electronic Engineer*, and said:

'At the last Annual General Meeting I drew attention to our work within the Council of Engineering Institutions. Acceptance by the fifteen Chartered Engineering Institutions of the Common CEI Examination has led to problems in establishing the mutual acceptance of exempting qualifications and the admission to membership of mature candidates who have demonstrated in practical terms their value to the engineering profession.

'In these matters, the engineering profession as a whole must speak with one voice, but speak with full regard to the particular problems of engineers by recognizing both academic achievement and practical ability in such fields as production techniques, marketing and management.

'The need for British unanimity of view on standards for professional recognition is emphasized by the agreement made at the time of the British entry into Europe. We are now faced with the basic problem of balancing the European concept of "quantity", i.e. number of years of university studies—against the British practice of 'quality', i.e. qualifications by experience and Institution examination. Fortunately for engineers, FEANI (the European Federation of National Associations of Engineers) has already organized a *de facto* recognition of British qualifications for those who wish to practise in Europe.

'Your Council's views on such important issues depend very largely on the advice of the Education and Training, Academic Standards and Membership Committees. I commend the Reports of these three Committees to every member.

'The changeover from an Examinations Committee to a Committee responsible for looking at the Academic Standards open to the radio and electronics engineer has already been justified.

'The need to retain the Education and Training Committee is shown by its vigilance in ensuring proper ventilation of problems affecting recruitment to our branch of the engineering profession.

'The weight of the views of the IERE is also reflected in the Report of our Membership Committee. I urge all members throughout the world to read the comments in the Annual Report on membership recruitment.

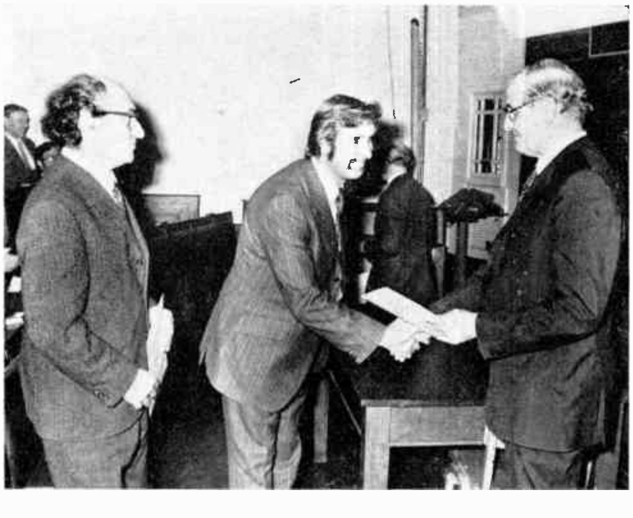
'Your Council continues to be concerned with ways in which the Institution can continue to provide services to a very large number of overseas members who are outside Europe.

Photograph above. After the Director and Secretary, Mr. Graham D. Clifford, has read the notice convening the meeting, the President signs the Minutes of the previous Annual General Meeting.

During the discussion on the accounts the Director, Mr. Clifford, deals with a point while the Treasurer and President refer to the balance sheet.

The President congratulates Dr. M. J. Buckingham, joint author of the paper which received the Clerk Maxwell Premium as the most outstanding contribution to the Journal during the year.

The President congratulates Mr. R. T. Irish when presenting him with the Marconi Award.



Mr. A. J. Allen and Mr. P. Atkinson are presented with their Local Section Award.

The President stands by while Mr. J. Langham Thompson (Past President) presents the Premium which he endowed in 1961 to Mrs. J. M. Brown of UWIST. This was the third occasion on which this Premium for the outstanding paper on Control Engineering had been awarded to Professor D. R. Towill and members of his Dynamic Analysis Group.

During the interval between the conclusion of the Annual General Meeting and the presentation of his Inaugural Address the President, Dr. Ieuan Maddock, talks with three of his predecessors and their officers of the Institution. To be seen in the picture are (left to right) Mr. H. F. Schwarz, Mr. G. A. Taylor, Mr. S. R. Wilkins, Major General B. D. Kapur and Mr. A. A. Dyson. Also in the picture, beyond Mr. Taylor and Mr. Wilkins, is Sir Leonard Atkinson.

This is reflected in the Report headed "Overseas Relations". Through CEI and our own activities in the Commonwealth there are well-established lines of communication, but as I was reminded quite recently by one of our members in America, we in the radio and electronics profession should particularly look at promoting a more alive association with our colleagues in America.

'Sadly I must report that at this moment there seems little hope of achieving common studentship registration through the Council of Engineering Institutions. This idea, originally promoted by my immediate predecessor, Mr. Harvey Schwarz, must be followed up. We are not discouraged by the immediate failure of our proposal any more than we were deterred by the idea of having such a Council of Engineering Institutions and the registration of professional engineers which was first proposed by Sir Louis Sterling—when he was the Seventh President of the Institution.*

The President then asked whether any member wished to comment on the report. Major-General Brahm D. Kapur (Fellow), a Past Chairman of the Council of the Indian Division, rose to express his appreciation for all the work that went into the well-being and advancement of the Institution and its members, which had been so amply illustrated in the Annual Report. General Kapur said that he was grateful to all the members who devoted so much valuable time serving on Council and its Committees: the Indian Division derived much inspiration from the energetic activities of the Institution in the United Kingdom. As no further questions were asked, the President moved from the Chair that the Annual Report of the Council be adopted. This proposal was unanimously approved.

Auditors' Report, Accounts and Balance Sheet

Before calling on the Honorary Treasurer, the President said that having now retired from professional life, Mr. Taylor had asked the Council to relieve him of his duties so as to make way for a younger man. The Council had reluctantly accepted Mr. Taylor's decision, and wished to record their gratitude to a colleague whose unselfish efforts had contributed so much to the progress of the Institution (Applause).

The President then formally called on Mr. Taylor to present the Accounts and Balance Sheet for the year ended 31st March 1973, together with the Auditors' Report.

In his own inimitable manner, George Taylor said he would resist the opportunity to recount his 20 years as Honorary Treasurer in a lengthy 'swan song'. He asked only to be recorded as having found it eminently worthwhile to have been Treasurer during the important years leading to the grant of a Royal Charter to the Institution,† and the subsequent twelve years of the Institution's further progress. Mr. Taylor continued: 'I was tempted to relate my report tonight to the last report of my predecessor—Mr. S. R. Chapman—when he presented the Accounts of the Institution for 1952–53. I am not quoting at length, however, but found it very interesting to see from what Mr. Chapman then said—"The success of the Institution must always lie with the growth of membership", that we were then concerned about increasing costs, and in his last year as Treasurer he was able to assure members that every reasonable economy was being exercised. Mr. President, I could use similar words in presenting the accounts twenty years later!

Mr. George Taylor presents his final report as Honorary Treasurer of the Institution.

'I suppose that it is always the burden of the Treasurer to report escalating costs and the need for economy. I have, however, the supreme advantage of having served the Institution during the momentous years of its growth. The pound was worth more in those days,‡ of course, but Chapman's last accounts showed that we had a total income of nearly £22,000 and that we managed to have a surplus of £2,500. The figures before you in last year's accounts testify to the growth of our Institution. As in every other form of life, however, the growth in income is inevitably accompanied by a

* *J. Brit. I.R.E.* 3, November 1942, p. 33 *et seq.*

† George Austin Taylor was named in the Grant of the Royal Charter as the Honorary Treasurer of the Institution.

‡ On the basis of the Retail Price Index the purchasing power of £1 has fallen since 1952 to 48p in 1972.

growth of expenditure, and it is my duty to report such facts to our members.

'Having refused the temptation to be loquacious on my last appearance, I beg all members to study carefully the report of our Finance Committee. Last year, and indeed the year before, I reported to the membership that we were doing everything to avoid increasing subscriptions. It is obvious that we can no longer delay, but I think we can all face up to our responsibilities by meeting the modest increases which were agreed at the Council Meeting this afternoon. The details will, of course, be circulated to all members,* but overall we believe that by making an average increase of £3 per member we shall not only be able to overcome our present difficulties but, I hope, to meet inflation for the next two to three years.

Dr. J. Stephen receives the Lord Rutherford Award from Mr. Dyson.

'In all my years as Treasurer I have never found our members failing to understand the increases that have had to be made from time to time, and with this step I believe that I may give up the reins of Treasurership in the confident knowledge that our Institution will grow from strength to strength.'

Mr. Taylor then said that he would be pleased to answer questions about the accounts. After a brief discussion on the reduction in Institution's liabilities compared with last year, he proposed the adoption of the Auditors' Report, Accounts and Balance Sheet. This was seconded by Mr. S. R. Wilkins (Fellow) and unanimously approved by the meeting.

Election of Council for 1973-74

The President referred to the nominations made by Council and circulated to corporate members by notice dated 26th June 1973. There being no opposing nominations, a ballot had not been necessary, and Mr. Dyson was sure that members would like to join him in congratulating Dr. Ieuan Maddock on becoming the 23rd President of the Institution (Applause).

The Heinrich Hertz Premium for the outstanding paper on mathematical and physical electronics has just been presented by the President to Dr. J. R. James and Lieutenant I. N. L. Gallett. The Institution Editor, Mr. F. W. Sharp, is also in the photograph.

Air Commodore S. M. Davidson, Professor W. Gosling and Mr. A. St. Johnston had been re-elected as Vice-Presidents and were joined by two new Vice-Presidents in Dr. P. A. Allaway (Fellow), Chairman of EMI Electronics Ltd., and Professor G. B. B. Chaplin (Fellow), Chairman of the Department of Electrical Science at the University of Essex.

Mr. S. R. Wilkins (Fellow), a member of the Finance Committee and a Vice-President in 1972, had been elected to succeed Mr. G. A. Taylor (Fellow) as Honorary Treasurer.

Ten new Ordinary Members of Council had been elected, namely Mr. R. C. Hills, His Royal Highness the Duke of Kent, Mr. P. L. Mothersole, Mr. K. G. Nicholls, Mr. J. Powell (Fellows), Mr. H. Blackburn, Professor D. W. Lewin and Mr. M. M. Zepler (Members) and, for the first time under the revised Bye-laws to represent the class of Associate, Mr. G. Phillips, and for Associate Members, to be represented by Mr. C. R. Fox.

Mr. R. J. Westcott receives the Leslie McMichael Award from Mr. Dyson.

* *The Radio and Electronic Engineer*, **43**, No. 11, p. 703, November 1973.

Mr. Dyson expressed his appreciation for the support he had received from all the members of Council, and he gave particular thanks to the twelve members of Council who were now retiring in accordance with the Bye-laws.

Appointment of Auditors and Solicitors

The President said that he proposed to combine Items 5 and 6 of the Agenda, namely the appointment of Auditors and of Solicitors. Mr. Dyson asked for approval to the re-appointment of Gladstone, Jenkins and Company as the Institution's Auditors, and that their remuneration be at the discretion of Council, and that Braund and Hill be re-appointed as Solicitors to the Institution. The President's motion was carried unanimously.

Presentation of Premiums and Awards

The formal business of the meeting having been concluded, the President passed to the presentation of premiums and awards to the authors of outstanding papers published in the Institution's Journal during the year. Mr. Dyson called upon the Institution's editor, Mr. F. W. Sharp, to announce the names of the prize winners and their papers.

Full details of these were published in an Appendix to the Annual Report. Mr. Sharp drew the attention of the President and members to the fact that twelve of the sixteen premiums were being awarded for 1972. At least one of the recipients of each of the premiums was present with exception of the winners of the Charles Babbage Award who were both resident in Canada.

The President congratulated the prize winners and presented them with the books or scientific instruments which they had chosen. He called upon Mr. J. Langham Thompson (Past President) to present personally two of the premiums which he had endowed, the J. Langham Thompson Premium for the outstanding paper on Control Engineering and the Dr. V. K. Zworykin Premium for Medical and Biological Electronics.

The Meeting closed at 6.40 p.m. and after a short interval to allow non-member visitors to assemble in the lecture theatre, the new President, Dr. Ieuan Maddock, presented his Inaugural Address 'The Engineer on Tap or on Top'. (This will be published in the January 1974 issue of *The Radio and Electronic Engineer*.) A vote of thanks to Dr. Maddock for his original and thought-provoking Address was moved by Mr. A. St. Johnston (Vice-President) and approved with enthusiasm.

ANNOUNCEMENTS

Index for 1973

The index to Volume 43 of *The Radio and Electronic Engineer* will be prepared early in 1974 in its customary format. In order to make some economies in the Institution's use of paper copies will only be sent automatically to libraries and other organizations which subscribe to the *Journal*. Members of the Institution who wish to obtain a copy of the index either for including in bound volumes or for reference purposes may obtain a copy free of charge on application to the Publications Department, Institution of Electronic and Radio Engineers, 8–9 Bedford Square, London WC1B 3RG. As a limited number only will be printed it would be helpful if such requests are made as soon as possible. Indexes will be incorporated in the bound volumes supplied by the Institution.

Colloquium on Leaky Feeder Radio Communication Systems

A Colloquium on Leaky Feeder Radio Communication Systems is to be held at the University of Surrey on Tuesday, 9th April 1974, organized jointly by the University, the National Coal Board, British Rail and the IEE and the IERE.

Visits to working installations in a coal mine and on British Rail are being arranged as part of the programme and will take place on the previous day. Overnight accommodation will be available at the University of Surrey as required.

The proceedings will be confined to two-way mobile radio communications using h.f., v.h.f. or u.h.f. techniques where coupling is predominantly via the leakage fields associated with coaxial or 2-wire transmission lines. Such systems are expected to find increasing application underground, in mines and tunnels, and overground on transportation systems where the required area of coverage is of limited lateral extent. The emphasis will be on scientific and engineering principles.

The Colloquium has been planned to follow a Conference on Radio Communication in Mines, Roads and Tunnels, organized by the Institut National des Industries Extractives (INIEX) which takes place in Liège on 1st–5th April 1974 and is concerned with systems and applications aspects. Delegates to the Conference are being invited to join the British tour programme.

Further details on the Colloquium may be obtained from Miss A. J. Perkins, Department of Electronic and Electrical Engineering, University of Surrey, Guildford, Surrey GU2 5XH. (Telephone: Guildford (0483) 71281.)

New Director for EEA

Mr. M. S. Ollivant, M.B.E., D.S.C., has been appointed Director of the Electronic Engineering Association in succession to Captain R. A. Villiers, C.B.E., who retired at the end of the year. Mr. Ollivant who has been Assistant Director of the Association for the past two years, joined EEA in 1970 on retirement from the Royal Navy in the rank of Captain.

During his service career Captain Ollivant had close links with the electronics industry, serving as an Applications Officer at the Admiralty Signal and Radar Establishment, Deputy Director of the Navigation and Direction Division at the Admiralty, and Captain of HMS *Dryad*, the Navigation and Aircraft Direction Training School. He spent two years in command of HMS *Protector*, engaged in scientific exploration in the Antarctic, and completed his naval career in command of the commando carrier HMS *Albion*.

The Clerk Maxwell Lodge

Mr. I. W. C. Robertson, A.F.R.Ae.S., has been elected as the 19th Master of the Clerk Maxwell Lodge for the year 1973–74.

The Lodge is supported by a number of members of the Institution and information regarding Lodge Meetings can be obtained from the Secretary, Mr. S. J. H. Stevens, B.Sc., C.Eng., F.I.E.R.E., The Birches, Park Close, Fetcham, Leatherhead, Surrey.

The British Electronics Industry in 1972

According to the latest survey of the electronics industry, just published by the Electronics Economic Development Committee,* during 1972 total sales of electronics products increased considerably after the 1971 recession – by 17% over 1971 at current prices. This high growth rate nearly matches the pre-1971 recession expansion during the late 1960s in money terms, although the latest figures particularly take account of rising prices.

The consumer goods sector was the main growth area. The colour television market strengthened further in 1972 and contributed to a record demand for components. Sales of colour television sets nearly doubled the 1971 figure to reach nearly £207M in 1972; imports were over £41M, Japanese sets alone representing nearly 9% of the total home market. Audio equipment sales increased, particularly sales of gramophone turntable units.

Telecommunications sales of approximately £320M represented a further rise over the very large increase in sales recorded for 1971. Computers and communications equipment only just held their own, in money terms, with 1970 levels sales, being £220M and £250M. In this year's edition, for the first time, it has been possible to treat industrial and scientific instruments as a separate sector: sales of industrial and scientific instruments were almost £350M, approximately £150M of which were for the export market.

The Trade Balance

The Survey shows that the worsening of the small trade deficit in 1971 to over £70M in 1972 was largely owing to the overwhelming home demand for colour television sets and components. While the computer sector's exports of peripheral units nearly trebled in 1972 (contributing to a halving of the computer trade deficit) and the industrial and scientific instruments sector slightly improved its export surplus, all the other sectors ended 1972 with a slightly worse trade balance than at the end of the previous year.

Total exports as a proportion of total sales of the whole industry fell from 29% in 1971 to 27% in 1972. The import share of the home market increased to 28% in 1972 from 25% in 1971. Imports of components, consumer goods, computers and industrial and scientific instruments took as much as one-third of the share of the home market of their respective sectors.

Communications

Turning now to a more detailed examination of the Survey, it is pointed out that the radio communications, radar and navigational aids sector is heavily dependent on public purchasing policies at home and abroad. The major buyers are defence departments, civil aviation and port authorities, public utilities and local authorities and the shipping industry. Projects involving radar and navigational aids usually span a period of several years, and therefore trends are difficult to detect and analyse. The overall figures seem to indicate however a slowing down in the home market, which is consistent with reduced Government expenditure in the field of civil air traffic control and defence. In overseas markets, it is difficult to explain the trends other than as random annual variations on a steady export business in a.t.c. and defence equipments from this country.

Breaking down the overall figures, a downward trend is visible in UK manufacturers' deliveries of airborne radar and navigational aids; US import penetration has deepened and UK exports are down from 1971, widening the already negative trade balance. In contrast, ground communications equipment sales improved by 20% to £56M, reflecting rising sales of military and civilian mobile radio equipment. Market expansion did, however, push imports up 16% to £1·6M – Japan's contribution rose 50% to £¼M – and syphoned off over 16% of 1971's £20M exports.

The bright spot in this area is the marine business: world shipping is expanding and UK companies are reaping some of the benefits. UK manufacturers' home market sales of shipborne communications equipment rose a further 20% over 1971 levels and deliveries of radar and other navigational aids were over 30% higher. Moreover, exports of marine communications equipment, but not marine radar and navigational aids, continued to rise, and imports in both categories increased, especially communications equipment from Japan and West Germany and marine radar from Italy.

There is little demand for public broadcasting equipment at present from either the BBC or the IBA and the decline in home market sales has continued. Licences have recently been granted to local commercial radio stations but their demand for transmission and studio equipment will be relatively small. Moreover, government decisions on a fourth television channel and on community antenna television have been postponed. Fortunately exports in 1972 were buoyant due mainly to a general rise in the market following the recession which started in the United States some three years ago. Some countries, in particular members of the Commonwealth are moving towards colour transmissions.

Instruments

The survey of the industrial and scientific instrument sector shows that total sales and work done during 1972 amounted to some £346M. Estimates of previous years' sales at current prices for the sector show that the 1972 level was relatively unchanged compared with 1971, after a period of steady growth since 1968 at around 10% per annum. Exports amounted to £154M in 1972, representing over 40% of sales. This is in fact the only sector showing a consistent rise over the period 1970–72. Process measuring and control instruments amounting to £139M account for nearly one-third of the sales of the scientific instruments. It is estimated that this sector has been the fastest growing, sales having increased at between 15 and 20% since 1968. The major market for these types of instrument is made up by four major industries, namely chemicals, power, petroleum, and iron and steel: it is estimated that these four industries, alone, account for about 80% of process control applications.

Computers

A significant proportion of computer manufacturers' effort is devoted to software, payment for which is to a large extent included in hardware sales. While no separate information on all software activities exists, DTI's Computer Services Enquiry does cover identifiable computer services provided to

* 'Annual Statistical Survey of the Electronics Industry' HMSO, November 1973, price 65p, postage paid.

clients. The results of the enquiry include a large number of bureaux and software houses and in 1972, nearly £80M worth of billings to all types of clientele in 1972 were identified. This represented an increase of 15% over 1971. One-quarter of the billings to clients was for the service for custom-built programs, and a further 30% was for the other activities of computer processing including package programs, time hire and remote access. Professional services accounted for 20% and the remainder of billings were for data preparation and other activities. Nearly one-third of total billings were for parent or associate companies. Of the remaining net outside billings some 70% of work done was for UK clients other than public services. The organization providing these services employed nearly 15,000 full-time staff and 800 part-time staff. Programmers and analysts accounted for one-third of total staff and another third were in data preparation.

Industrial R&D

The statistics for research and development are two years behind the other figure. It is estimated that about £147M was spent on scientific research and development in electronics in the financial year 1969–70: £130M was in private industry, £13M in public corporations and £¾M in research associations. The amount spent in private industry was 8% higher than in the previous year although part of this expenditure is to be accounted for by cost inflation.

It should be pointed out that the contribution from Government is not a measure of total research and development effort by Government in electronics. It refers only to Government funds provided for research and development carried out within private industry, and therefore excludes expenditure by Government on research relevant to electronics but carried out within organizations such as public corporations, universities, and Government research establishments. These figures also exclude the research and development done by private industry on customer account. This research and development is considered as part of the industry's output and is included in production statistics.

The Industry's Labour Force

Recovery in output of the electronics industry in 1972 was achieved while the labour force was reduced slightly. However, an additional ten thousand people were employed in the consumer goods sector between 1971 and 1972. All other sectors saw a contraction in the number employed.

Over half the employees in employment in the electronics industry are concentrated in the South East of England, about 11% in the Midlands and about 10% in the North West. The telecommunications sector is exceptional in that as few as a quarter of its employees are in the South East, with the Midlands, North West and Northern regions taking a larger share than other sectors. The Scottish electronics industry employs about 7% of the total labour force in electronics in Great Britain. The only regions where there appeared to be a high pressure of demand for labour in 1972 was for women in the South East, East Anglia and the South West.

The earnings of male manual workers averaged £33 a week. This was £1·50 a week less than the average across all manufacturing industries. Average weekly earnings without overtime for men were £28 a week, and for women just under £18 a week. The hourly rate for women manual employees is slightly higher than the average in manufacturing industry.

Safety in Industry

Discussing general conditions of employment and in particular the labour losses due to accidents, the rather surprising point is made that although the order of risk in the various sectors of the electronics industry is low compared with manufacturing industry as a whole, the incidence of severe accidents per 100,000 employees is rising slightly. Thus in electronics and scientific instruments, the figures for 1970–72 were respectively 200, 230 and 240, in industry as a whole the figures are falling: 670, 590 and 580 for those years. The components sector has the highest rate of notifiable accidents, about 28 per 1,000 employees, almost half in total of the accidents experienced in the whole industry.

Ownership of the Industry

Some of the first results of the Electronics EDC's analysis of Company financial performance in the electronics industry are shown in the survey. Summary tables show that over one-third of the top 200 companies in the electronics industry are foreign-owned and they produce one-third of the 200 companies' sales. In the components and computer sectors, foreign-owned companies provide as much as half of the sectors' total sales.

What is the Electronics Industry?

Electronics has always proved a difficult subject area to define—so also is the electronics industry. However, the Survey has adopted as a primary criterion in deciding whether or not a product (system, assembly or sub-assembly) is electronic the importance of active electronic compounds in its functioning.

Television Interference from Ignition Systems

New regulations controlling interference from the ignition systems of internal combustion engines* were laid before Parliament at the end of July by Sir John Eden, Minister of Posts and Telecommunications. They extend the frequency range controlled by the present regulations in order to give greater protection to the whole range of frequencies now used by television. They have been made after consultation with the Minister's Advisory Committee on Wireless Interference from Ignition Systems and came into force on October 1st.

The existing Wireless Telegraphy (Control of Interference from Ignition Apparatus) Regulations made in 1952 require assemblers, importers and users of ignition systems forming part of combustion engines (other than in aircraft) to ensure that the field strengths of electromagnetic energy radiated at frequencies between 40 MHz and 70 MHz do not exceed specified limits. These regulations will still apply to ignition apparatus assembled in engines before 1st October 1973.

The new regulations which extend the frequency range to 250 MHz conform with the agreed international standards for radio interference suppression with which all vehicles manufactured on and after 1st October 1973 have to comply under the Motor Vehicles (Construction and Use) (Amendment) (No. 4) Regulations 1972.†

* The Wireless Telegraphy (Control of Interference from Ignition Apparatus) Regulations 1973, SI No 1973/1217 published by HM Stationery Office.

† SI No 1972/1734 published by HMSO on behalf of the Department of the Environment.

Letters to the Editor

The Institution's Council does not necessarily agree with views expressed by correspondents. Correspondence of a technical nature, or on any matter of interest to electronic and radio engineers, is welcomed.

From: A. Brown, C.Eng, M.I.E.E., M.I.E.R.E.
and F. G. Diver, M.B.E., C.Eng., F.I.E.R.E.

The Next Stage in Television?

The editorial in the October 1973 issue of *The Radio and Electronic Engineer* includes statements regarding the case for satellites which are rather misleading.

For a country of the size of the United Kingdom, it would be not only uneconomic, but also somewhat impracticable to beam transmissions only to outlying areas. It must be remembered that, taking account of the restricted size of the satellite transmitting aerial and the feasible stabilization and aerial pointing accuracy, the minimum practicable beamwidth is likely to be 0·5 deg, and such a beam would cover, for example at least the whole of Scotland. It is therefore misleading to suggest that satellites would be suitable for supplying outlying areas, at least as far as the United Kingdom is concerned.

I would also query the statement that s.h.f. satellite transmissions have the potential for a vast number of channels. Considering the band 11·7–12·5 GHz, which was allocated to the broadcasting satellite service by the World Administrative Radio Conference in 1971, the current policy of the European Broadcasting Union is that this band would be used to provide national services. If all the European countries are to share the available bandwidths, studies indicate that it would be practicable to provide four or possibly five television channels per country. The 41–43 GHz and 84–86 GHz bands could also be used, in principle, but suitable technology is unlikely to be available for a considerable time. In any case, even these bands could not be said to provide 'vast' numbers of television channels for each country.

I realise that my comments apply to only a small part of the editorial, and what I have said does not affect its main conclusions.

A. Brown

European Broadcasting Union,
Technical Centre,
Avenue Albert Lancaster 32,
1180 Bruxelles, Belgium.

5th December 1973.

[Overconciseness in dealing with complicated technical arguments meant that it may not have been made sufficiently clear that the use of broadcasts from satellites was not contemplated as a solution for the United Kingdom.

Regarding Mr. Brown's second point, the development of the higher s.h.f. bands will, of course, depend very much on technological advances. While the potential number of channels using satellites does not compare with waveguide or optical-fibre transmission techniques, a considerably greater number of national channels than the present v.h.f. and u.h.f. bands would be feasible in principle: their allocation to different purposes within a particular country would be a limiting factor to the effective increase.—*Editor*.]

Quality and the Engineer

The twelfth Annual General Meeting of the National Council for Quality and Reliability has been held recently. I draw satisfaction from the current emphasis placed on quality and the widespread use of the word—a desirable state of affairs which did not exist when NCQR was formed. Not that NCQR can be given the sole credit. But it can now be seen how timely and far-sighted was the concerted action which brought NCQR into being. The inspiration and support which it received from the professional ranks helped to ensure that the corporate thinking on the merits of the quality theme, as well as on its philosophies and techniques, was sound and pertinent to current needs. No longer has NCQR the pioneering role but its continued existence is a necessity if only to serve as a focal point for the extremely wide range of interests involved.

But in one respect I find use of the word 'quality' a little too widespread. I deplore the trend to use the term Quality Engineer in job advertising and earnestly hope that it will not be adopted as a formal title within any organization.

In making this observation I am well aware that *The Quality Engineer* is the title of the Journal of the Institute of Quality Assurance. This is a special usage always to be seen in context with the scope of that Institute. The latter, be it noted, is not the Institute of Quality Engineers.

I make no objection to such titles as Quality Director, Quality Manager, Quality Assurance Engineer or Quality Control Engineer; these are very acceptable job descriptions. An engineer with overall responsibility for the quality of products is in any event much more aptly termed Quality Manager. I am sure that those among his staff with appropriate tasks to perform need not look askance at the respectable word 'test' or even the seemingly old-fashioned 'inspection'!

Let those tempted to designate a man, chartered or otherwise, as a Quality Engineer reflect a maxim which has had universal use and acceptance throughout NCQR's life—'Quality is everybody's business'. Most certainly is it the concern of every engineer.

F. G. Diver

117 Oldfield Road,
Stannington,
Sheffield S6 6DU

15th December 1973

[Mr. Diver has been the IERE's representative on NCQR since the Council was formed in 1960; he is a past chairman of the Institution's Technical Committee.—*Editor*.]

Members' Appointments

CORPORATE MEMBERS

Mr. D. K. Baker, B.Sc.(Eng.) (Member 1964) has been appointed Divisional Managing Director, Instruments, with Advance Electronics Limited, Bishops Stortford. Mr. Baker has been with Rank Strand Electric for the past four years as an Engineering Manager.

Col. J. A. Baker (Member 1968), until recently on the staff of the Chief Signal Office, BAOR, has been appointed Project Manager WAVELL and BATES with the Procurement Executive Ministry of Defence.

Major D. E. Blenkinsop, REME (Member 1972) who has been Officer Commanding Ground and Airborne Instrumentation at the Royal Artillery Range in the Hebrides for the past two years, has been posted as Officer in Charge of the Aircraft Technical Service Unit, REME, at the Army Aviation Centre, Middle Wallop.

Mr. J. Cotterell (Member 1948, Associate 1943), Principal of Llandaff Technical College, Cardiff, since 1954, has been appointed to the Engineering Industries Training Board as one of the five 'Educational Members'. Mr. Cotterell is a past chairman of the South Wales Section.

Mr. M. W. Dudley (Member 1970, Graduate 1964) has been appointed Product Engineer with IBM UK Manufacture at Havant. He joined IBM in 1968 and has latterly been a Product Testing Engineer at Hursley.

Mr. K. G. Dunn (Member 1970, Graduate 1968) has been promoted to Senior Lecturer in Electrical Engineering at York College of Further Education. He joined the College in 1970 as a Lecturer II in electronics.

Mr. R. C. Fawell, B.Sc. (Member 1971) has moved from Marconi-Elliott Avionic Systems where he has worked as a Project Engineer for the past two years to join the Solartron Electronic Group as a Sales Engineer.

Mr. D. M. Fidler, B.Sc. (Member 1961, Graduate 1960) is now joint Managing and Technical Director of Tridem Transformers and Electronics Ltd., Ringwood, Hants. Mr. Fidler was Chief Designer with Gardners Transformers Ltd. for some twelve years. He served on the Components and Circuits Group Committee from 1968 to 1972.

Mr. D. E. G. Gibbons (Member 1969) has been appointed Senior Signals Officer in charge of peripheral engineering in the Civil Aviation Authority's Directorate of Data Processing. He was previously in the National Air Traffic Control Service.

Sqn. Ldr. D. F. Grimston, RAF (Member 1971) has been posted to RAF Gan, as Officer Commanding 6 Signals Unit. For the past two years he was RAF member of the Joint Signals Staff, U.K. Commanders-in-Chief Committees.

Mr. H. R. Holliday (Member 1972, Graduate 1962) has formed a manufacturing engineering consultancy—Coupar Angus Associates in Birmingham. Mr. Holliday has been for the past 15 years on the staff of the Royal Radar Establishment, latterly as a Higher Scientific Officer.

Major N. Rome (Member 1970) has recently retired from the Army where his final appointment was DAD EME Electronics, HQ BAOR and has joined Marconi Radar Systems Ltd. as a Radar Sales Engineer.

Mr. B. R. Veale (Member 1969, Graduate 1966) has joined Garrett Manufacturing Ltd. at Rexdale, Ontario as General Manager of Microelectronics Products. Before emigrating to Canada in 1970, Mr. Veale was with Mullard Southampton Works as a Unit Manager.

NON-CORPORATE MEMBERS

Mr. J. N. Elven (Associate 1966) joined Rank-Hilger at Margate earlier this year as General Sales Manager and has now been appointed Marketing Manager. He was previously with Standard Telephones and Cables.

Mr. A. H. Harrold (Graduate 1967) is now with Pentec Services, Johannesburg, a firm of consultant engineers in which he is the partner responsible for instrumentation.

Mr. W. J. Lloyd (Graduate 1970) has retired from REME where he was latterly Warrant Officer I at 73 Field Workshop (Aircraft) and is now at Garnett College where he is taking a teacher's training course for the certificate of education.

Mr. J. C. Newman (Graduate 1970) is now Project Engineer with Delamain Ltd, Ammanford, South Wales. He was previously with the Marconi Company at Chelmsford.

Mr. A. R. Pakes (Graduate 1969) is now Lecturer in Electrical Engineering, South East Berkshire College of Education. His previous appointment was as a Quality Engineer with British Aircraft Corporation, Stevenage.

Mr. J. A. Rutter (Associate 1961) who has held engineering appointments with broadcasting stations in Libya and Mexico is now in the Head Office Engineering Section of the New Zealand Broadcasting Corporation, Wellington.

Obituary

The Council has learned with regret of the deaths of the following members.

Oswald Francis Mingay, M.B.E. (Fellow 1945) died on 8th April 1973, aged 77 years.

Mr. Mingay was one of the founders of the Institution of Radio and Electronics Engineers Australia and was its General Honorary Secretary from 1932 to 1940. He later held the offices of Vice-President and Deputy President. His early career was spent with the Australian Post Office and he then moved into industry as Radio Manager of Burgin Electric Co. In 1930 he founded Australian Radio Publishers, subsequently the Mingay Publishing Company, of which he was managing director until his retirement in 1965. During the war years he served with the Australian Corps of Signals, for much of the time being concerned with the production of radio equipment for forces in the South West Pacific area.

Soon after the war Mr. Mingay spent some time in London where he became well known to engineers in the Radio Industry and particularly to those in the technical publishing field. From 1950 until his retirement Mr. Mingay was the representative of the IERE Council in Australia.

His contributions to the Australian IREE were recognized by election as an honorary life member. Some years ago he endowed an award for the most outstanding paper delivered at a divisional meeting of the IREE in any year; this is now to be renamed the Mingay Memorial Award.

William Reginald Curry (Member 1955) died on 29th August 1973, aged 61 years. He leaves a widow and grown-up family.

Educated at Emanuel School, Wandsworth, London, Mr. Curry joined the Post Office Engineering Department in 1930 as a Youth-in-Training and continued his studies at the Regent Street and Northampton Polytechnics in London. He was successively promoted to Skilled Workman, Inspector, Assistant Engineer and, in 1952, to Executive Engineer. His early Post Office career was concerned with telephone equipment but from 1940 his main activity was with the Radio Branch, particularly from 1952 to 1954 on work associated with the new P.O. Radio Station at Rugby and with Cable and Wireless stations.

In 1954 he transferred to the then Ministry of Supply Directorate of Electronic Production (Radar) as an Engineer III engaged in the provision and repair of airborne radars. He was subsequently promoted to Engineer II and his duties within the Procurement Executive extended to cover navigational equipment. After his retirement in 1972 Mr. Curry undertook private consultancy work.

Forthcoming Institution Meetings

London Meetings

Wednesday, 16th January

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Colloquium on DOPPLER VOR SYSTEMS

IERE Lecture Room, 2.30 p.m.

Second-generation Doppler VOR
By M. S. Whitney and T. R. Tapsell (*DTI*)

A New Doppler VOR System
By F. Thomas (*Plessey Radar*)

A Versatile v.h.f. Power Amplifier with Loop Controlled Amplitude Modulation
By M. Jackson (*Plessey Radar*)

Doppler VOR Commutator
By M. Chambers (*Plessey Radar*)

Further details and registration forms from Meetings Secretary, IERE.

Wednesday, 23rd January

COMPONENTS AND CIRCUITS GROUP

Colloquium on RECENT DEVELOPMENTS IN TURNTABLE DESIGN

MEETING POSTPONED

Thursday, 24th January

EDUCATION AND TRAINING GROUP

Training the Euro-Engineer

By F. R. J. Langridge (*Engineering Employers Federation*)

IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

Monday, 28th January

JOINT IEE/IERE COMPUTER GROUP

Designing Machines for People—Human Factors in Computer Design

By Dr. C. R. Evans (*NPL*)

IEE, Savoy Place, W.C.2, 5.30 p.m. (Tea 5 p.m.).

Wednesday, 30th January

COMMUNICATIONS GROUP

Colloquium on ACTIVE AERIALS

IERE Lecture Room, 2 p.m.

An Electrically Small Receiving Monopole Sited above a Transmitting Antenna—the Coupling Problem

By J. F. Goodey (*ASWE*)

Some Design Aspects of H.F. Active Aerials
By A. K. Roberts (*Plessey*)

Autonull—a Method of Overcoming Co-site Problems using an Active Aerial
By M. M. Zepler (*Plessey*)

Short Integrated Active Aerials
By Dr. T. S. M. Maclean (*Birmingham University*) and Dr. P. A. Ramsdale (*Lanchester Polytechnic*)

H.F. Active Antennae—Performance Requirements and Realization
By B. M. Sosin (*Marconi*)
Further details and registration forms from Meetings Secretary, IERE.

Wednesday, 6th February

AUTOMATION AND CONTROL SYSTEMS GROUP

Digital Phase Lock Loops

By K. Thrower (*Racal*) and P. Atkinson (*Reading University*) IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

The digital phase-locked loop is widely used for frequency synthesis and control. The lecture is concerned with the design, performance and practical applications of the loop. The shortcomings of the conventional Type 1 loop are discussed and it is shown how the Type 2 loop, which has integral control, overcomes these. In considering the applications of the digital phase-locked loops the problems of oscillator design, noise, jitter and frequency range are considered.

Wednesday, 20th February

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Situation Display

By A. Harrison (*Kelvin Hughes*)
IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

The use of an optical storage medium capable of retaining a radar picture for periods up to 15 or 30 minutes has led to the achievement of a mode of display of radar data on a moving ship, originally suggested many years ago but never successfully carried out. A simple ship's head-up relative display is optically projected on the image retaining panel, which is viewed by a closed-circuit television system. The panel is moved electro-mechanically to correspond to the ship's movements as indicated by log and gyro. The unique combination of electronic, optical, and electromagnetic techniques has produced a radar display with properties of real daylight viewing by more than one person, and the indication of position, course, and speed of other ships, in a manner extremely cost effective and operationally very attractive.

Wednesday, 27th February

COMMUNICATIONS GROUP

Data Communications—The Next Ten Years

By P. T. F. Kelly (*Post Office Telecomms.*)

IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

Data communications is now just over 10 years old and there are now over 30 000 data terminals in the U.K. making use of data transmission facilities. Information is being obtained on likely growth patterns, data signalling rates requirement, and data traffic patterns, in order to assist in the planning of future Datel services. The next ten years could see the introduction of new point to point, multipoint and switched digital data services, perhaps initially on an experimental basis. Some indication of the Post Office proposals as currently foreseen and the impact that these could have on the existing services is given.

Thames Valley Section

Thursday, 24th January

New Sources of Power

MEETING CANCELLED

Thursday, 14th February

Mini-Computers—Wherever Next?

By M. Judd (*Data General*)

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

A critical review of the development of mini-computers covering the main technological advances which have brought about the current range of machines will be given. Extrapolation of present trends in both hardware and software will be given to predict the architecture of the next generation of machines, with particular emphasis on new applications.

East Anglian Section

Thursday, 24th January

JOINT MEETING WITH IEE

Planning for U.H.F. Television Coverage

By A. L. Witham (*IBA*)

University Engineering Laboratories, Trumpington Street, Cambridge, 6.30 p.m. (Tea 6 p.m.).

Since the end of 1969, three u.h.f. television services have been available in colour to a large and increasing proportion of the country. The lecture describes some of the problems which have been encountered in attempting to provide a high quality of air signal to as many viewers as possible. As the unserved areas decrease in size, the planning techniques require increasing refinement and service criteria need closer definition.

Thursday, 7th February

Advances in Display Technology

By D. W. Byatt (*Marconi*)

Civic Centre, Chelmsford, 6.30 p.m. (Tea 6 p.m.).

Current display techniques applied to data handling displays such as radar and computer control systems are described together with the application of new materials. In particular the use of light-emitting semiconductors and electro-optic chemical materials will be discussed.

Wednesday, 13th February

JOINT MEETING WITH IEE

Digital Measurement Techniques

By R. Vivian (*IBA*)

Ipswich Civic College, 6.30 p.m. (Tea 6 p.m.)

The ability of an equipment or channel to transmit faithfully a video signal appearing at its input terminals may be investigated in many ways. One of the most obvious would be to apply audio testing techniques and plot amplitude and phase responses against frequency.

However, the results of this simple if tedious process cannot always be related easily to subjective assessments of visual impairments. Early attempts at resolving these difficulties resulted in the practice of supplementing the frequency domain information with transient response curves, but as the applied waveforms often contain frequency components outside the normal passband, the interpretation was still not as straightforward as might have been supposed originally.

Further efforts to improve the correlation between measured and subjective degradations focused attention on measurements in the time domain, using waveforms whose frequency distributions were well defined.

Southern Section

Wednesday, 30th January

JOINT MEETING WITH IEE

Colloquium on COMPACT MOBILE TRANSMITTERS—IMPROVEMENTS FOR HIGH STRESS ENVIRONMENTS

The Cinema, Plessey Company, West Leigh, Havant, 4 p.m.

Thursday, 31st January

Meteorological Telecommunications Systems Engineering

By C. E. Goodison (*Meteorological Office*)

Farnborough Technical College, 7 p.m. (Refreshments available in College Refectory from 6.30 p.m.).

Meteorological telecommunications systems consist of different scales of size and sophistication ranging from world encompassing systems which are computer controlled to simple 50-baud teleprinter links. Following the pattern of present day developing communication systems the present picture of meteorological telecommunications is one of rapid change. The lecture will describe some of these systems and the subjects will include the Global Telecommunication System of the World Weather Watch, its Main Trunk Circuit and particularly the Regional Telecommunication Hub at Bracknell.

Wednesday, 6th February

Charge Coupled Devices

By Dr. J. D. E. Beynon (*Southampton University*) University of Surrey, Guildford, 6.30 p.m.

Although the charge-coupled device was conceived only three years ago it is already challenging many conventional integrated circuit techniques, particularly in the memory and solid-state imaging fields. This is because of the device's extreme simplicity which is leading to circuits having high packed density, low power dissipation and low cost per function. The lecturer will explain the operation of the charge-coupled device and describe some of the techniques used for fabricating c.c.d. circuits. Some of the c.c.d.'s many present and future applications will be discussed.

Friday, 8th February

Design of Displays for Radar Data Handling Systems

By C. E. G. Cooke (*Plessey Radar*)

Isle of Wight Technical College, 7 p.m.

Traditional radar displays utilize a plan position indicator with a long persistence phosphor. Modern techniques of plot extraction, tracking and flight plan processing generate large amounts of data which must be displayed. Three types of display result—'raw radar', 'all synthetic', and 'time shared'. The paper discusses the design of such displays and the compromises involved.

Wednesday, 13th February

Submillimetre Waves

By Professor D. Harris (*Portsmouth Polytechnic*)

Portsmouth Polytechnic, Anglesea Road, 6.30 p.m.

The wavelength from about 1 mm to 0·1 mm is now being explored. The lecture will outline the development of this region, present the state of the art so far as generation, detection and modulation devices are concerned, consider the properties of materials and the atmosphere at these wavelengths, and discuss the possibility of low-loss waveguides for submillimetre operation. Applications, including special purpose communications and image forming systems, will be discussed and possible developments outlined. A demonstration of submillimetre lasers, techniques and waveguides will be given after the lecture.

Tuesday, 19th February

Project Management in the 1970s

By R. H. Bradnam (*Urwick Technology Management*)

Bournemouth College of Technology, 7 p.m. (Refreshments in Refectory 6.30 p.m.).

The paper will describe the factors involved in setting up a 'new project', and will review current practices and procedures of project management. The problems associated with running a project will be discussed, and brief mention will be made of possible future trends.

Kent Section

Wednesday, 30th January

Medical Electronics

By W. J. Perkins (*National Institute for Medical Research*)

Lecture Theatre 18, Medway and Maidstone College of Technology, Chatham, Kent, 7 p.m.

Thursday, 28th February

Multiphonic Organs

Lecture by J. H. Asbery

Lecture Theatre 18, Medway and Maidstone College of Technology, Chatham, Kent, 7 p.m.

The principle of the multimonophonic organ (usually abbreviated to multiphonic) involves the use of a small number of oscillators, the frequency of these being determined according to the keys pressed. Attention will be drawn to the relative advantages of a.c. resistive, a.c. capacitive and d.c. keyboard switching and to systems using the divider principle. Aperiodic frequency multipliers may be used as an alternative to dividers. Tone forming by use of non-linear elements and modulation of one footage by another will be mentioned. Some of the techniques can be used where a conventional melodic section is provided in a polyphonic organ. While the concept of the multiphonic organ is over a quarter of a century old, its commercial exploitation has hitherto been inhibited by lack of inexpensive components of sufficient stability.

Yorkshire Section

Thursday, 17th January

JOINT MEETING WITH IEE

The Problems of Radio Systems above 10 GHz

By Dr. P. A. Watson (*University of Bradford*)

University of Leeds, 7 p.m. (Refreshments 6.30 p.m.).

The effects of rainfall, multipath propagation and other lower atmospheric phenomena on the propagation of radio waves above 10 GHz will be discussed and related to the design of systems at these frequencies. Some possible future satellite and terrestrial radio relay systems using these frequencies will be described.

Thursday, 21st February

JOINT MEETING WITH IEE

Mini-Computers and Their Application

By N. Emslie (*Hewlett-Packard*)

Leeds University, 6.30 p.m.

The first part of the presentation will explain to an engineering audience the principles of operation of mini-computers with emphasis given to the important architectural features which influence the performance of complete systems. This will be followed by a brief review of the concepts of software and comparison between scientific programming languages. Examples will be shown of the applications in engineering and science where mini-computers have made major contributions.

East Midland Section

Thursday, 17th January

Interfacing Strategy on Real Time Computer Systems with particular reference to CAMAC

By H. Bisby (*AERE, Harwell*)

Lecture Theatre 'A', Physics Block, Leicester University, 7 p.m. (Tea 6.30 p.m.).

CAMAC is a definitive style for implementing the interface conditions which exist when many channels of input/output information share a common data-processor/controller. The features of CAMAC are described to indicate the applicability of CAMAC-compatible equipment and programming to real-time situations. Some of these are illustrated by typical systems which use either a computer or a simpler device as the central processor/controller.

Wednesday, 13th February

Opto-Electronics

By M. Miller (*Texas Instruments*)

Lecture Theatre 'A', Physics Block, Leicester University, 7 p.m. (Tea 6.30 p.m.).

West Midland Section

Thursday, 24th January

JOINT MEETING WITH IEE

Electronic Ignition Systems

By R. L. Rivers (*Mobelec*)

Lanchester Polytechnic, Coventry, 7 p.m.

Thursday, 28th February

JOINT MEETING WITH IEE AND IPOEE

Low Frequency Navigational Aids

By A. Brooker-Carey (*Decca Navigator Co.*)

Post Office Training College, Stone, Staffs, 7 p.m.

South Midland Section

Monday, 14th January

JOINT MEETING WITH IEE

The Use of Radar in Meteorology

By Dr. R. Starr (*RRE, Malvern*)

Winter Gardens, Malvern, 7.30 p.m.

Tuesday, 19th February

Integrated Circuits for Radio Receivers

By J. Bryant (*Plessey Semiconductors*)

B.B.C. Club, (Evesham) 7.30 p.m.

South Western Section

Wednesday, 23rd January

New Electronic Components using Amorphous Semiconductors

By Dr. J. Allison (*University of Sheffield*)

Lecture Room 4E3.10, University of Bath, 7 p.m. (Tea 6.45 p.m.).

Thursday, 31st January

JOINT MEETING WITH IEE

High Fidelity Sound Reproduction

By R. L. West (*Polytechnic of North London*)

Main Hall, Plymouth Polytechnic, 7 p.m. (Tea 6.45 p.m.).

The lecture and demonstrations will be largely historical in nature, not only tracing the gradual improvement in sound quality over the years, but also showing how and why these developments came about. The lecture will cover the whole field of reproduced sound and highlight the most important landmarks on this journey.

Tuesday, 5th February

More and More for Less and Less (Integrated Circuits)

By P. C. Newman (*Allen Clark Research Centre, Plessey*)

The College, Regent Circus, Swindon, 6.15 p.m. (Tea 5.30 p.m.).

The art of integrated circuit design, since its inception about twelve years ago, has consisted of squeezing quarts first of all into pint pots, then into half pints, gills, fluid ounces, and so on. The reasons for doing this lie in the demands of system designers and sales managers.

The lecture will survey past accomplishments and future expectations. A 'crystal-ball' prediction of ultimate limits will be attempted.

Wednesday, 27th February

JOINT MEETING WITH IEE AND RAeS

Digital Instrumentation

By A. R. Owens (*University of Wales, Bangor*)

Queen's Building, University of Bristol, 6 p.m. (Tea 5.45 p.m.).

North Eastern Section

Monday, 4th February

JOINT MEETING WITH IEE

Change of Subject

Trends in Semiconductors

By Dr. K. J. Dean (*S.E. London Technical College*)

Merz Court, University of Newcastle upon Tyne, 6.15 p.m. (Tea 5.45 p.m.).

North Western Section

Thursday, 17th January

The Texan Stereo Amplifier

By R. Mann (*Texas Instruments*)

Lecture Theatre R/HIO, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.).

Wednesday, 27th February

JOINT MEETING WITH IEE

Data Transmission—Present and Future

By Messrs. Brenton and Wright (*Post Office*)
Lecture Theatre R/HIO, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.).
The lecture will comprise an introduction to data transmission, a section on modem design and the prospects for future development.

Merseyside Section

Wednesday, 13th February

Digital Systems Technology: Its Influence on Modern Electronic Systems

By A. K. Porter (*Liverpool Polytechnic*)

Department of Electrical Engineering and Electronics, University of Liverpool, 7 p.m. (Tea 6.30 p.m.).
The paper appraises the revolution which has occurred in the past decade in the use of digital technology in electronic systems. Applications in instrumentation, communications, and computer technology are reviewed with an indication of trends in the development of new components and design techniques.

Northern Ireland Section

Wednesday, 9th January

CHANGE OF DATE

Optoelectronics

By M. Miller (*Texas Instruments*)

Cregagh Technical College, Montgomery Road, Belfast 5, 7 p.m.

Wednesday, 13th February

The Ergonomics of Electronic Equipment
By G. R. Dickson

The Polytechnic, Jordanstown, 7 p.m.

South Wales Section

Wednesday, 13th February

ANNUAL GENERAL MEETING at 6.30 p.m.

Electronic Music

By K. Winter (*University College, Cardiff*)

Large Shandon Lecture Theatre, University College, Cardiff, 7 p.m.

Long before the advent of so-called electronic music, apparently non-musical sounds had been employed in various works. Today tape recorders and synthesizers have expanded the available range of sounds to such an extent that composers are no longer content with traditional orchestration. They demand an expansion of instrumental possibilities. It is this interaction between live performer and tape/synthesizer in the concert hall or on another level between composer and machine which is stimulating much new music today.

Scottish Section

JOINT MEETINGS WITH BRITISH COMPUTER SOCIETY (PUBLIC LECTURE)

From Morse Code to Data Networks

By E. B. Stuttard (*Racal Milgo*)

Wednesday, 16th January

Napier College of Science and Technology, Edinburgh, 7 p.m.

Thursday, 17th January

Boyd Orr Building, Glasgow University, 7 p.m.

The two basic problems of telephone communication are transmission and switching. The transmission characteristics of the telephone network, designed for the human voice, are shown to have created the need for data modems whose complexity increases with the data rate required. The main modulation and equalization techniques are reviewed. The switching problem is then examined and a comparison is made between various types of leased line network, the public switched network, multiplexing and concentrating. The principles of these techniques are described and system examples are shown. The lecture closes with a look at digital transmission techniques and the various national digital networks being planned or implemented throughout Europe.

JOINT MEETINGS WITH IEE

Concorde Electronics

By H. Hill (*BAC*)

Monday, 11th February

Room 406, James Weir Buildings, University of Strathclyde, Glasgow, 6 p.m.

Tuesday, 12th February

South of Scotland Electricity Board Showrooms, 130 George Street, Edinburgh, 6 p.m.

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

Thick Film Circuits

G. V. PLANER and L. S. PHILLIPS. Butterworths, London 1972. 21·5 × 14 cm. 152 pp. £4·00.*

CONTENTS: Applications of thick film circuits. Substrates. Conductor patterns. Resistor patterns. Printed capacitors and insulating layers. Printing procedures. Firing procedure. Hybrid circuits—attachment of chip and other discrete devices. Trimming and test procedures. Environmental protection. Circuit design concepts. Future trends.

The authors set out to provide an understanding of the materials and techniques involved in thick film processes. They succeed remarkably well in that the book is very readable and extremely easy to understand—in fact it requires absolutely no prior knowledge of microelectronics.

One of the most useful features of the book is the list of references at the end of each chapter which add up to an extensive bibliography of the subject. The utility of the book would, however, have been improved by the inclusion of a trade index giving the reader an indication of where to buy equipment.

The authors give a reasonable comparison between the differing micro-circuit technologies and show that there is an increasing role for thick films in hybrid technology. Overall the book provides a good review of present day techniques and of the standards achieved in thick film circuitry.

J. C. ANDERSON

(Dr. Planer is Managing and Technical Director of G. V. Planer Ltd. and Mr. Phillips is Senior Executive Scientist with the Company.)

Magnetic Recording

CHARLES E. LOWMAN. McGraw-Hill, Maidenhead, Berkshire 1972. 22·5 × 16 cm. 285 pp. £6·95.*

CONTENTS: Why magnetic recorders? The development and advantages of magnetic recording. Introduction to the basic elements of the magnetic recorder. The theory of magnetism. Magnetic heads. Magnetic tape. The recording and reproducing process. The transport. Direct record and reproduce signal electronics. Frequency-modulation record and reproduced electronics. The television recorder. Cassette and cartridge systems.

This work is a practical guide to the technology of magnetic recorders used in such fields as audio recording, broadcast and closed-circuit television, instrumentation and computer data systems. The design concepts, manufacturing procedures and operating parameters for a wide variety of magnetic heads, tapes, disks, etc. are discussed. Mathematical treatment is kept a minimum.

(Mr. Lowman is with the Ampex Corporation.)

Electronic Systems for Radio, Television and Electronics Mechanics

RHYS LEWIS. Macmillan, London 1973. 22·5 × 16 cm. 223 pp. £2·40.

CONTENTS: Electronic signals. System subunits. System block diagrams. Active and passive components. Power supplies. Amplification. Oscillation. Mixing and detection. Switching and counting. Transmission and propagation. System faults. Self-test questions.

Intended for students studying basic radio, television and electronics, the book presents the fundamental theory and application of electronic systems in non-mathematical terms. Objective test questions similar to those set in the C. & G. examination papers are included.

(Mr. Lewis is a Senior Lecturer at the Openshaw Technical College, Manchester.)

Field Effect Transistors

N. R. BIJLSMA, P. BURWELL and E. G. EVANS (Editors). Mullard, London 1972. 22 × 16 cm. 131 pp. £1·80.

CONTENTS: Introductory survey. The junction field-effect transistor. The MOS field-effect transistor. Electrical properties. Circuits with field-effect transistors.

The work is designed to introduce the user of f.e.t.s. to its operating principles, characteristics and terminology of these devices so that their special potentialities can be recognized. Relative structures and principles involved in both junction and insulated-gate f.e.t.s. are discussed. Electrical properties are dealt with in some detail and in conclusion circuit configurations and typical application are discussed. Engineers with the Philips Organization, from Germany, England and the Netherlands have contributed to this book.

Theory of Electrical Transport in Semiconductors

B. R. NAG. Pergamon Press, Oxford 1972. 26 × 18 cm. 227 pp. £4·50.*

CONTENTS: Electrons in semiconductors. Collision processes in semiconductors. Transport coefficients. Relaxation time for the collision processes. Physical parameters related to transport. Transport characteristics of the important semiconductors. Theory of high-field transport.

This book is written to provide a text on the various aspects of the theory of conduction in semiconductors and is suitable for undergraduate students and other new entrants in the field. The latest development, namely, the hot-electron conduction, is dealt with and an account of the theory and experiments related to the phenomenon is given.

(Professor Nag (Fellow 1964) is at the Centre of Advanced Study in Radio Physics and Electronics, University of Calcutta.)

Handbook of Wiring, Cabling and Interconnecting for Electronics

CHARLES A. HARPER (Editor-in-chief). McGraw-Hill, Maidenhead, Berkshire 1973. 22·5 × 16 cm. 1142 pp. £14·75.*

CONTENTS: Soldered, welded and mechanical terminating systems. Basic selection of connector systems. Hookup wires, multiconductor cables and associated terminating devices. Coaxial cable and connector systems. Wire and cable for communication systems. High-voltage wiring and connector systems. Magnet wire and insulated conductors. Rigid printed wiring and connector systems. Flexible printed wiring and connector systems. Flat conductor cable and connector systems. Formed high-frequency circuits. Microelectronic interconnection systems.

This book represents a great wealth of material in diverse interconnexion fields and discusses such traditional technologies as magnet wire and coaxial cable, as well as newer techniques such as printed circuits, flat cable and microelectronic interconnexions. There is a wide coverage of competing techniques and these data are generally tabulated so that the user may compare different techniques at a glance. A large bibliography is included.

(Mr. Harper is with the Westinghouse Electric Corporation, Baltimore, Maryland.)

Electrical Principles and Science for Electrical Installation Students

4th and 5th Years.

G. WATKINS and W. E. JONES. Blackie, Glasgow 1972. 18·5 × 12 cm. 150 pp. £1·40 paperback.

CONTENTS: D.C. distribution. Magnetic and inductive circuits. Time constant. Direct-current generators and motors. Electrostatics. Capacitance and resistance in series. Time constant. Single-phase a.c. circuits. Single-phase and three-phase motors. Three-phase circuits. Power-factor improvement. A.C. distribution. Utilization of electrical power. Single-phase transformers. Space and water heating. Tariffs. Illumination.

Digital Integrated Circuit D.A.T.A. Book

Edition No. 14. D.A.T.A. Inc., New Jersey, U.S.A. 1972. 28 × 21·2 cm. 554 pp. £16·10.

More than 12,700 types from 84 worldwide manufacturers are included with nearly 1800 outline and schematic drawings permitting specific i.c. selections. A special circuit term thesaurus clarifies the varying terminology used to describe integrated circuits.

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.

Understanding Technology

Charles Süsskind. Johns Hopkins University Press, Baltimore 1973. 22·5 × 15 cm. 164 pp. £3·30.

Contents: Birth, coming of age and rise of modern technology. The computer technology. Some aspects of contemporary technology. Ideologies of technology. Technology as a social force and ethical problem. Challenges.

Engineers increasingly resemble physicians in their concern for the human condition, but the engineering code of ethics still deals mainly with business relations among practitioners. This is the view expressed in a book which forcefully suggests that engineers must not be mere spectators. They play a crucial part not only in opening new doors to knowledge, but also in deciding which doors to open. Therefore it is doubly important for engineers to be guided by the highest moral principles. A Hippocratic Oath for Engineers is put forward: 'I will not use my professional knowledge contrary to the laws of humanity; I will endeavour to avoid waste and the consumption of nonrenewable resources'. This American publication clearly has great interest to Chartered Engineers in Great Britain in view of current discussions revolving around the new CEI Code of Conduct (which forms part of the IERE Bye-Laws).

The author is well known for his informed and articulate papers and books on the historical development of modern technology and he discusses the effects of technology—including, of course, electronics—on fields with which it is not usually associated—new foods, the health sciences, the fine arts, education and humanities. These sections in particular are both fascinating and highly informative. Passing then to broader social and political implications, he traces such 'ideologies of technology' as technocracy, the managerial revolution, utopias and anti-utopias, and the work of contemporary thinkers. 'Technology is too important to be left to the engineers' says Süsskind in a paraphrase of Clemenceau's saying about war and the generals and he believes that only by the concerted efforts of all can the abuses of technology be avoided.

F.W.S.

(Professor Süsskind (Member 1950, Associate 1940) has been a member of the Berkeley Engineering Faculty of the University of California since 1955. Several of his papers on electron devices have been published in the Journal.)

Fifty Photoelectric Circuits

P. S. Smith. Butterworths, London 1972. 22 × 14 cm. 83 pp. £2·30.*

Contents: Switching with photoelectric cells. Photographic applications. Novel circuits. Industrial uses. Domestic applications. Automobile applications. Control circuits. Special devices.

Design details of circuits incorporating a number of basic applications of photoelectric devices are explained. These include simple light-measuring instruments, switching circuits for operating lights, counting units and smoke detecting circuits for use with fire alarms.

(Mr. Smith was formerly a Technical Product Engineer with Mullard Ltd.)

Elements of Linear Microcircuits

T. D. Towers. Butterworths, London 1973. 22 × 15 cm. £2·80.*

Contents: What a linear microcircuit is, how it is made and packaged. Survey of types and selection. Handling and safety precautions. Three generations of operational amplifiers. Everyday uses of monolithic operational amplifiers. Audio amplifiers. Radio and intermediate frequency amplifiers. Wideband amplifiers. Voltage regulators. Amplitude modulated receivers. F.m. radio receivers. Television receivers.

Based on a series of articles written by the author for *Wireless World*, this book gives guidance for selection of commercially-available devices. Throughout the book the emphasis is on applications and practical problems encountered in the design of electronic equipment. The book would be found useful by students at H.N.C. and higher levels, as well as being a useful reference for practising engineers.

(Mr. Towers (Member 1964) is a Director of Newmarket Transistors, Ltd.)

Techniques of Circuit Analysis

G. W. Carter and A. Richardson. Cambridge University Press, London 1972. 22·5 × 16 cm. 548 pp. £5·00.

Contents: Direct current circuits. Ideal circuit elements. Alternating current circuits and the use of phasors. The use of complex numbers in sinusoidal analysis. Introduction to the theory of transients. Reduction methods and network theorems. The frequency characteristics of elementary circuits. Two-port networks. The harmonic analysis of alternating quantities. Techniques of transient analysis. Fourier transforms. The complex plane in circuit analysis and synthesis. Distributed circuits.

Written primarily for undergraduate students the book provides instructions and practice in the methods of analysis which are essential in solving electrical circuit problems. Laplace transforms, matrix algebra, Fourier integrals and complex plane are explained and analysis of distributed circuits and transmission lines under steady-state and transient conditions is included. Examples are worked out to illustrate the methods described.

(Professor Carter and Mr. Richardson are at the University of Leeds.)

Feedback Control Theory for Engineers. (Second edition).

P. Atkinson. Heinemann Educational Books, London 1972. 20·5 × 14 cm. 467 pp. £2·95.

Contents: Introduction to control engineering. Graphical representation of signals. The use of complex numbers in the solution of vector problems. The principles of mechanics and simple electrodynamics using the SI system. The solution of linear differential equations with constant coefficients. Equations of physical systems. Control system components. The dynamics of a simple servomechanism for angular position control. Transfer functions. Introduction to frequency response methods. Series compensation using the Nyquist diagram. Parallel compensation using the inverse Nyquist diagram. Logarithmic representation of frequency response functions. Process control systems. Analog computing and simulation. Appendix: The Laplace transforms.

The book now includes a chapter on analogue computation and simulation which incorporates sections on digital simulation of control systems and hybrid computers.

(Mr. Atkinson (Member 1962) is a Lecturer in Applied Physical Sciences, University of Reading.)

Continued from opposite page

STUDENTS REGISTERED

BRIGHT, Stephen Maxwell. *London, N.W.4.*
ELSTOB, Harry Leigh. *Hemel Hempstead, Hertfordshire.*
FLETCHER, Ronald Irwin. *Claverton Down, Bath, Somerset.*
GODDARD, Robin. *Corsham, Wiltshire.*
THOMAS, Desmond Morgan Brice. *Shrewsbury, Shropshire.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

AYIVORH, Samuel Clifford. *Kumasi, Ghana.*
GREEN, Robert Frederick Dennis. *New South Wales, Australia.*
LUK, Kam Ming. *Hong Kong.*
SAH, Shanker Lal. *Dehra Dun, India.*
SALEEMI, Mohammad Hanif. *Tripoli, Libya.*

Direct Election to Member

BAGALKOTI, D. R. *Surathkal, India.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

MUNASINGHE, Upali Dharmadewa. *Gampola, Sri Lanka.*

STUDENTS REGISTERED

CHIA, Hoo Nguan. *Singapore 19.*
KULARATHNE, Aluthwatte Durayalage Piyadasa. *Narammala, Sri Lanka.*
LUK, Koon Min. *Kowloon, Hong Kong.*
PAY, Loo Seng. *Geylang, Singapore 14.*
WONG, Tat Wai. *Kowloon, Hong Kong.*

Notice is hereby given that the elections and transfers shown on Lists 166, 167 and 168 have now been confirmed by the Council.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 6th and 20th November 1973 recommended to the Council the election and transfer of 134 candidates to Corporate Membership of the Institution and the election and transfer of 16 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: 6th November 1973 (Membership Approval List No. 169)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

ADAMS, Edward. *Greenock, Renfrewshire.*
ANDERSON, William. *Earley, Berkshire.*
BARBER, John Dennis. *Radcliffe-on-Trent, Nottingham.*
BARTLETT, Alan Charles. *Weybridge, Surrey.*
BERMINGHAM, Alan Patrick. *Evesham. Worcestershire.*
BISHOP-MILLER, William Barrie. *Riseley, Bedford.*
BHATTACHARYA, Sabyasachi, B.Sc., Ph.D. *Gosta Green, Birmingham.*
BRAIN, Ronald Alan. *London, S.E.9.*
BROWN, Roy. *Welling, Kent.*
BUTT, Sidney John. *St. Albans, Hertfordshire.*
CAME, David Kestell. *Plympton, Plymouth, Devon.*
CHUNG, Paul Cheuk Pui, M.Sc. *East Bergholt, Suffolk.*
COOPER, Allan Leslie. *Plymouth, Devon.*
CROOKS, William Ronald. *Liverpool.*
DAVIES, Anthony Royston. *Cheltenham, Gloucestershire.*
DAVIES, Peter Frank, M.Sc. *Stevenage, Hertfordshire.*
DEAL, George Keith. *Halstead, Essex.*
DENNIS, Jack. *Burton-on-Trent, Staffordshire.*
DOWNING, Terence Ernest. *Brentwood, Essex.*
DULEY, Jeffrey Ernest Charles. *South Ruislip, Middlesex.*
EAST, George Frederick. *Twickenham, Middlesex.*
ELDER, Fergus. *Bishopbriggs, Glasgow.*
EVANS, Gordon Jack. *Bath, Somerset.*
FOLWELL, David John. *Bradville, Milton Keynes.*
FORSTER, Frederick Ronald. *Amersham, Buckinghamshire.*
FOSTER, Peter. *Bootle, Lancashire.*
FRAMPTON, Peter Guy. *Wokingham, Berkshire.*
GHALLEY, Jaspal Singh. *Tewkesbury, Gloucestershire.*
GILLMAN, Brian Francis Robin. *Codicote, Hertfordshire.*
GOODSELL, Roger Keith. *Hassocks, Sussex.*
HANNAH, David Anthony. *Writtle, Essex.*
HAY, Eric. *Edinburgh.*
HOARE, Anthony Michael Edward. *Farnborough, Hampshire.*
HUDSON, Nicholas James, Lieutenant, R.N. *Waltham Chase, Hampshire.*
JELLY, Christopher. *Biggin Hill, Kent.*
JENKINS, John George. *Hatfield, Hertfordshire.*
JOHN, Ellis Hugh. *Welwyn Garden City, Hertfordshire.*
JOHNSON, John Edward. *Crawley, Sussex.*
JONES, Brian. *Hucclecote, Gloucester.*
JONES, David Llewellyn. *Totnes, Devon.*
JONES, Kenneth Hugh. *Abbots Langley, Hertfordshire.*
JONES, Thomas Christopher. *Harrow, Middlesex.*
KEY, Malcolm. *Woodbridge, Suffolk.*
KING, Rodney Charles. *Hastings, Sussex.*
KNEESHAW, Robert William. *Frenchay, Bristol.*
KNIPE, Victor Thomas Alfred. *Erith, Kent.*
KNOWLES, Roger William. *Woodthorpe, Nottingham.*
LANGLEY, Roy Thomas. *North Harrow, Middlesex.*
LOUKES, Roger Keith. *Ely, Cambridgeshire.*
NEWMAN, David William. *Maldon, Essex.*
NEWTON, Raymond Joseph. *Hunger Hill, Bolton, Lancashire.*
NICHOLSON, Brian Steven. *West Hagley, Worcestershire.*
OLIVER, Oswald Hamilton. *Southall, Middlesex.*
OLIVER, Robert. *Christchurch, Hampshire.*
OLIVER, Robert James. *Herne Bay, Kent.*
OWEN, Raymond Barry. *Tonbridge, Kent.*
RAYMOND, Walter Kenneth. *Culcheth, Lancashire.*
RICHARDS, Nigel Bruce Cheffers. *Bookham, Surrey.*
ROGERS, Alan Charles. *Feltham, Middlesex.*
SHARP, John William. *Caversham Park, Reading, Berkshire.*
SMITH, William Henry. *Welwyn Garden City, Hertfordshire.*
SURTEES, John Anthony. *Biggin Hill, Kent.*
TANSLEY, Brian Arthur. *Snodland, Kent.*
TEMPLETON, James Greenshields. *Prestwick, Ayrshire.*
WHITFIELD, Robert Stephen. *Liss, Hampshire.*
WILSON, George David. *Crawley, Sussex.*
WITT, James. *Coventry, Warwickshire.*
YORK, Raymond William Rex. *Cheddar, Somerset.*

Direct Election to Member

BALE, David James, B.Sc. *West Molesey, Surrey.*
BURRETT, David John. *Crawley, Sussex.*
GARNETT, John Graham. *Kidsgrove, Stoke-on-Trent.*
HUNT, Leonard. *Ledbury, Herefordshire.*
JAMES, Keith. *Cove, Hampshire.*
LEWIS, Gareth Martin, B.Sc. *Croydon, Surrey.*
MACPHAIL, Neil. *Kidsgrove, Staffordshire.*
MEENEHAN, Michael, M.Sc. *Coombe Dingle, Bristol.*
PEEVOR, Raymond Colin, B.Sc., M.Sc. *Carshalton Beeches, Surrey.*
PROCTER, Kenneth John Michael, Squadron Leader, B.Sc. *High Wycombe, Buckinghamshire.*
WHEELER, Peter David, B.Sc., M.Sc. *Hook, Hampshire.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

DAVIES, Huw. *Caerphilly, Glamorgan.*
PATTERSON, Michael Anthony. *Arborfield, Reading, Berkshire.*

Direct Election to Associate

FAWCETT, Alan James. *Sleaford, Lincolnshire.*
WILLIAMS, Elwyn Morgan. *Pontyclun, Glamorgan.*

STUDENTS REGISTERED

ADEYEMO, Olatunde Christopher. *London, S.E.17.*
BLACKFORD, Donald Stewart. *Little Baddow, Chelmsford, Essex.*
BUTLER, Ronald Rhett. *Guildford, Surrey.*
MACBETH, John Robert. *Harrow, Middlesex.*
PARMAR, Pravin. *London, N.W.2.*
PLOWMAN, Susan Rosemary (Miss). *Reading, Berkshire.*
ROBERTS, Martin Henry. *Leicester.*
WELPTON, Noel. *Wheatley, Oxon.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

ADEKOYA, Moses Adewumi. *Kano, Nigeria.*
AGBANA, Elisha Ibitayo. *Kaduna, Nigeria.*
AKINDELE, Gabriel Oludotun. *Lagos, Nigeria.*
BAIDEN, Frederick Adams. *Accra, Ghana.*
DEMETRIOU, Theodoros, M.Sc. *Nicosia, Cyprus.*
HELLER, Gideon. *Kiryat-Bialik, Israel.*
HUNT, Walter John. *Nairobi, Kenya.*
IYAMABO, Adolphus Ehiakamen. *Lagos, Nigeria.*
JEYASINGHAM, V. *Seremban, Malaysia.*
JIYA, Edward Ayodele. *Lagos, Nigeria.*
OKEKE, Christian Nnaemeka. *Ikeja, Nigeria.*

Direct Election to Member

SUBRAMANIAN, Vaidyalingom, Wing Commander. *Bangalore, India.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

TAY, Joo Thong. *Singapore 3.*

Direct Election to Graduate

SLADE, Ivan. *Schloss Schule, Germany/B.R.D.*

STUDENTS REGISTERED

AU YONG, How Kai. *Kuala Lumpur, Malaysia.*
KURUPPU, Premalal Tejkumar. *Boralaugamuwa, Sri Lanka.*
NGOWI, Rodrick Henry. *Nairobi, Kenya.*
TANG, Chi Wai. *Hong Kong.*
WONG, Chung Hoo. *Port Dickson, N.S., Malaysia.*

Meeting: 20th November 1973 (Membership Approval List No. 170)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

BEDDOE, Neville Ernest. *London, S.W.19.*
BELL, Reginald. *Bletchley, Milton Keynes, Buckinghamshire.*
CAMERON, John Charles. *Berkhamsted, Hertfordshire.*
CLIFF, William Neil, M.Sc. *Henlow, Bedfordshire.*
CRANDON, Christopher. *Stanford-le-Hope, Essex.*
DOYLE, George Brian. *Wirral, Cheshire.*
GINN, Michael Graham. *Chelmsford, Essex.*
GRICE, Roy. *Guernsey, Channel Islands.*
HALL, Peter Michael. *London, E.C.2.*
HIND, David. *Lower Darwen, Lancashire.*
HOLT-KEENE, John Holford. *East Gomeldon, Wiltshire.*
HOWARTH, Alan Henry. *Sale, Cheshire.*
HUGHES, Stephen Wishart. *Pontypool, Monmouthshire.*
HUNT, Victor Royce. *Wokingham, Berkshire.*
JACKSON, Eric. *Leatherhead, Surrey.*
JACKSON, Hugh. *Darlington, County Durham.*
JOLLY, William Alfred. *Binley, Coventry.*
JURY, Anthony William. *Alstone, Tewkesbury, Gloucestershire.*
LANE, Michael Anthony. *Richmond, Surrey*
LAWRENCE, Alan Richard. *London, N.W.2.*
LAWRENCE, Trevor Richard. *Aldershot, Hampshire.*
LEWIS, John Alfred. *Totton, Southampton, Hampshire.*
LINTON, Roger Leslie. *East Grinstead, Sussex.*
LORD, Chrisopher. *Ashford, Middlesex.*
LOVETT, Walter Reginald. *Braintree, Essex.*
LOWNDES, John Dennis. *Knebworth, Hertfordshire.*
McELROY, Antony. *Greenock, Renfrewshire.*
McLEAN, Duncan Callam. *Bridge of Allan, Stirlingshire.*
McLEOD, Eric Malcolm. *Marton, Teesside.*
PAULINE, Leslie. *Liverpool, Lancashire.*
RILEY, George Charles. *Banbury, Oxfordshire.*
SAMIOTIS, Demetrios. *London, N.4.*
THROWER, Ivan Walter. *Borough Green, Kent.*
WIDDOWSON, Neville Bruce. *Redditch, Worcestershire.*

Direct Election to Member

McGINNES, Martin Screen. *Silvertonhill, Hamilton, Lanarkshire.*
REDHEAD, Peter Harvey. *Northwood Hills, Middlesex.*
TEASDALE, Dennis. *Middlesbrough, Teesside.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

MORTON, Henry McKnight. *Beeston, Nottinghamshire.*
WEST, Michael Graham. *Crawley, Sussex.*

Direct Election to Graduate

ANLEY, Barry Snowden. *Troon, Ayrshire.*
CLEMINSON, Peter. *Wokingham, Berkshire.*
HOWARD, Gordon William Mackinnon. *Bordon, Hampshire.*
HODGES, David Oliver. *Colchester, Essex.*
LAWRENCE, Peter Jamie. *Ipswich, Suffolk.*
PARRATT, Neville John, B.Sc. *Romford, Essex.*
PIPER, Stanley John. *Burnham, Buckinghamshire.*

Continued on opposite page

British Electronics Exports

Some recent contracts gained in Europe, Asia, the Americas and Australasia

AUSTRALIA

Labgear Ltd, Cambridge, has won the first of a series of orders from Australia for television service equipment to assist the conversion to colour broadcasting in Australia in 1975. The £35,000 order includes 520 colour bar generators, for use by service engineers when setting up and testing colour receivers. This equipment is the export version (v.h.f. on 625 lines) of the model developed by Labgear in collaboration with Granada TV Rental Ltd of Bedford.

CTF-Adaps Ltd, an Australian computer service bureau, has ordered three medium-sized Honeywell Series 2000 computers to cope with the general expansion of its business. The computers, Model 2050s together valued at £750,000, are being shipped from Honeywell's Newhouse, Lanarkshire, computer factory. CTF-Adaps has Honeywell 125 systems installed in its data centres in Sydney, Melbourne and Adelaide. The Model 2050s will replace these installations, though some of the peripheral units of each Model 125 will be retained.

BRAZIL

British colour television broadcasting equipment worth more than £200,000 is to be used for producing educational television programmes in South America under an export contract awarded to EMI. The Television Division of a leading subsidiary, EMI Sound & Vision Equipment Ltd, of Hayes, Middlesex, is supplying the Brazilian foundation, Fundação Centro Brasileiro de TV Educativa, with full broadcast-quality colour television systems including an outside broadcast vehicle and two EMI '2005' colour cameras.

The Television Division of EMI has already supplied '2005' cameras and other colour systems to television stations in Brazil which provided the colour pictures for the inauguration of Brazil's first colour television service in 1971.

Empresa Brasiliera de Telecomunicações (Embratel) has placed an order for fifty Marconi H1060 1 kW transmitters to be installed in new and existing shore stations. Early this year another GEC-Marconi company, Eddystone Radio Ltd, won a contract to supply sixty receiving sets as part of the same programme. The chain of Brazilian coast stations will therefore be largely equipped with transmitting and receiving equipment from Marconi Communication Systems Ltd.

BULGARIA

Rupert Neve & Co, of Melbourn near Cambridge and Kelso in Roxburghshire, has sold a 16-channel console for Bulgarian Radio and Television and a 24-channel console for the Congress Hall, Bucharest. This is Neve's first order from Bulgaria and more are expected.

FEDERAL REPUBLIC OF GERMANY

As a part of an agreement with Siemens for the supply of graphical display equipment, two further orders worth over £60,000 have been received by Ferranti. Under the agreement Ferranti is supplying graphical displays and display control modules, the present orders being for the Ferranti type MD4 display equipment which will be used with Siemens 300 process computer systems.

GREECE

Companies in the George Kent Group are to supply telemetry systems and other instrumentation valued at £70,000 for the first stage of a water management scheme under development in the River Pinios in the Peloponnesus, Greece. The Pinios project is for the control or irrigation of local farmland, and the contract covers the supply of instrumentation for the scheme's central control room and the automatic control of five remote pumping stations. It will incorporate the Dataflex acquisition/control systems for digital transmission of data and control signals over telemetry lines.

THE NETHERLANDS

A £160,000 order for data transmission equipment from Control Data Corporation of Holland is announced by Racal-Milgo Ltd. The equipment order includes T-16 time division multiplexers together with auto-equalized modems for 4800 bit/s and 9600 bit/s transmission speeds and will form part of the European end of an international time-sharing network. Model T-16 time division multiplexers provide for up to sixteen asynchronous channels and interleave low-speed data to form a single high-speed serial data stream which can then be transmitted, via high-speed data modems, over a telephone channel to a remote location where it is decoded to its original speed and character structure.

Bailey Meters & Controls Ltd. have received an order for full automatic control of four 225,000 t tankers being built by Verolme United Shipyards, Holland. The boiler control systems will be composed of the Bailetronic 370 modular system, which is capable of continuous operation in an ambient temperature of 70°C. Due to heat dissipation by the equipment itself the modules are designed to operate at a temperature of 85°C without effect on the system performance.

JAMAICA

Racal-Mobilcal has received an order to provide its Syncal 30 high frequency transmitter-receivers for installation in American-made armoured personnel carriers. The order comes from the Jamaican Defence Force and includes 100 watt amplifiers and vehicle harness systems, together with antenna tuning systems.

JAPAN

Redifon Flight Simulation Ltd has won a £1·6M order from Japan Air Lines. This contract for a Boeing 747 flight simulator incorporates the Redifon Duoview visual system. Redifon claims to be the only manufacturer in the world to have built simulators for the complete range of Boeing civil aircraft.

POLAND

Teofilow—a Polish textile complex—has placed an order with International Computers Ltd for a 1903T computer valued at £700,000. The computer will be installed at Teofilow's computer centre at Lodz, where the complex has a textile mill and four garment factories.

ROMANIA

Marconi Marine has received an order from Navimpex, the Romanian national shipping industry import and export agency, for the supply of complete communications installations for five new buildings at the Galatz shipyard. All five vessels will have similar installations based on the 1800 W *Conqueror* single-sideband transmitter and associated *Apollo* single-sideband digital display receiver.

SWITZERLAND

A specially designed K70 computer-based control system valued at £30,000 has been supplied to the Technikum Beider Basel, an engineering college in Switzerland, by Kent Automation Systems Ltd of Luton. The K70 equipment has been designed to incorporate as many variants as possible on the K70 interface to assist students in mastering the latest advances in process control using d.d.c. methods.

YUGOSLAVIA

An export order for two ICL computers—a 1902A and a 1904S with a total value of £550,000—has been placed with International Computers Ltd by Rudarsko-Metalurski Kombinat, Zenica, Yugoslavia's major iron and metallurgical combine. The 1904S is to be installed at the combine's steelworks at Zenica where it will be used for a wide range of applications. The 1902A computer is to be installed at the Hasan Brkic Metallurgical Institute where it will be used for scientific computing associated with iron and steel production.

An order for a 1903S computer system valued at £250,000 for the city of Rijeka has been received by ICL. The new computer is being supplied to the city's computer centre which operates as an independent organization to provide computing services for both local government and industry.