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The Electronics Industry and Higher Education

NATIONAL objectives in education are usually strongly influenced by social pressures and until recently there has been a steadily increasing demand for more and more places in higher education. The nature of the provision has been conditioned mainly by the 'Robbins principle'—that students should read the subject of their choice, and this, at a time when 'science' has temporarily lost its appeal, has resulted in a preferential expansion in non-science-based subjects. Thus, while on the whole, the broad National objectives have been met, a serious shortage has developed of students wishing to fill the available University places in the physical sciences and engineering.

With the development of the Polytechnics it was argued that more young people would be given 'relevant' education and that this might help to redress the worsening balance and increase the supply of graduates able and willing to contribute science-based skills to manufacturing industry. Unfortunately, the Polytechnic expansion rate has also been disappointing—for much the same basic reason that accounts for the University situation. It is often overlooked that the main source of supply from which undergraduate students in natural science and engineering is drawn is determined, numerically, largely by the numbers of sixth-form students reading mathematics and physics at A-level and that these numbers have not changed significantly since 1965. Thus, however many new courses are set up, they will not attract more people into engineering—for experience in many places has demonstrated that people who read 'Arts' A-levels seldom wish to change discipline subsequently, even when attractive opportunities to receive engineering education are afforded to them.

In this situation, therefore, with the Electronics Industry facing relatively optimistic growth prospects, it is essential to ensure that those students who do enter our Universities and Polytechnics shall receive not only an *education*, in the broad sense, but, further, a preparation which is also seen to be useful in the eyes of their future employers. To achieve this preparation satisfactorily, a continuing dialogue between the Educational and Industrial sectors is essential. At the present time, things are changing rapidly in 'education', as well as in 'industry', and the topic of 'The Electronics Industry and Higher Education' for the Conference to be held under IERE auspices at the Royal Holloway College from 28th–31st March 1974 is particularly timely. Conferences concerned with the Engineering Industry as a whole occur with reasonable frequency, but the Electronics Industry differs in many respects from other manufacturing sectors. The generalized conclusions of such conferences are always of interest—though they seldom, by their very nature, are of use in dealing with the problems of specific subject areas and it was for this reason that it was decided to make electronics the subject of special attention on this occasion.

The Conference will discuss the present state of manpower needs in the industry in the context of present National educational policies and will examine typical job profiles, with a view to seeing how appropriately present degree courses produce suitably qualified engineers. Ways in which the Educational sector can effectively collaborate, and thus 'refresh' itself, through research, consultancy and staff interchange will also be explored in depth. The speakers are widely drawn from Industry, Government and Education, and the views of the Science Research Council and NRDC will also be presented. The programme of the Conference is given in detail on page 104 of this issue.

Much has happened in recent years to bring about more genuine and realistic collaboration across the 'interface' and criticisms of education in electronics are now noticeably less strident than they were. Nonetheless, we still have much to learn from one another both within the Educational sector and across the frontiers with Industry. Occasions such as that which the forthcoming Conference will provide are, therefore, worthy of serious consideration by both Industry and Educators alike.

G. D. SIMS

Contributors*
to this issue

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Dr. Champa Bhushan received the B.Eng. degree in telecommunications from Jadavpur University, Calcutta, in 1964; the M.Eng. degree in electronics from Birla Institute of Technology and Science, in 1966; and the D.Eng. degree in electrical engineering from Sir George Williams University, Montreal, Canada, in 1972. She was a lecturer from 1966 to 1968 in the Department of Electrical Engineering, Birla Institute of Science and Technology; and from 1968 to 1969 in the Department of Electronics and Communications, University of Roorkee. At present she is working at Bell Northern Research in Ottawa.

Professor B. B. Bhattacharyya received his B.Tech. and M.Tech. degrees from the Indian Institute of Technology, Kharagpur, and the Ph.D. degree in electrical engineering in 1968 from Nova Scotia Technical College. From 1959–1965 he held appointments as lecturer at Kharagpur and at the Indian Institute of Technology, Madras. He joined the Electrical Engineering Department of the Nova Scotia Technical College in 1967 and in 1968 moved to the University of Calgary as an Assistant Professor in Electrical Engineering. Dr. Bhattacharyya joined Sir George Williams University, Montreal, in 1970 as an Associate Professor of Electrical Engineering.

* See also pages 69, 72 and 84.

Mr. R. W. J. Barker is a graduate of Nottingham University. After industrial experience he entered technical education in 1967 at the then West Ham College of Technology and from 1970 to 1973 he was a Senior Lecturer in the Department of Electrical and Electronic Engineering at Portsmouth Polytechnic. He was recently appointed Senior Lecturer in the Electronic and Electrical Engineering Department at the University of Sheffield. This is the third paper which he has contributed to the *Journal* in collaboration with Mr. B. L. Hart.

Mr. B. L. Hart (Member 1961, Graduate 1955) read physics at Queen Mary College, University of London, and after 10 years' experience in industry he joined the staff of West Ham College of Technology, now a part of the North East London Polytechnic. He is currently Senior Lecturer in charge of the Post-graduate Electronics Laboratory at the Polytechnic. A fuller note on his career was published in March, 1972 in connexion with a previous paper.

Dr. Joseph Watson has been on the staff of the Department of Electrical and Electronic Engineering of the University College of Swansea since 1961 and is now a Senior Lecturer; before going to Swansea he was Manager of Process Control Electronics with Hilger and Watts Ltd. A graduate of the University of Nottingham, where he also gained his doctorate, Dr. Watson has spent several years in North America. He obtained the S.M. degree from M.I.T. and in 1967–68 held a Visiting Associate Professorship at the University of California at Davis and he is at present spending several months at the University of Calgary as a Visiting Professor in the Department of Medicine.

Dr. Watson is author or co-author of three books and nearly 20 papers, his main field of interest having been in electronic instrumentation, more recently for medical applications.

Dr. D. R. Tanner obtained his B.Sc. degree in electrical engineering at the University College, Swansea, in 1971, and then followed the course in electronic instrumentation, obtaining an M.Sc. degree in 1972. For the past two years he has been working as a Senior Research Assistant in the Department of Electrical Engineering at Swansea on an oceanographic project aimed at developing an instrument to measure the mobility of sand grains beneath the sea.

Some observations on distance dependence in tropospheric propagation beyond radio horizon

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SUMMARY

This paper shows analytically that the distance exponent, n, obtained by comparing power levels at two points beyond radio horizon depends upon (i) the region beyond horizon where these points are located (diffraction or forward scatter region), (ii) the relative positions of these points, (iii) the initial gradient of refractivity and (iv) heights of the antenna terminals. In the diffraction region, the exponent critically depends upon (ii), (iii) and (iv) above. In the forward scatter region, the absolute value of n critically depends upon (iii) and (iv) but not on (ii). Observed results on variation of n over three paths (200 to 700 km) to illustrate these points have been presented. It has been shown that for paths with complicated geometry, angular distance exponent depends much less on path geometry and initial gradient. From the transmission loss data, possible values for spectral slope, m, in the relation $\phi(K) \propto K^{-m}$ and scattering parameter C_n^2 have been indicated.

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1 Introduction

Field strength observed at locations beyond normal radio horizon varies with distance in diverse manners depending upon the dominant mechanism of propagation. Smooth earth diffraction theory predicts an exponential decrease in signal level of 0·75 dB per km at 500 MHz and 1·5 dB per km at 4000 MHz.[1] Beyond the diffraction region, the long-term distance exponent for tropospheric forward scatter signal has been reported[2] to vary from −6 to −9 depending upon the climate.

The specific features of all forward scatter theories is the power law dependency on distance and this is often reported[3] to be in poor agreement with experimental results. In order to adjust with observed results, different authors have assumed different forms of height variation for scattering intensity.[4, 5]

The significance of distance dependence has not been adequately explained in published literature. Specifically, if received power levels at two identical receiving equipments located at distances d_1 and d_2 (both beyond radio horizon) from a common transmitting system are designated as P_{R1} and P_{R2} then a simple distance law would relate them by:

$$P_{R1}/P_{R2} = (d_2/d_1)^n \qquad (1)$$

where n is the distance exponent.

However, the value of n depends not only on whether the receivers are in the diffraction or forward scatter region, but in each of these regions it also depends upon (i) the complete path geometry which can be described more effectively by angular distance[4] as a single parameter, and (ii) initial refractive index gradient or the Earth's effective radius. For the forward scatter region, the value of n depends also upon (iii) the number of narrow beam widths involved[6] and (iv) the spectrum of fluctuations of refractive index field. Further complications are introduced for transhorizon propagation when partial reflexion becomes an important mechanism. In this case, angular power spectrum depends most critically on the thickness of reflecting layer.[7] The variation of received power with distance in these cases has to be determined from the angular power spectrum based on the geometry of the path and the layers.

The purpose of the present paper is to discuss how distance exponent depends upon the complete geometry and initial refractive index gradient and then present the results of experimental observations on distance exponent together with its variations. These observations were made in the v.h.f. band (120 MHz) over a path length range of 200 km to 700 km (angular distances up to 45 milliradians) for a period of three years. From the results obtained, certain inferences have been made about possible values of the slope of refractive index spectra and scattering parameter, C_n^2.

2 Geometrical Effects

Geometrical effects on distance exponent are most prominent on paths with elevated antenna terminals. For a transhorizon path, this point is clear from the expression of received power which is given by[2] (Fig. 1)
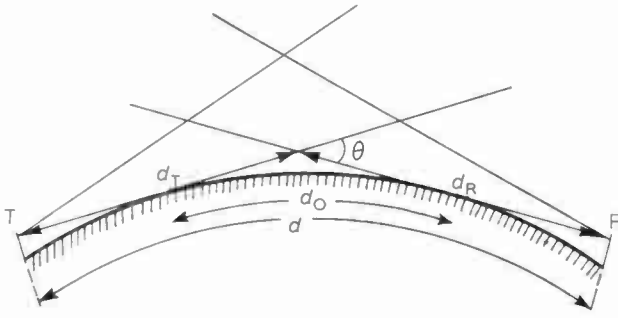
Fig. 1. Schematic path geometry.

$$P_R = P_T \frac{\lambda^2 G_T G_R}{16\pi^2} \int_v \frac{1}{d_T^2 d_R^2} 2\pi k^4 \phi_n(k\theta) \, dv. \qquad (2)$$

where P_R is the received power

P_T is the transmitted power

G_T, G_R are the gains of receiving and transmitting aerials

d_T, d_R are distances from transmitter and receiver to scatterer

$k = 2\pi/\lambda$ is the wave number of electric field

λ is the wavelength of the radio wave

$\phi_n(k\theta)$ is the normalized fluctuation spectra of refractivity

θ is the scattering angle and

dv is the elemental scattering volume.

If, as given by Du Castel,[2] the scattering volume is considered small and the integral is replaced by a product, the received power can be shown as proportional to the following expression:

$$P_R \propto \frac{\theta^{\frac{5}{3}}}{d_T^2} \frac{C_n^2}{d_R^2} \qquad (3)$$

where C_n^2 is a scattering parameter as introduced by Tatarski.[8]

Although the assumptions made above to arrive at expression (3) are oversimplified, the expression nevertheless serves to illustrate the geometrical effects.

Assuming further that average value of C_n^2 varies inversely as the square of the height[4,5] one can approximately write that

$$C_n^2 \propto d_0^{-4} \qquad (4)$$

where d_0 is the distance between the horizons as shown in Fig. 1. Combining (3) and (4)

$$P_R \propto d_0^{-2.66} d_T^{-2} d_R^{-2}. \qquad (5)$$

Thus, if P_{R1} and P_{R2} are the received powers at distances d_1 and d_2 we have, using (5),

$$\frac{P_{R1}}{P_{R2}} = \left(\frac{d_{02}}{d_{01}}\right)^{2.66} \left(\frac{d_{T2}}{d_{T1}}\right)^2 \left(\frac{d_{R2}}{d_{R1}}\right)^2 \qquad (6)$$

where subscripts 1 and 2 are associated with paths d_1 and d_2 respectively.

For a symmetrical geometry, $d_T = d_R \simeq d/2$ so that,

$$\frac{P_{R1}}{P_{R2}} = \left(\frac{d_{02}}{d_{01}}\right)^{2.66} \left(\frac{d_2}{d_1}\right)^4. \qquad (7)$$

Expression (7) will give a simple distance law only if $d_{02}/d_{01} = d_2/d_1$, which is rarely the case. In general, (assuming $d_2 < d_1$) $d_2/d_1 > d_{02}/d_{01}$ so that $(d_{02}/d_{01})^{2.66}$ will be equivalent to $(d_2/d_1)^n$ where n is larger than 2·66. Thus, depending upon the relative values, observed distance index will vary.

Further complications are introduced in case of air/ground geometry where the distance term d for the moving terminal consists essentially (in case of a level flight over smooth terrain) of a constant part and a variable part so that,

$$d_T = d_C + d_V \qquad (8)$$

where C and V signify constant and variable respectively. Thus, $d_T^2 = d_V^2(1 + d_C/d_V)^2$ which is proportional to $\theta^2(1 + d_C/d_V)^2$. For low antenna heights, $d_T^2 \propto \theta^2$. However, if d_C is a large fraction of d_T, this proportionality no longer holds. Taking a specific case where $d_C = 0.7 \, d_T$ it can be shown that if P_R is proportional to d_T^{-n} it is proportional to θ^{-m} such that $n/m \simeq 2.5$. It is quite clear from above that different forms of distance dependence will result purely due to geometrical effects.

Although analytical discussion presented above is related to transhorizon propagation, different path geometries will lead to different distance exponents for the diffraction region as well. This is so because the attenuation for the diffracted field basically consists of four terms, two of which are essentially related to antenna heights, one to path length and another to Earth's constants, polarization and radio frequency.[9]

3 Effect of Initial Gradient of Refractivity

Apart from geometry, the initial gradient of refractivity, unless accounted for, affects the distance exponent differently in different regions.

Expression (3) above can be written slightly differently as

$$P_R \propto \frac{a^{3.66} d_0^{-2.66}}{d_T^2 d_R^2} \qquad (9)$$

and for a symmetrical geometry

$$P_R \propto a^{3.66} d_0^{-2.66} d^{-4} \qquad (10)$$

where a is the Earth's effective radius.

The Earth's effective radius is related to the initial gradient of refractive index by the following expression:

$$a = \frac{R}{1 + R/\rho \frac{d\rho}{dh} \cos\beta} \qquad (11)$$

where R is the geometrical radius, ρ is the refractive index and β is the elevation angle of the radio ray.

If we compare the received power levels at distances d_1 and d_2 (d_1 being greater) in the forward scatter region, the power ratio r_1 for a standard initial gradient say, will be given by:

$$r_1 = \frac{P_{R1}}{P_{R2}} = \left(\frac{d_{02}}{d_{01}}\right)^{2.66} \left(\frac{d_2}{d_1}\right)^4. \qquad (12)$$

The ratio r_2 for the same distances when initial gradient

is higher is given by:

$$r_2 = \left(\frac{d'_{02}}{d'_{01}}\right)^{2 \cdot 66}\left(\frac{d_2}{d_1}\right)^4 . \qquad (13)$$

From (12) and (13),

$$\frac{r_1}{r_2} = \left(\frac{d_{02}}{d'_{02}}\right)^{2 \cdot 66}\left(\frac{d'_{01}}{d_{01}}\right)^{2 \cdot 66} . \qquad (14)$$

Since the first term on right-hand side of equation (14) is larger than the second term, $r_1 > r_2$. Thus when initial gradient ΔN is more negative, the power ratio is reduced and consequently the distance exponent increases. As an illustration about the order of change expected, let us consider path lengths of 715 and 550 km. With typical antenna heights, values of d_{01} and d_{02} are respectively 395 and 230 km when initial gradient of refractivity is -40 N-units/km. Assuming antenna heights unchanged, these values reduce to 325 and 160 km respectively when initial gradient is -80 N-units/ km. Calculated distance exponents in these two cases are approximately 9 and 11, resulting in a 20% change due to the change of initial gradient. Higher fractional change is possible for geometries different from what we have considered here.

4 Details of Experiments

The results of observations on distance exponent presented here were obtained over three air/ground paths in which radio frequency was in the v.h.f. band (118 to 126 MHz). The path lengths ranged from 200 to 700 km covering a normal angular distance of up to 45 milli-radians. Details of the paths are given in Table 1. Path descriptions in the Table include important features of terrain, average horizon distances, total distance covered by line-of-sight and distance between the two radio horizons.

To the author's knowledge, all published results on distance exponent are based on spot measurements or on basis of compilation of data over various paths. However, the results presented here are based on measurements over a period of three years (June 1964 through May 1967) at a large number of points covering a range from 200 to 700 km. The data, therefore, represent the most comprehensive and systematic observations on this problem.

Two sets of observations were made. In one, signal strength for aircraft transmissions was recorded in a ground-based receiving system (so-called air/ground
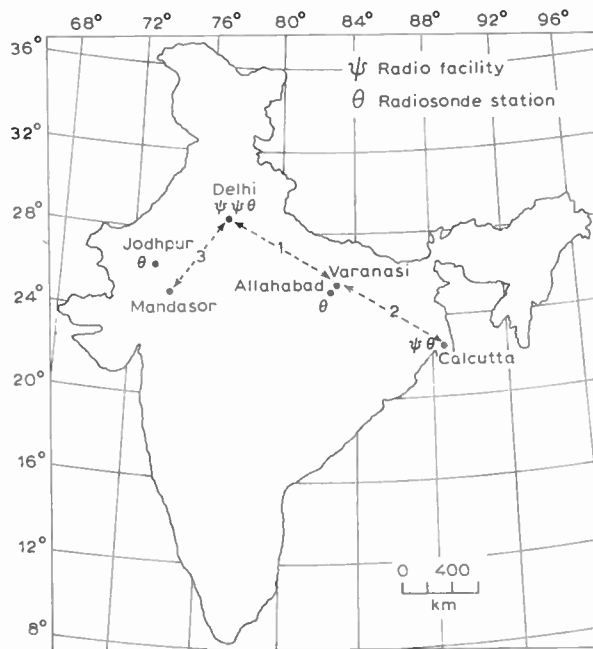

Fig. 2. Radio paths and associated facilities.

Table 1

Details of path geometries

Path designation	Important terrain features	Radio horizon distance from transmitting antenna (km)	Radio horizon distance from receiving antenna (km)	Total line of sight distance (km)	Distance between the radio horizons (km)
1. Delhi/Varanasi air/ground path (airborne transmitter, receiver at Delhi)	terrain fairly smooth, average sea-level altitude 150 m	variable in the range of 280 to 400 km	$33 \pm 3 \cdot 5$	variable in the range of 310 to 430 km	in the range of 0 to 400 km
2. Delhi/Varanasi ground/air path 1 (transmitter at Delhi, airborne receiver)	as above	$33 \pm 3 \cdot 5$	in the range of 130 to 230 km	in the range of 160 to 260 km	in the range of 0 to 300 km
3. Calcutta/Varanasi ground/air air path 2 (transmitter at Calcutta, airborne receiver)	smooth up to 250 km from Calcutta followed by a range of small hills from 250 km to 400 km and smooth thereafter	$22 \cdot 5 \pm 2 \cdot 5$	in the range of 100 to 200 km	in the range of 120 to 220 km	in the range of 0 to 350 km
4. Delhi/Mandasor ground/air path (transmitter at Delhi, airborne receiver)	smooth up to 40 km from Delhi after which small isolated hills were present almost throughout	28 ± 3	in the range of 120 to 220 km	in the range of 145 to 245 km	in the range of 0 to 290 km

path). These observations were made during the period June 1964 through May 1967. The aircraft height was in the range of 5 to 10 km. In the second set, signal due to ground-based transmitters were recorded in a specially-equipped *DC*-3 aircraft flying at level in the height range of 1 to 3 km, on long routes (called here as ground/air paths). Such observations over each path were made for about a year during the period mentioned above. Figure 2 shows in a map the location of radio installations over the paths and also the radiosonde stations, data from which were utilized for determination of refractive index profile characteristics. Radiosonde data were available at several height levels from 900 up to 50 mb including significant levels. The initial gradient of refractive index was determined from surface data and the radiosonde data.

Briefly, the ground based transmitting system consisted of a transmitter with 1 kW output, an antenna system having a gain of 20 dB and a receiving system with a sensitivity of 0.5 µV input. The calibrated a.g.c. of the receiver was put to a recorder. The airborne transmitter power was 25 W and the receiver had a sensitivity of 1.5 µV. The recording equipments had a time-constant of about 1 s.

5 Variation of Distance Exponent with Initial Refractivity Gradient ΔN

Variation of instantaneous received power with distance was converted to distance exponent and the mean value calculated over a period of four hours was designated as n. The dependence of n on ΔN has been studied over the entire distance range of observations. Some of the typical results are presented in three groups corresponding approximately to (i) diffraction region (ii) transition region and (iii) forward scatter region.
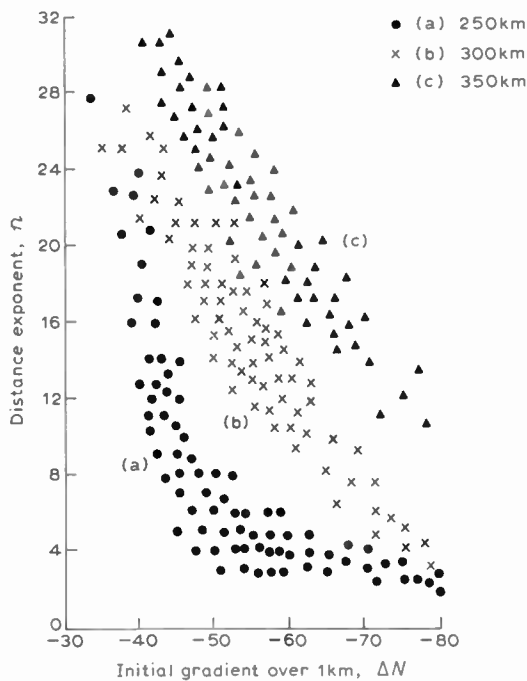

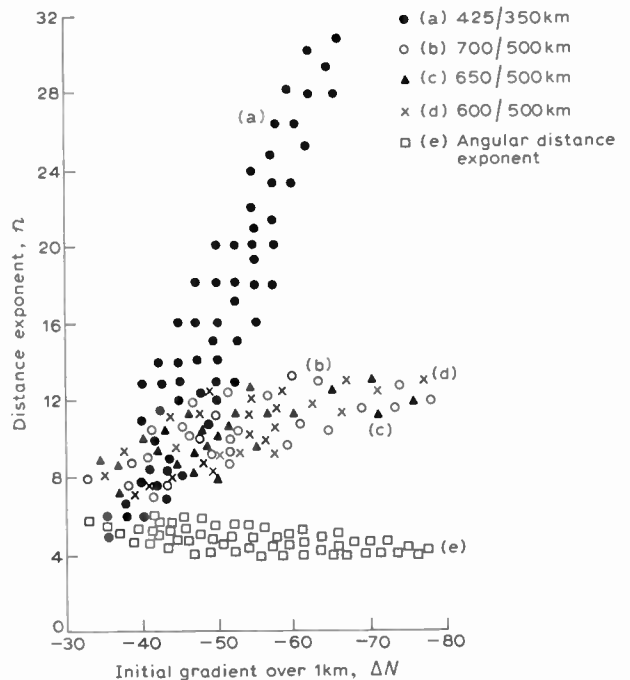**Fig. 3.** Distance exponent for diffraction region.

Fig. 4. Distance exponent for transition and scatter region.

5.1 Diffraction Region

Figure 3 shows the variation of distance index with ΔN for path lengths of up to 350 km. The index was calculated by comparing power at the far point with a reference point normally corresponding to the limit of line of sight. The exponent starts with a very high value for lower values of ΔN and gradually comes down. One observes here that n depends upon the pairs of points which are considered to determine its value, and to this extent, absolute value of n is unpredictable. In all the three cases considered, the exponent decreases with increase of ΔN but the initial rate of decrease is faster for the lower curve corresponding to a path length of 250 km. The rate of fall in the value of n in different regions of the gradient appears to depend upon the distance pairs in question.

5.2 Transition and Forward Scatter Region

A transition zone is considered here to be that region where contributions due to both diffraction and forward scatter are, at times, significant. Curve (a) in Fig. 4 shows index variation for this case. In contrast to the diffraction region, here the exponent n increases with the rise of ΔN. The distance pairs shown in the figure are such that at 350 km diffraction mechanism is dominant under standard gradient conditions while forward scatter is dominant at 425 km. When the gradient value changes from standard to superstandard, signal strength increase at 350 km is relatively more than at 425 km and consequently the observed distance exponent obtained from the ratio of received powers at these points increases. This explains the shape of the curve rising rapidly with ΔN. As in the case of the diffraction region, here again absolute value of n is unpredictable since this critically depends upon the pairs of distances (shown in the diagram) in question.

The lower curve shows the results for the forward scatter region. There are three sets of points on this curve designated as (b), (c) and (d). The distance exponent lies between 7 and 13·0 and does not appear to change with change of position of the farther point. The rise of n with increase of ΔN appears to be asymptotic. The most noteworthy feature in this curve is the high value of n when compared to d^{-7} law. This is purely due to the path geometry and shows the importance of the first term on the right-hand side of equation (6). The wide divergence of the points around the mean curve is to a considerable extent, caused by aircraft height difference for paths of same lengths. The change of n with ΔN is quite significant.

It is concluded from these observations that for diffraction region and near-scatter region, n varies widely with change of initial gradient of refractivity and it also depends critically on pairs of points at which power levels are compared and so the absolute value of n is unpredictable. For the scatter region beyond 25 milliradians, variation in n due to change in ΔN is comparatively less although it is still appreciable. Further, n does not depend upon positions of the pairs of points although for the same distance pairs, divergence is introduced due to difference in aircraft altitude and as a consequence of large antenna heights, the mean distance exponent is much higher than can be explained by homogeneous turbulence theory.

6 The Angular Distance Exponent

From the above results, it is apparent that if the observed distance exponent obtained from measurement of received power levels at two points beyond radio horizon is used for prediction or for testing the validity of a propagation model, effects of path geometry (antenna heights and terrain effects in particular) and initial gradient of refractivity must be taken into account. In fact, the observed distance exponent may be substantially different from that given by a propagation model for forward scatter mechanism due to these effects and more so when one or both of the points in question are situated in the near-scatter region.

Variation of distance exponent due to effects of initial gradient and geometry can be considerably reduced if instead of linear distance angular distance is considered. By this artifice, effects of antenna height and initial gradient are substantially taken into account. The reduction in range of variation with height and initial gradient is shown by the results represented by curve (e) in Fig. 4. The exponent values in this curve pertain to the same geometries and refractivity gradients as in the case of the upper curves (b), (c), (d). It may, therefore, be concluded that angular distance dependence of transmission loss rather than path distance dependence is a better criterion for testing a propagation model.

7 Some Remarks on Refractivity Spectra

The three-dimensional spectral function $\phi_n(k\theta)$ in expression (2) is given, according to homogeneous turbulence theory by:

$$\phi_n(k\theta) = 0.033 \, C_n^2 (k\theta)^{-\frac{11}{3}} \qquad (15)$$

where C_n^2 is a scattering parameter given by:

$$C_n^2 = 5.3 \langle \delta\varepsilon^2 \rangle_{av} \, l_0^{-\frac{1}{3}} \qquad (16)$$

where $\langle \delta\varepsilon^2 \rangle_{av}$ is the mean square value of permittivity fluctuations and l_0 is the input scale, that is, the scale associated with maximum energy density.

Measurements of various characteristic properties of scattered radio wave have shown[10] that the power 11/3 of $k\theta$ in expression (15) which is associated with homogeneous, isotropic turbulence, varies considerably from summer to winter and also changes with geographical position. Gjessing[10] gives the variation of the median value from 3·3 to 5.

Variation of $\langle \delta\varepsilon^2 \rangle_{av}$ with height is quite complicated and as stated earlier, various authors[4,5] have tried different models to fit in with observed radio propagation data. There is considerable amount of divergence even in the reports on experimental values of $\langle \delta\varepsilon^2 \rangle_{av}$. For example, during early experiments carried out in the USSR at heights up to 5 km, no drop in the value of $\langle \delta\varepsilon^2 \rangle_{av}$ with altitude was noted.[3] On the other hand, Borresen and Gjessing[11] have reported considerable decrease of scattering intensity with height, the mean rate being 6 to 8 dB per km. For the higher wave number region, Lane observed that at heights up to 2 km, there was considerable inhomogeneity in the refractive index structure even in non-anticyclonic conditions.[12] He concluded that fluctuations in refractive index with height between 0 and 2 km cannot be represented by a model of $\langle \delta\varepsilon^2 \rangle$ decreasing steadily with height.

An attempt has been made here to derive the average value of m in the expression $\phi_n(K) \propto K^{-m}$ where $K(\simeq k\theta)$ is the spatial wave-number, under the following assumptions:

(i) direction of K is essentially vertical,

(ii) spectral slope m does not change over the wave number range in our experiments, i.e. 0·03 m^{-1} to 0·06 m^{-1} approximately, and

(iii) there is no appreciable variation in scattering intensity with change of horizontal position of scattering volume as observed by Borresen and Gjessing.[11]

With these assumptions, whenever transmission loss changes with aircraft position, it can be ascribed to change of θ and height of scattering volume. In our experiment, both of them change simultaneously. Following Gordon,[13] the scattered power can be expressed by converting the volume integral in (2) as a height integral giving:

$$\frac{P_R}{P_{FS}} = pd^2 \int_0^\infty \frac{\sigma(0,\chi)}{d_T^2 d_R^2} \frac{4}{3} d_0 \, h \, dh \qquad (17)$$

where P_R is the forward scatter power

P_{FS} is the power that would be received under free space condition with same terminal equipments at same distance

p is the factor accounting for ground reflexion effects

$\sigma(\theta,\chi) = 2\pi k^4 \phi_n(k\theta)$ is the scattering cross-section, being the scattered power per

unit power density of incident wave in the scattering volume per unit solid angle in the direction of receiver from scatterer, per unit scattering volume

d_0 is the distance between the two radio horizons and

h is the height of elementary scattering volume.

Finally expression (17) can be rewritten as

$$\frac{P}{P_{FS}} \simeq 43\cdot8 \frac{d^2 d_0 b}{d_T^2 d_R^2} \int_0^\infty (h_0 + h)^{-n'} \left[\frac{d_0}{10\cdot04(h+2h_0)}\right]^m h \, dh \tag{18}$$

where m is the spectral slope,

n' is the height exponent for C_n^2

and b is a constant associated with C_n^2 which also ensures dimensional adjustments. In deriving expression (18), we have taken d_T and d_R as essentially constant for one geometry.

In expression (18) a value of 10^{-6} can be assigned to the constant b without affecting the forms of terms inside the integral. There are now two unknowns n' and m in (18) and from the data, it is not possible to deduce their values separately. However, by assuming discrete values for m, the author has determined a set of values for n' and consequently C_n^2. Table 2 shows the results for two months, namely, August (higher value in Table) and April (lower value in Table) which correspond respectively to the most favourable and unfavourable months for propagation. Results therefore emphasize the extremes of variations as well. The unit of C_n^2 here is in appropriate power of centimetres, the value of power being decided by slope of spectra.

These sets of values of C_n^2 can be compared with the values obtained or quoted by other workers.

(i) Crain[14] evaluated $\Delta n^2/l$ at different heights and values of this parameter varied from 6×10^{-13} to 10^{-15} per metre between 5000 and 10 000 ft of height.

(ii) Norton, Rice and Vogler[4] estimated from Cheyenne Mountain radio data values of $\Delta n^2/l$ to lie between 10^{-14} to 4×10^{-15} (per metre) for heights from 1 to 2 km. Their estimate for transmission paths in remainder of USA was 8×10^{-14} to 2×10^{-14} for the above height range.

(iii) Lane[12] indicated that when $\Delta n^2/l$ was unusually large, its value was about 10^{-11} per metre in the above height range. Normal values would be several orders lower than this.

(iv) Tatarski[8] evaluated C_n at $1\cdot5$ km altitude and his estimate gave $C_n = 8 \times 10^{-8}$ to $4\cdot5 \times 10^{-9}$ $(cm)^{-1/3}$.

(v) Gjessing[10] considered a value of $C_n^2 = 10^{-12}$ $(cm)^{-2/3}$ to be typical.

(vi) Clifford and Strohbehn[15] quoted $C_n^2 = 4 \times 10^{-14}(m)^{-2/3}$ for calculating optical variance.

Values of C_n^2 in the range of 10^{-13} to 10^{-14} around a height of 1 km would therefore appear to be consistent with the above observations. In our case, this order of values for C_n^2 would indicate a spectral slope of $4\cdot5$ to 5 in this wave-number range (0·03 to 0·06 m^{-1}) and is therefore higher than that given by homogeneous turbulence theory.

8 Conclusions

It is concluded that the distance exponent obtained by comparing received power levels at two points in diffraction region may vary from 2 to more than 30 depending upon the initial gradient and relative positions of the pairs of points. In the forward scatter region, the exponent has been found to vary from 7 to about 13

Table 2

Mean values of C_n^2 at various heights for several assumed values of spectral slope m

| Height in metres | Slope m | | | | |
	3·00	3·66	4·00	4·50	5·00
500	$5\cdot47 \times 10^{-11}$	$1\cdot48 \times 10^{-11}$	$7\cdot04 \times 10^{-12}$	$2\cdot30 \times 10^{-12}$	$7\cdot00 \times 10^{-13}$
	$6\cdot21 \times 10^{-12}$	$1\cdot58 \times 10^{-12}$	$7\cdot20 \times 10^{-13}$	$2\cdot26 \times 10^{-13}$	$6\cdot86 \times 10^{-14}$
707	$3\cdot33 \times 10^{-11}$	$9\cdot46 \times 10^{-12}$	$4\cdot90 \times 10^{-12}$	$1\cdot71 \times 10^{-12}$	$5\cdot80 \times 10^{-13}$
	$3\cdot60 \times 10^{-12}$	$9\cdot51 \times 10^{-13}$	$4\cdot72 \times 10^{-13}$	$1\cdot62 \times 10^{-13}$	$5\cdot30 \times 10^{-14}$
945	$2\cdot00 \times 10^{-11}$	$6\cdot48 \times 10^{-12}$	$3\cdot30 \times 10^{-12}$	$1\cdot25 \times 10^{-12}$	$4\cdot18 \times 10^{-13}$
	$1\cdot79 \times 10^{-12}$	$5\cdot50 \times 10^{-13}$	$2\cdot70 \times 10^{-13}$	$9\cdot86 \times 10^{-14}$	$3\cdot35 \times 10^{-14}$
1208	$1\cdot23 \times 10^{-11}$	$4\cdot09 \times 10^{-12}$	$2\cdot30 \times 10^{-12}$	$8\cdot83 \times 10^{-13}$	$3\cdot41 \times 10^{-13}$
	$9\cdot00 \times 10^{-13}$	$2\cdot76 \times 10^{-13}$	$1\cdot50 \times 10^{-13}$	$5\cdot78 \times 10^{-14}$	$2\cdot14 \times 10^{-14}$
1514	$6\cdot96 \times 10^{-12}$	$2\cdot44 \times 10^{-12}$	$1\cdot40 \times 10^{-12}$	$5\cdot76 \times 10^{-13}$	$2\cdot40 \times 10^{-13}$
	$4\cdot83 \times 10^{-13}$	$1\cdot62 \times 10^{-13}$	$9\cdot00 \times 10^{-14}$	$3\cdot53 \times 10^{-14}$	$1\cdot45 \times 10^{-14}$
1760	$4\cdot29 \times 10^{-12}$	$1\cdot59 \times 10^{-12}$	$9\cdot00 \times 10^{-13}$	$3\cdot84 \times 10^{-13}$	$1\cdot60 \times 10^{-13}$
	$2\cdot98 \times 10^{-13}$	$1\cdot00 \times 10^{-13}$	$5\cdot70 \times 10^{-14}$	$2\cdot36 \times 10^{-14}$	$9\cdot86 \times 10^{-15}$
2000	$3\cdot14 \times 10^{-12}$	$1\cdot14 \times 10^{-12}$	$6\cdot87 \times 10^{-13}$	$3\cdot00 \times 10^{-13}$	$1\cdot26 \times 10^{-13}$
	$2\cdot00 \times 10^{-13}$	$7\cdot42 \times 10^{-14}$	$4\cdot10 \times 10^{-14}$	$1\cdot75 \times 10^{-14}$	$7\cdot24 \times 10^{-15}$

depending upon the initial gradient and path geometry. It has been shown analytically that antenna height and Earth's effective radius can change the observed distance index by a large amount. Since in the forward scatter region, the distance exponent and form of turbulent spectra are related, it is essential to take into account these effects of path geometry and refraction to arrive at the correct value of distance exponent. It is also shown that for asymmetrical paths subject to wide variation in initial gradient, angular distance exponent is subject to much less variation. Above conclusions are based on transmission loss measurements over three paths varying in length from 200 to 700 km over a period of three years.

For small wave-number range (0·03 to 0·06 m^{-1}) in the present case, the form of spectrum is not well known.[15] If, however, homogeneous turbulence theory is assumed to be applicable, it is found that higher values (4·5 to 5·0) of spectral slope m in the relation $\phi(k) \propto K^{-m}$ lead to more realistic order of values for average value of scattering parameter C_n^2.

9 Acknowledgments

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Since 1964, Mr. Majumdar has been engaged in research on tropospheric propagation beyond the radio horizon and has already published seven papers on this subject. He is currently employed as a Scientist in the National Physical Laboratory of India and is working on microwave propagation phenomena and related characteristics of the medium.

Improved function generation in pipe flow simulators

R. W. WILLIAMS, Ph.D., F.Inst.P. *

1 Introduction

In the direct analogue technique for the study of fluid distribution networks, a 'pipe cell' is a two-terminal electronic circuit which behaves as a non-linear resistance whose voltage-current law is an analogue of the pressure/flow law in a pipe. Thus the pipe cell is required to obey the law $E = kI^n$ where E is the potential difference across the cell and I is the current through it. In a practical cell, the pipe law should be followed reasonably accurately, say to within 1% of full scale, and a wide range of variation of the hydraulic resistance parameter k should be accommodated. The cell action should be bidirectional, i.e. when the polarity of E reverses, the direction of the current flow should also reverse, automatically. A previous paper[†] has described such a cell, based on the use of an integrated circuit operational amplifier in combination with a law-forming network of resistors and biased diodes. Further work has resulted in the simplification of the function-generating aspects of the cell and a new design is now described.

2 The Function-generating Requirement

The action of the previous cell depends on the use of a function—generating network of resistors and biased diodes designed to follow precisely the same law as that of the cell as a whole, namely $E \propto i^n$, where E is the cell voltage and i is the current drawn by the network. The approach now adopted is based on the generation of a law differing from that of the cell itself and it may be explained with the aid of Fig. 1. The cell voltage is applied to the input terminals of the 'function generator'.
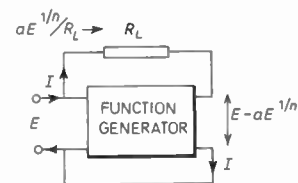

Fig. 1. Principle of new cell.

This generator is designed to yield an output voltage of the form $E - aE^{1/n}$, where a is a constant. The input impedance of the function generator is assumed for practical purposes to be infinite and its output impedance is taken to be negligibly small. As described thus far, the device draws no current from the source E and a cell current does not as yet exist. But if a resistor R_L is connected between the input and output terminals of the generator as shown in Fig. 1, then the action of the function generator maintains a voltage $E - (E - aE^{1/n})$, i.e., $aE^{1/n}$ across R_L. In consequence a current I, equal to $aE^{1/n}/R_L$, flows through R_L and returns to the source E via the output stage of the function generator, provided that this is capable of sinking it. I thus constitutes the cell current and the cell law is $E = (R_L/a)^n . I^n$.

† Williams, R. W., 'Bidirectional action in pipe flow simulators', *The Radio and Electronic Engineer*, **41**, No. 11, p. 490, November, 1971; British Aircraft Corporation Limited and Williams, R. W., British Patent 1,289,062.

SUMMARY

The function-generating requirements in an existing design of 'pipe cell', which simulates the hydraulic behaviour of pipes, are reviewed and a simpler method of meeting them is explained. A cell design embodying the new approach is described.

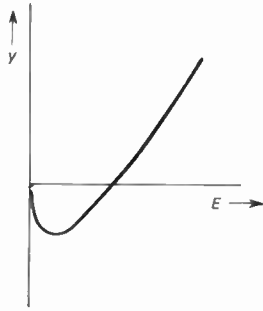* British Aircraft Corporation, Stevenage, Hertfordshire.

Fig. 2. General form of the function $y = E - aE^{1/n}$.

The circuit of Fig. 1 has two key features:

(i) The function-generating task is substantially eased by changing the form of the function from an nth power law to the expression $E - aE^{1/n}$,

(ii) The bypass resistor R_L provides a path for cell current to flow and it serves as the means of controlling the value of the hydraulic resistance simulated by the cell, which is $(R_L/a)^n$.

2.1   The Nature of the Function $E - aE^{1/n}$

The function $y = E - aE^{1/n}$ is characterized (Fig. 2) by a dip near the origin in which y attains a peak negative value, followed by a more or less linear portion of the curve for the remainder of the range of E. If the constant a is suitably chosen, the peak negative value of y becomes very small indeed. Figure 3 is a plot of y for $a = 0.600$ and $n = 1.85$ over the range 0–13 V of E. It will be apparent that the dip has virtually disappeared, and the three-segment piecewise linear approximation OABC which has been drawn on Fig. 3 is clearly an accurate realization of the function y. The range of E chosen for Fig. 3 was selected in the light of the known operating range of linear integrated circuits such as the 741 operational amplifier. The value of 0·6 for a is the result of a relatively crude optimization in which y was plotted for various values of a (n being kept at 1·85) and the best outcome selected in terms of smallness of the negative
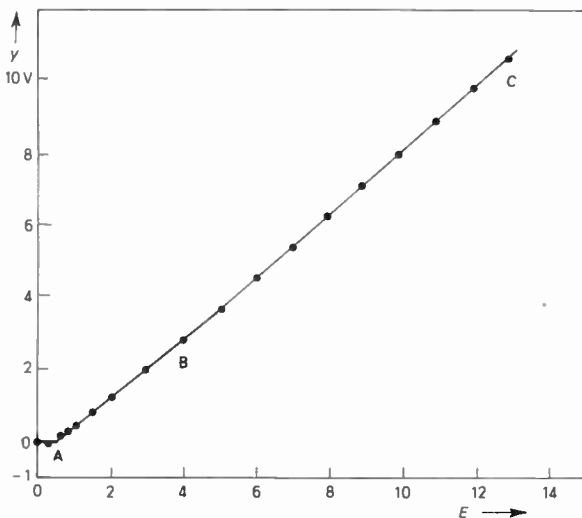

Fig. 3. Plot of y for $a = 0.600$ and $n = 1.85$.

peak value of y combined with the degree of linearity of the major portion of the plot.

2.2   Realization of the Function

The three segments of the approximation OABC in Fig. 3 are readily realized without having to resort to a conventional resistor and biased diode law-forming network. The breakpoint which terminates the horizontal segment OA occurs at a voltage of about 0·6 V. This is the threshold voltage of a silicon diode; Bellamy and West† state that the type 1N916 diode is equivalent to a source of 0·59 V in series with 30 Ω for currents greater than 1 mA. The two segments OA, AB may therefore be generated by an operational amplifier arrangement in which the input arm is a silicon diode in series with a resistor.

Referring to Fig. 4, the cell voltage E is first buffered by the operational amplifier A1 used as a non-inverting voltage follower. The function generation is performed with the aid of the summing inverting operational amplifier A2, the input arm of which comprises the diode D1 in series with R1. Generation of the second segment AB starts only when the output of A1 is great enough to overcome the threshold of D1, and the slope of AB is determined by the value of the feedback arm R2 in relation to R1. To secure clean generation of the first
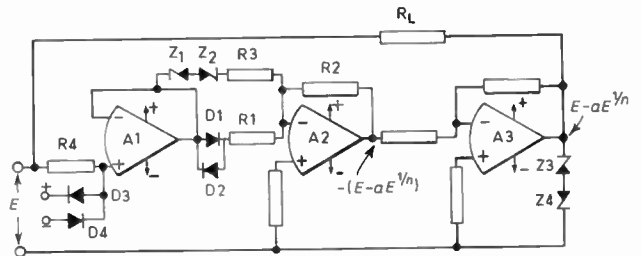

Fig. 4. Circuit diagram of new cell.

breakpoint, the current through R1 should be of the order of milliamperes, hence the use of the buffer A1. In the absence of A1, the current through R1 would form an appreciable fraction of the total current through the cell and so result in a substantial error in the realization of the cell law.

The second breakpoint occurs at a voltage of about 5·3 V and the segment BC may therefore be generated by using a second input arm R3 in shunt with the main arm $D_1 R_1$. A Zener diode Z1 of appropriate Zener voltage value is in series with R3; when the output of A1 exceeds the Zener value, current flows through R3 and BC is generated, its slope being determined by the value of R3 in relation to R1 and R2.

The output of A2 is thus seen to be the desired approximation OABC to the function y of Fig. 3. It is, however, of reversed polarity and the cell circuit must therefore be extended by the addition of the inverting amplifier A3, the input and feedback arms of which are equal so that its gain overall is unity. Finally, the cell circuit is com-

† Bellamy, N. W. and West, M. J., 'A high-accuracy four-quadrant time-division analogue multiplier', *Electronic Engineering*, 42, No. 509, p. 30, July 1970.

pleted by the connexion of the bypass resistor R_L from the input terminal of the cell to the output of A3.

As described, the cell functions only for one polarity of the cell voltage E, which has to be positive with respect to the common line in the circuit of Fig. 4. Automatic reversal of the cell action is secured simply by the addition of a second 'threshold' diode D2 connected in opposition across D1 and a second Zener diode Z2 in series opposition to Z1.

The input voltage of A1 is prevented from exceeding the safe limits by the addition of the series resistor R4 and the diode clamp D3, D4. The clamping voltages for D3 and D4 are derived within the cell power supply. The output of A3 and the cell power supply are protected by the connexion of a pair of Zener diodes Z3, Z4 (11 V), back-to-back between the output terminal and the common line. When the voltage applied to the cell exceeds its maximum working voltage, the Zener diodes divert the bypass current, which now rises above its normal value, to the common line and prevent its passage through A3 and the power supply.

3 Cell Performance

The circuit of Fig. 4 has been constructed and found to work well using type 741C operational amplifiers, the $+15$ V and -15 V rails for which were provided by a simple Zener-stabilized supply. (Since the cell has to be a fully floating two-terminal device, it must have its own self-contained power unit.) Z1 and Z2 were $5 \cdot 6$ V Zener diodes; R1 and R3 were each made up of a fixed resistor in series with a pre-set which was adjusted during test to obtain the closest approximation overall to the desired law.

For a fixed value of R_L, the value of E for a given value of I followed the $1 \cdot 85$ power law within the limits of $\pm 1 \%$ of full scale over the range 0–13 V of E. When the polarity of E was reversed at various values of E, the consequent changes in the value of I did not exceed $\pm 0 \cdot 4 \%$ of full scale. Regarding the hydraulic resistance range, the input bias current drawn by A1 is negligible in comparison with the cell current and in practical terms the maximum hydraulic resistance is determined by the maximum value of the bypass resistor R_L. In the earlier cell, the current drawn by the law-forming network imposed the limit on the maximum realizable resistance. The minimum hydraulic resistance is determined by the maximum current which the output stage of A3 can accept as a sink. The experimental cell had a hydraulic resistance range of 56 000 : 1, over which range there was no appreciable variation in the accuracy of the cell law.

4 Conclusions

A review of the function-generating requirement in the design of pipe flow simulators has led to a fresh approach, resulting in an easement of the requirement and a simplification of the function-generating circuit. It is true that the new circuit employs two additional integrated circuit operational amplifiers but the overall improvement in performance, including the enlargement of the hydraulic resistance range, is considered to be worthwhile.

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Dr. R. W. Williams has been local head of research at the Stevenage Works of the British Aircraft Corporation for the past five years. He joined the company in 1949 and after several years working on guided missile systems moved to the Industrial Products Group. He obtained his first and second degrees from the University of Wales. Dr. Williams has also written several papers including one in the Institution's *Journal* and he is the author of a book 'Analogue Computation'. A fuller note on his career was published in November, 1970.

Monolithic voltage-ratio-circuit performance limits

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and

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SUMMARY

An expression, involving the degree of mismatch in the respective parameters of similar integrated circuit components, operated at similar current levels, is presented for the error limits of a new voltage-ratio-circuit having a small source loading and a low output resistance. Using currently available s.i.c. process technology, an output error < 1% is readily achieved. Experimental results with a matched transistor array indicate a performance superior to this for a typical case.

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1 Introduction

Certain circuit families, e.g., voltage regulators, analogue/digital converters and emitter-coupled logic, require temperature-independent reference sources for voltage level comparison purposes. Such sources are normally derived, via a divider network, from a Zener diode reference unit or the rail supply voltage. The former case applies particularly to timing circuits, employing a comparator, in which the output pulse duration is required to be sensibly independent of rail supply tolerances and variations.

One class of temperature-compensated divider network, previously described,[1] is shown in Fig. 1. If I_B is regarded as negligible compared with I_1, then $I_1 \simeq I_2$ and $(V_0/V_i) = (m-1)/(m+1)$, for $(R_1/R_2) = (n+1)/(m-1)$. The circuit is flexible in application and its performance in monolithic form quite satisfactory.[2] However, it can be improved if the effect of non-zero I_B is cancelled or partially-cancelled. In a scheme already considered,[3] a discrete p-n-p device was used for this purpose, but complementary transistors with high, matching, β values are not conveniently realized in an n-p-n-optimized monolithic process technology. Although this problem can be minimized by increased circuit complexity, an 'all-n-p-n' transistor scheme is preferable.

This paper discusses the fundamental limitations of one such scheme, shown in Fig. 2 for which (I_1/I_2) is unity—
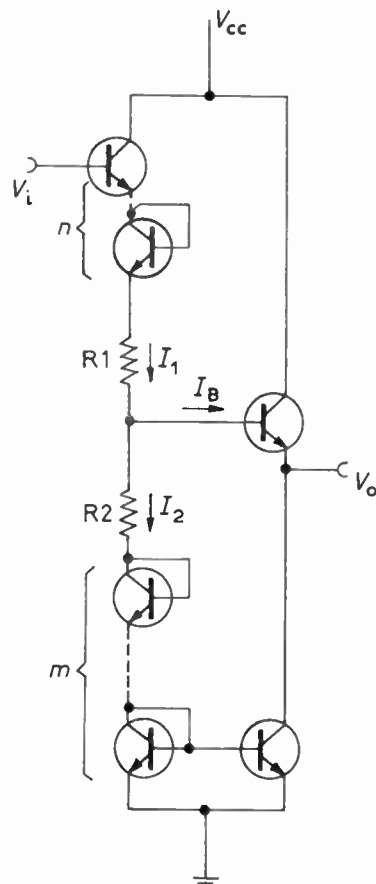

Fig. 1. Previously described voltage-divider network.[1]

to an accuracy well within the tolerance currently obtainable on the ratio of adjacent monolithic resistors—and $(V_0/V_i) = (m+1)/(m+n+2)$, for the design choice, $(R_1/R_2) = (n+1)/(m+1)$.

2 Theory

Assume, initially, that the transistors have a common value, δ, for their base-emitter voltage drops. Then, from Fig. 2,

$$V_0 = V_i - (n+1)\delta - I_1 R_1 \qquad (1)$$

$$V_0 = (m+1)\delta + I_2 R_2. \qquad (2)$$

Let

$$I_1 = \eta I_2 \qquad (3)$$

and

$$y = \eta(R_1/R_2) \qquad (4)$$

where η is a numeric, shown later to be very close to unity and sensibly independent of I_1. Manipulation of (1) through (4) yields

$$V_0 = [V_i + (m+1)y\delta - (n+1)\delta]/(1+y). \qquad (5)$$

If, by design

$$(m+1)y = (n+1) \qquad (6)$$

then

$$V_0 = V_i/(1+y) = V_i(m+1)/(m+n+2)$$
$$\neq f(I, \text{temperature}). \qquad (7)$$

For the design choice (6), inspection of (5) indicates that the worst case error ΔV_0, in V_0 due to a maximum difference, $\Delta\delta$, in transistor V_{BE} matching is given by

$$(\Delta V_0/V_0)_1 = (n+1)\Delta\delta/V_i. \qquad (8)$$

Similarly, the error resulting from an uncertainty, Δy, in y is

$$(\Delta V_0/V_0)_2 = (\Delta y/y) \times \{[(n+1)\delta/V_i] - [(n+1)/(m+n+2)]\} \qquad (9)$$

The total fractional error, $(\Delta V_0/V_0)$, is the algebraic sum of $(\Delta V_0/V_0)_1$ and $(\Delta V_0/V_0)_2$.

Considering the definitions of y and η given by (3) and (4), we have

$$(\Delta y/y) = [\text{fractional error in } (R_1/R_2)] + (\Delta\eta/\eta). \quad (10)$$

To calculate $(\Delta\eta/\eta)$, we proceed as follows: From Fig. 2,

$$I_1 - I_2 = I_B - I_4. \qquad (11)$$

Furthermore,[4]

$$I_3 = I_2[1 - (2/\beta_6) \pm (\Delta\delta/V_T)] \qquad (12)$$

where β_6 is the common-emitter d.c. current gain of Tr6, $V_T = $ thermal voltage $\simeq 26$ mV at $T = 300$ K, and $\Delta\delta$ here relates to the V_{BE} mismatch in the 'current-mirror' transistors Tr5, Tr6.

Using appropriate transistor number subscripts for β and common-base d.c. current gain, α, it follows from (11) and (12) that,

$$I_1 = I_2\{\alpha_3 + (1-\alpha_7)$$
$$\times [1 - (2/\beta_6) + (1/\overline{\beta_3+1}) \pm (\Delta\delta/V_T)]\}. \qquad (13)$$
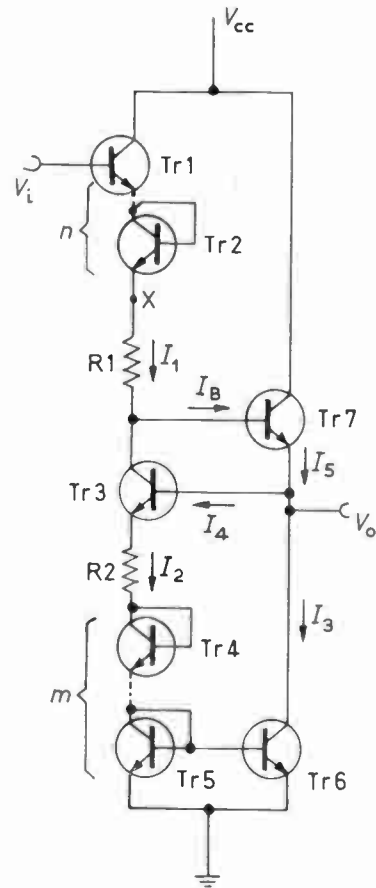

Fig. 2. New voltage-divider network.

For experimental test, Tr2, Tr4 omitted. Tr3, Tr5, Tr6, Tr7 comprise part of a SL 3145 (Plessey), Tr1 a selected unit of a second SL 3145 array, $R_1 = R_2 = 2.2$ kΩ (nominal), matched to within 1 part in 10^4, $V_{cc} = 12$ V.

In view of the order of the approximations involved it is legitimate to write

$$(\alpha_3 - \alpha_7) = \pm \Delta\alpha \simeq \pm(1/\beta)(\Delta\beta/\beta) \qquad (14)$$

$$(1 - \alpha_7) = (1 - \alpha) \simeq (1/\beta) \qquad (15)$$

$$(1/\beta_6) \simeq 1/(\beta_3 + 1) \simeq (1/\beta). \qquad (16)$$

Thus, (13) reduces to

$$(I_1/I_2) \simeq [1 - (1/\beta^2) \pm 1/\beta\{(\Delta\beta/\beta) + (\Delta\delta/V_T)\}]. \quad (17)$$

With the typical values of β, $\Delta\delta$, considered later, there is no significant error in assuming that,

$$|(1/\beta)| \ll |\Delta\beta/\beta| + |\Delta\delta/V_T|. \qquad (18)$$

For this case,

$$(I_1/I_2) \simeq [1 \pm (1/\beta)\{(\Delta\beta/\beta) + (\Delta\delta/V_T)\}]. \qquad (19)$$

Clearly,

$$(I_1/I_2) = \eta = 1 \text{ (nominal)}. \qquad (20)$$

The fractional uncertainty in η due to mismatches in β, δ is thus,

$$\Delta\eta/\eta = \Delta\eta = \pm(1/\beta)\{(\Delta\beta/\beta) + (\Delta\delta/V_T)\}. \qquad (21)$$

The total fractional error in V_0 is thus simply calculable, using (8), (9) (10) and (21), from a knowledge of the process parameters β, $(\Delta\beta/\beta)$, tolerance on (R_1/R_2), δ,

$\Delta\delta$, and the circuit choice for V_i and y (and hence n, m).

Table 1

Process	β	$\frac{\Delta\beta}{\beta}$	(R_1/R_2) match	δ (V)	$\Delta\delta$ (mV)	$\frac{\Delta\eta}{\eta}$	$\left(\frac{\Delta V_0}{V_0}\right)_1$	$\left(\frac{\Delta V_0}{V_0}\right)_2$
		%	%			%	%	%
A	40	10	3	0.75	2.5	0.25	0.083	1.6
B	75	5	1.5	0.75	1	0.067	0.033	0.78

Table 1 gives the magnitudes of acceptable yield parameters and their percentage variations for two s.i.c. processes, A and B, together with $\Delta V_0/V_0$ components calculated from (8) (9), (21) for $y = 1$, i.e. $(V_0/V_i) = 0.5$ (nominal), with $n = m = 1$, for the case $V_i = 6V$. This is a typical system value in digital-analogue conversion and can be obtained in practice by the use of a temperature-compensated Zener diode—formed from a forward biased p-n junction in series with the reverse biased base-emitter junction of a transistor operating in the breakdown region—driven by a constant current source to isolate possible V_{cc} variations.

For generality, A and B do not refer to one manufacturer's performance capabilities: A is intended to be indicative of what has been possible for some years, whilst B is meant to be more representative of current processing technology.[5] The conclusion from Table 1 is that $(\Delta\eta/\eta)$ contributes little to $(\Delta V_0/V_0)_2$ in comparison with the tolerance on R_1/R_2.

3 Results

Tests were made on a circuit having $(V_0/V_i) = 0.5$ (nominal) i.e. $n = m = 1$, and $R_1 = R_2$, constructed from a monolithic bipolar array (Plessey SL 3145). To isolate, and check the contribution of $(\Delta\eta/\eta)$ to $(\Delta V_0/V_0)_2$, a test was devised to measure η for $0.2\,mA < I_1 < 2\,mA$, this being considered a useful current range for practical applications. The upper limit represents a compromise between minimum input current and minimum resistor size. The input current, I_{in}, of Tr1 is given by $I_{in} \simeq (V_i - 4\delta)/2\beta R_1$, for $R_1 = R_2$. If $I_{in} = 50\,\mu A$, corresponding to $I_1 = 2\,mA$ and $\beta = 40$, then $V_i(max) = 12\,V$ implies $R_1 \simeq 2\,k\Omega$. This is a reasonable size for monolithic circuit realization; a smaller I_{in} is obviously possible at the expense of chip area. Accordingly, two discrete $2.2\,k\Omega$ (nominal) resistors were selected to match, on a measurement bridge, to within 1 part in 10^4 for use as R_1 and R_2.

Furthermore, matched $1.5\,k\Omega$ (nominal) 'current viewing' resistors were inserted in series with the bases of Tr7 and Tr3. This is necessary because $\Delta\eta$ is small and a determination of it from the subtraction of two relatively large quantities (I_1, I_2) would give unacceptable error.

From (3), (11) and the interpretation of $\Delta\eta$ as the amount by which η departs from unity, it is seen that

$$(\Delta\eta/\eta) = \Delta\eta = (I_B - I_4)/(I_1 + I_B - I_4) \qquad (22)$$

or,

$$(\Delta\eta/\eta) \simeq (I_B - I_4)/I_1. \qquad (23)$$

The test procedure was as follows: with Tr1 omitted, an input voltage, V_x, from a stable, variable, power supply was applied at point X (Fig. 2) and I_1, I_B, I_4 were measured by observing, with a digital voltmeter, the voltage drop across the appropriate monitoring resistors. Results are given in Table 2.

Table 2

V_x (V)	3	4	5	6	7
I_1 (mA)	0.207	0.420	0.643	0.862	1.08
$(\Delta\eta/\eta)$ %	0.39	0.29	0.26	0.25	0.24

V_x (V)	8	9	10	11	12
$(I_1$ (mA)	1.30	1.51	1.72	1.95	2.17
$\Delta\eta/\eta$ %	0.24	0.25	0.25	0.28	0.30

Tests were then made to determine the total error $(\Delta V_0/V_0)$ with Tr1 in circuit and the resistors inserted to monitor I_B and I_4 shorted out. The transistor array incorporating Tr3, 5, 6 and 7, is fabricated in such a way that it is not possible (because of substrate earthing problems) to utilize the remaining transistor for Tr1. It was necessary, therefore, to select a transistor from a similar array, so that its V_{BE} matched the average V_{BE} of Tr3, 5, 6, and 7, to within a few millivolts.

Table 3 gives the measured $(\Delta V_0/V_0)$ as a function of V_i.

Table 3

V_i (V)	3.0	4.1	5.1	6.1	7.1
$\frac{\Delta V_0}{V_0}$ %	+0.66	+0.21	+0.12	+0.072	+0.044

V_i (V)	8.1	9.1	10.1	11	12
$\frac{\Delta V_0}{V_0}$ %	+0.021	*	*	−0.036	−0.062

* Indicates error within matching of R_1, R_2, i.e. within 1 part in 10^4.)

4 Discussion

Table 2 shows that (i), $(\Delta\eta/\eta)$ is sensibly independent of I_1 over a useful range of currents; this implies small variation of β with current and good tracking, and (ii), the matching of R_1 and R_2 will be the main contributor to the error term $(\Delta V_0/V_0)_2$ in a monolithic realization of Fig. 2. However, the value of 0.25% for $(\Delta\eta/\eta)$ is unexpectedly high since the SL 3145 is a recent device type. On checking, it was found that for the particular sample used, at $V_x = 6V$, $\beta_7 \simeq 46$, $\beta_3 \simeq 41$, i.e. a difference of approximately 10%. The parameters are close to those listed for Process A in Table 1 and the figure quoted for $(\Delta\eta/\eta)$ in the Table (0.25%) agrees favourably with the experimental figure. The worsening of $(\Delta\eta/\eta)$ at higher voltages could be attributed to the

modest collector output resistance of the high f_T devices used.

The data in Table 3 are evidence of (i) the partial cancellation of the curly-bracketed terms, both of which are positive, on the right-hand side of (9), and (ii) the partial cancellation of $(\Delta V_0/V_0)_1$ and $(\Delta V_0/V_0)_2$.

Improved circuit performance is possible if Tr3, Tr7, Tr5, Tr6 are arranged to operate at similar collector-emitter potentials. This could be achieved by the use of amplified-diode structures[6] acting as coarse Zener diodes, which could be inserted in series with the collectors of Tr6 and Tr7 to make the collector-emitter potentials of these transistors nearer that of Tr3, Tr5. This modification would be suitable only for a limited range of V_i values, since Tr6 and Tr7 would saturate for large input voltage changes.

One restriction in the analysis was the inequality (18), the validity of which falls for the case of very close transistor matching; e.g. $(\Delta\beta/\beta) = 2\%$, with $\beta = 72$, and $\Delta\delta = 0.25$ mV. However, for such a situation, $(\Delta\eta/\eta)$ as calculated from (21) has a value which is orders of magnitude better than currently achievable tolerances on (R_1/R_2).

Equation (20) is important practically because it means $y = 1$, and the ratio (R_1/R_2) is required to be a simple, exact fraction.

Only the case $(V_0/V_i) = 0.5$ (nominal), corresponding to $m = n = 1$, has been considered in detail, theoretically and experimentally, to establish the validity of the approach. Since, in the design technique discussed,

$$(V_0/V_i) = (m+1)/(m+n+2)$$

and

$$(R_1/R_2) = (n+1)/(m+1),$$

a table is easily drawn up to give guidance on the choice of minimum m and n for a given voltage division ratio. Equations (8), (9), (10) and (21) then permit a calculation of $(\Delta V_0/V_0)$ for the arrangement adopted.

In conclusion, the technique described is amenable to certain variations. Thus in Fig. 3, the 'Wilson current mirror'[7]—comprising the three bottom transistors—is used to obtain $(V_0/V_i) = 0.5$. The accuracy of the configuration can be shown to be virtually the same as that discussed above, but it does require an additional transistor.
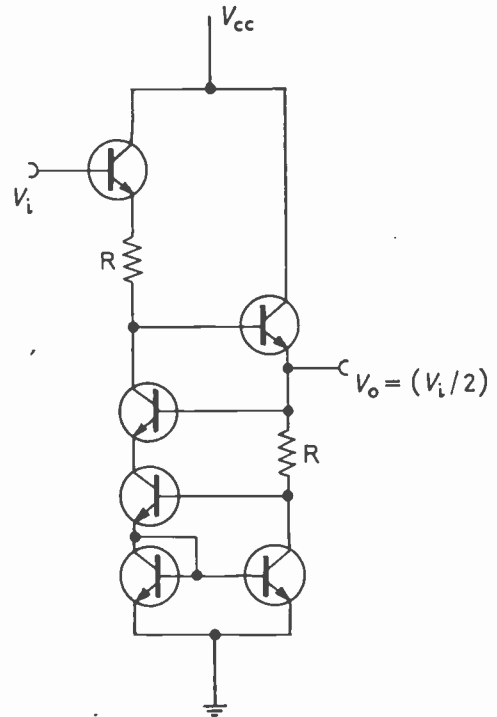

Fig. 3. Circuit variant for $(V_0/V_i) = 0.5$.

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Maximum tolerable local oscillator noise in a 12-GHz f.m. satellite television receiver

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SUMMARY

When considering the design of receivers for direct broadcasting of f.m. television from satellites at 12 GHz, a common requirement in all variations is for a local oscillator which supplies several milliwatts of power at approximately 11·7 GHz. A specification is derived for the maximum allowable noise on the local oscillator when the device used is an avalanche diode oscillator or a Gunn effect oscillator. Spectra are derived for these two cases for two margins of safety. These results are than applied to two measured spectra from different Gunn effect oscillators and it is shown how the measured spectra compare with the calculated specification. Practical tests using a coaxial varactor-tuned Gunn oscillator indicate that there is no degradation in picture quality when using such devices. In fact they suggest that the signal/noise criteria chosen for the local oscillator are rather more conservative than is necessary.

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1 Introduction

The possibility of broadcasting television signals from geostationary satellites directly to the home or to a small community receiving system has been much discussed in the last few years. For developed countries such broadcasts would have to make use of bandspace not previously used for this purpose. The 1971 World Administrative Radio Conference of the I.T.U. authorized satellite broadcasting, *inter alia*, in a band at 12 GHz. For some time the Mullard Research Laboratories have been studying various 12-GHz television receiving systems and experimental models of both double and single superheterodyne systems have been built and operated successfully and their performance measured[1].

In all systems studied there is a requirement for a local oscillator source at s.h.f. to mix with the incoming signal and form the intermediate frequency (or first intermediate frequency in the case of the double superhet). The exact requirements for this local oscillator will vary with the system chosen, e.g. it may be fixed tuned or variably tuned, but in all cases several milliwatts of power at approximately 11·7 GHz are required.

It is probable that the video information will be transmitted as frequency modulation of the s.h.f. carrier and so it is noise of a frequency-modulated nature that will determine the ultimate sensitivity of the system and the quality of picture obtained. The a.m. noise is assumed to be removed in the limiting amplifier preceding the discriminator. Calculations to find the minimum onboard satellite power, for a given picture quality etc., whilst allowing for the noise figure of the s.h.f. down converter, usually ignore any noise produced by the local oscillator itself and consequently assume that it is negligible. However any f.m. noise that is present on the signal from the local oscillator will be converted by the mixer to the intermediate frequency where it will be present both with the f.m. signal and other f.m. noise. Thus the problem is to find the maximum allowable f.m. noise on the local oscillator such that this noise is insignificant with respect to the f.m. noise already in the system (and allowed for), and to express this noise in a form suitable for comparison with measurements on an instrument such as a spectrum analyser.

2 The Local Oscillator Noise Spectrum

Several devices for producing the type of local oscillator signal required are currently available, but restricting the choice to those of the semiconductor variety there is basically the Gunn-effect oscillator or the avalanche diode transit-time oscillator. The avalanche oscillator, although slightly more efficient, operates at a higher voltage than the Gunn diode (namely at ∼ 60 V) and produces more power than required for a local oscillator. This is unattractive if one is considering using microstrip technologies where heat extraction is a problem. Moreover the lower operating voltage of the Gunn device makes it more compatible with other transistor circuits. Because of this and because Gunn devices can readily be mounted in microwave cavities, these have been the devices that have so far been utilized. In the future it is highly probable that transistors of the GaAs—f.e.t.

variety capable of oscillating at 12 GHz will be an economically viable proposition, but this paper is concerned mainly with the use of the Gunn effect oscillator (and to a lesser extent, mainly for comparison, the avalanche oscillator) as the local oscillator.

The noise in the Gunn oscillator is predominantly f.m., this noise being produced by the variation in the position of the formation of a domain on the Gunn chip and also variations in the time of formation. Experimental results,[2] indicate that the a.m. noise is very low. For example, in reference 2 it is shown that for a Gunn oscillator the a.m. noise in a 1 kHz band 100 kHz from the carrier is approximately 45 dB below the f.m. noise. In any event, the a.m. noise will be assumed to be removed by the limiter. Since the f.m. noise when demodulated manifests itself as a.m. noise superimposed on the video signal, it is evident that only the demodulated noise produced in the video bandwidth is of importance. For 625-line PAL, system I, this sets a limit on the demodulated noise modulation frequencies of interest to the range 10 kHz to 5 MHz.[3] Further, it is important to know the frequency distribution of the demodulated f.m. noise power in this bandwidth as produced by the local oscillator. This is because the eye of the viewer is less sensitive to high frequency noise in the final picture, so that low-frequency noise in the luminance channel (and in the demodulated chrominance channel) is more perceptible than a similar amount of noise at high video frequencies. For example, if we have two sources of microwave power producing the same total f.m. noise power in the demodulated bandwidth 10 kHz to 5 MHz but one is predominantly low frequency and the other predominantly high frequency, then it will be the former that is subjectively the noisier.

The sound channel will be assumed to be of similar quality to the video channel, i.e. the quality will be excellent. The degradation due to the local oscillator noise that will be allowed in the video channel will be very small and it will be assumed that the degradation in the sound channel is similar. As will be seen later, the main noise distribution considered is one where the r.m.s. noise voltage falls with the square root of the baseband frequency and so qualitatively it is seen that the subjective degradation will be less on the sound channel than on the video channel.

It is convenient to be able to describe the noise distribution by a mathematical relationship but unfortunately the measured f.m. noise spectra do not follow a simple law over the whole bandwidth. Indeed devices in low Q cavities ($Q \sim 50$) as used can give very different results.[4] However, it is possible to say from experimental results reported[4,5,6] that the f.m. noise produced by the Gunn effect device is mainly such that the deviation (Δf_{rms}) at a given frequency (f_m) removed from the carrier is proportional to $1/f_m^{\frac{1}{2}}$. According to Magarshack† the noise spectrum has this very strong ($1/f_m$) component, i.e. $\Delta f_{rms}^2 \propto 1/f_m$ near the carrier but at frequencies further away from the carrier it falls off more

rapidly according to the relationship $\Delta f_{rms}^2 \propto A[B + (f_m \tau)^2]$ where A and B are constants and τ is a relaxation time. For the purposes of this paper the simpler distribution will be assumed to be valid over the whole bandwidth. This will not lead to great inaccuracies because of the weighting effect as described before and in any case will produce a slightly tighter specification on the allowable f.m. noise than is perhaps necessary.

The avalanche oscillator, as mentioned before, will also be included for comparison and also because it is possible that such a device might be used. Again the a.m. noise for this device is of the order of 45 dB below the f.m. noise.[2] The f.m. noise however in this case shows a distribution which is essentially flat,[7] i.e. $\Delta f_{rms} = K$ (where K is a constant).

Therefore the local oscillator will be assumed to have an f.m. noise spectrum given by

(a) $\Delta f_{rms} \propto K$

(b) $\Delta f_{rms} \propto 1/f_m^{\frac{1}{2}}$.

3 Signal/Noise Ratios and Weighting Factors

It has been previously noted that the noise contribution from the local oscillator must be small with respect to the noise otherwise present and allowed for in the calculation of picture quality from satellite power etc. To obtain the tightest specification on the local oscillator noise the procedure is as follows. The signal/weighted-noise ratios required in the final video chain of the television receiver for a very good picture are taken to be[8]

48 dB in the luminance channel

and 42 dB in the chrominance channel.

These figures correspond to Grade 1 (excellent) on the EBU subjective quality scale where the signal/weighted-noise ratio is defined by

$$20 \log_{10} \frac{\text{peak to peak luminance (or chrominance) signal}}{\text{weighted r.m.s. noise voltage}}$$

This is the final signal/noise ratio required and should include all the noise contributions whatever their source. If the noise from the local oscillator is to be small then an appropriate signal/weighted-noise ratio due to this source is required. A signal/weighted-noise ratio due to the incoming noise has been chosen to be 48 dB. For the local oscillator two rather arbitrary values of noise, the noise margins, due to the oscillator alone are chosen, i.e. 10 dB below and 20 dB below the incoming noise level. This implies signal/weighted noise ratios when considering the effect of noise due to the local oscillator alone of 58 dB and 68 dB respectively. The total resultant signal/weighted-noise ratio, i.e. incoming and local oscillator noise, will then be 45·6 dB and 47·2 dB respectively for the two cases, i.e. a degradation of 2·4 dB and 0·8 dB respectively. The former may not appear at first sight to be a small change but in fact experimental results due to Weaver[9] indicate that with such a good initial picture quality a 2·4 dB increase in noise will cause a degradation of less than half a grade on the standard six-point subjective scale of impairment. The effect of a 0·8 dB change will therefore be imperceptible on the screen.

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† Magarshack, J., Laboratoires d'Electronique et de Physique Appliquée, Limeil-Brévannes. Private communication; see also ref. 4.

The signal/unweighted-noise ratios are found by applying the CCIR standard weighting network[3] to the noise signal present and thus deriving correction factors for the weighted noise voltages. The weighting network for the luminance channel 10 kHz → 5 MHz has a time-constant of 200 ns and the chrominance channel 3·5 MHz → 5·5 MHz has a similar network which is applied to the demodulated 1 MHz bandwidth chrominance channel. The correction factor inevitably depends upon the frequency distribution of the noise being considered. In the previous section it was shown that this is of two forms: (a) a flat distribution with frequency and (b) a distribution where the noise falls as $1/f^{\frac{1}{2}}$. The flat frequency distribution is identical with the normal a.m. case for which the weighting factors are:

<div align="center">

luminance channel   6·5 dB

chrominance channel 4·2 dB.

</div>

The appropriate weighting factor for case (b) must be calculated but before doing so it is worthwhile looking at how the weighting factors affect the relative signal/noise ratios of the luminance and chrominance channels. The standard result for the flat noise case has been given; a second standard result is for the case where the noise increases with frequency after demodulation ('triangular' noise) this being appropriate to the demodulated Gaussian noise when using an f.m. system in, for example, a television link.[3] Thus the signal/unweighted-noise ratios can be found as in Table 1.

The usefulness of comparing these results is seen by considering the channel which is most sensitive to noise. In case (b) the chrominance channel is approximately 6 dB more sensitive than the luminance channel. In case (a) however the luminance channel is more sensitive, i.e. 3·7 dB more than for the chrominance. From extrapolation of these results in a qualitative manner it can be

Table 1

Calculation of signal/unweighted-noise ratios

		signal/ weighted- noise	weighting factor	signal/ unweighted- noise
(a) flat noise	luminance	x	6·5	$x - 6·5$
	chrominance	$x - 6$	4·2	$x - 10·2$
(b) increasing	luminance	x	12·3	$x - 12·3$
triangular noise	chrominance	$x - 6$	0·3	$x - 6·3$

Table 2

Video-signal/unweighted-noise ratios for a Grade 1 picture and the two noise margins

	Signal/unweighted-noise ratios		
	Grade 1 picture (0 dB margin)	10 dB margin	20 dB margin
(a) flat noise	41·5	51·5	61·5
(b) decreasing noise ($\propto 1/f^{\frac{1}{2}}$)	46·5	56·5	66·5

concluded that where noise is *falling* with frequency it will again be the luminance channel which is the sensitive channel and consequently it is only necessary to calculate a weighting factor for this channel. This calculation is shown in Appendix 1 and the result is a weighting factor equal to 1·5 dB.

For the starting point of a very good (Grade 1) picture (video/weighted-noise = 48 dB) and the two noise margins chosen, the signal/unweighted-noise ratios for the two noise distributions of interest are therefore as shown in Table 2.

These are the figures relating to local oscillator noise alone that will be used in the remainder of this paper but before leaving this Section the possible use of pre-emphasis will be qualitatively considered. With an incoming noisy f.m. signal, the demodulated output will have a noise spectrum increasing with frequency. It has already been pointed out that this effectively makes the chrominance channel the most sensitive. It is certain therefore that pre-emphasis and subsequent de-emphasis in the region of the colour sub-carrier frequency will improve the signal/noise ratio by effectively reducing the noise. This will also modify the noise distribution of the demodulated local oscillator signal whether this is of the flat or decreasing triangular distribution. In the latter case the modification will be slight since the noise is already low at that point, the dominant noise being at low frequencies. For the flat noise case the effective reduction will be more marked, though not nearly as much as the effect of de-emphasis on the increasing triangular spectrum of demodulated Gaussian noise. It is concluded therefore from qualitative arguments that pre-emphasis and de-emphasis of the colour sub-carrier frequencies of the incoming signal will have very little effect on the contribution to the final picture of the noise of a local oscillator with a decreasing noise spectrum and only a small effect on the contribution of the noise of a local oscillator with a flat noise spectrum. Since the inclusion of an, as yet unspecified, amount of pre-emphasis and de-emphasis would greatly complicate the analysis but seems unlikely to make much difference to the results, its possible use will not be considered further here.

4 Determination of Δf_{rms} for the Two Noise Distributions and Two Safety Margins

The frequency modulated signal will have a total peak-to-peak deviation that will be the sum of the deviations produced by the sound sub-carrier, the sync pulses, the black-to-white transition, and the peak chroma. From the definition of signal/noise ratio, i.e.

$$\left(\frac{S}{N}\right)_{dB} = 20 \log_{10}\left(\frac{\text{luminance signal}}{\text{r.m.s. noise signal}}\right),$$

the deviation produced by the black-to-white transition is of most interest. Many calculations have been done using different assumptions but the consensus of opinion would appear to favour a total bandwidth for the f.m. channel of the order of 27 MHz with the deviation (p-p) corresponding to the luminance signal alone being of the order of 7·5 MHz.[10] Using this figure, Δf_{rms} for the two cases can be determined.

4.1 The Flat Noise Case

For this case with the two margins the signal to noise ratio can be written

$$\left(\frac{S}{N}\right)_{dB} = 20 \log_{10}\left(\frac{7 \cdot 5}{\Delta f_{rms}}\right) = 51 \cdot 5 \text{ dB or } 61 \cdot 5 \text{ dB}$$

where Δf_{rms} is expressed in MHz and is composed of all the noise power contribution in the 5 MHz video bandwidth.

Thus $\Delta f_{rms} = 20$ kHz or $6 \cdot 3$ kHz in a 5 MHz band. This is equivalent to $\Delta f_{rms} = 9 \cdot 0$ Hz or $2 \cdot 8$ Hz in a 1 Hz band (assuming $\Delta f_m \propto 1/f_m^{\frac{1}{2}}$).

These are the two values for Δf_{rms} for the two margins chosen in the flat noise case.

4.2 The Decreasing Noise Case

In this case the calculation is not quite as simple since the noise distribution varies with frequency. It is therefore necessary to integrate over the total baseband frequency range to determine the power. Defining the r.m.s. frequency deviation produced by the noise in a 1 Hz band at 10 kHz as $\Delta f_{0.01}$, then

$$\Delta f_{rms}^2 = \frac{10^4 \Delta f_{0.01}^2}{f} \text{ where } f \text{ is in Hz,}$$

and the demodulated power produced by the noise is given by

$$P_{noise} = K' \int_{0.01 \times 10^6}^{5 \times 10^6} \frac{10^4 \Delta f_{0.01}^2}{f} \, df$$

where K' is a constant.

The power due to the carrier over the black-to-white transition is given by

$$P_{carr} = K'(7 \cdot 5 \times 10^6)^2$$

evaluating these two expressions gives

$$P_{noise} = 10^4 \Delta f_{0.01}^2 \, K' \, 6 \cdot 22$$
$$P_{carr} = K' \, 56 \times 10^{12}.$$

Hence the signal to noise ratio is given by

$$10 \log_{10} \frac{P_{carr}}{P_{noise}} = 10 \log_{10} \frac{65 \times 10^{12} K'}{6 \cdot 22 \times 10^4 K' \Delta f_{0.01}^2}$$
$$= 56 \cdot 5 \text{ dB or } 66 \cdot 5 \text{ dB}$$

Thus

$$\Delta f_{0.01} = 45 \text{ Hz} \quad \text{or} \quad 14 \cdot 2 \text{ Hz in a 1 Hz bandwidth.}$$

These are the two values of the root mean square deviation for the noise evaluated at 10 kHz for the two noise margins as measured in a 1 Hz band. The noise voltage falls off above 10 kHz according to the relation: $\Delta f_{rms} = 10^2 \Delta f_{0.01}/f^{\frac{1}{2}}$.

5 Determination of Target Spectra

Having determined the values of Δf_{rms} for the four cases it is now necessary to relate these to a readily observable quantity. In this instance it is assumed that the spectrum of the device used as the local oscillator will be measured on a spectrum analyser with r.m.s. detection and hence the ratio of carrier to sideband levels
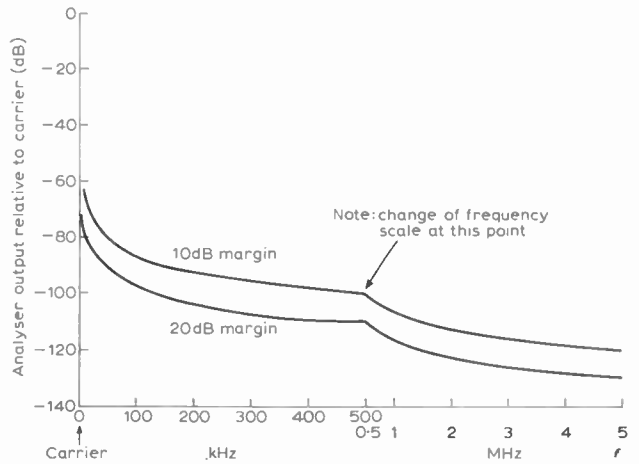

Fig. 1. Spectra for the 10 dB and 20dB noise margins in the flat noise case, i.e. $\Delta f_{rms} = 9 \cdot 0$ Hz or $2 \cdot 8$ Hz in a 1 Hz bandwidth.

will be a function of the r.m.s. deviation as calculated. It has already been pointed out that the a.m. noise is assumed to be negligible compared with the f.m. noise and consequently the spectrum observed on the analyser will be mainly due to f.m. noise components. Following the development given by Shields[11] the procedure is as follows. (See Appendix 2 for more details.)

A simple frequency modulated carrier (carrier frequency $= \omega_c$ rad/s and modulating frequency $= \omega_m$ rad/s) can be expressed as a series of Bessel functions:

$$J_0(m) \sin \omega_c t + J_1(m) \sin (\omega_c + \omega_m)t +$$
$$+ J_2(m) \sin (\omega_c + 2\omega_m)t + J_1(m) \sin (\omega_c - \omega_m)t +$$
$$+ J_2(m) \sin (\omega_c - 2\omega_m)t + \ldots$$

where m, the modulation index $= \Delta f/f_m$, with Δf the deviation and f_m the baseband frequency (Hz).

Now for

$$m \leqslant 0 \cdot 3 \qquad J_0(m) \simeq 1$$
$$J_1(m) \simeq m/2$$

and $J_2(m)$, $J_3(m)$ etc. are negligible.

Hence each modulating frequency produces two sidebands only, and by adding the noise sinusoids for each baseband frequency a complete spectrum can be obtained. At each frequency removed from the carrier:

$$\frac{V_{sideband}}{V_{carrier}} = \frac{J_1(m)}{J_0(m)} = \frac{m}{2} = \frac{\Delta f}{2f_m}.$$

Hence

$$\left(\frac{V_{sideband}}{V_{carrier}}\right)_{dB} = 20 \log_{10}\left(\frac{\Delta f}{2f_m}\right) \quad (m \leqslant 0 \cdot 3).$$

For $m > 0 \cdot 3$ the more conservative limit

$$\left(\frac{V_{sideband}}{V_{carrier}}\right)_{dB} = 20 \log_{10}\left(\frac{J_1}{J_0}\right)$$

is assumed.

For the two noise distributions of interest and their two margins these formulae are used to calculate the observed carrier/sideband ratios of the spectrum. The details are given in Appendix 2 and the results are shown in Figs. 1 and 2.
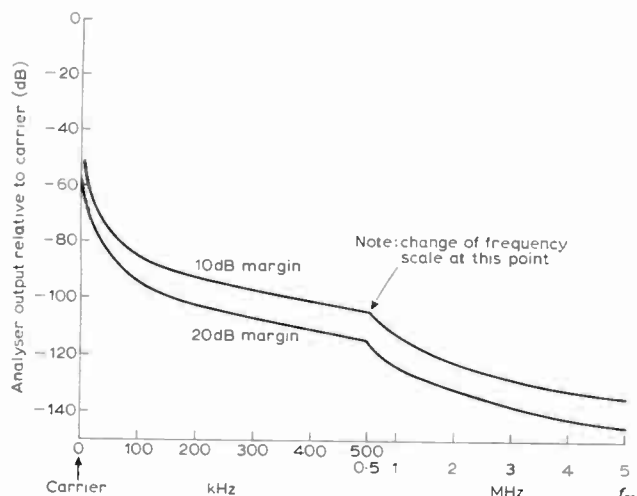
Fig. 2. Spectra for the 10 dB and 20 dB noise margins in the decreasing noise case, i.e. $\Delta f_{rms} = 45$ Hz or $14\cdot2$ Hz in a 1 Hz band 10 kHz from the carrier.

6 Experimental Results and Discussion

To assess the practical value of the target spectra derived according to the above theoretical analysis experimental investigations were made of the noise spectra of available Gunn oscillators and their subjective effect in an actual system.

The experimental observations were therefore in two parts. First the noise spectra produced by two different coaxial Gunn oscillators were measured on a spectrum analyser and the displays photographed. The two oscillators used were a commercially available Mullard mechanically tuned device, the CL8390, and an MRL experimental varactor-tuned oscillator (in which electronic tuning was achieved by application of a variable

Fig. 3. Observed spectrum for coaxial Gunn oscillator type CL8390. Centre frequency = 11·45 GHz, spectrum analyser bandwidth = 3 kHz.

voltage to a varactor diode mounted inside the cavity). The measured spectra are shown in Figs. 3 and 4. Although both devices had Q factors of a few tens the CL8390 had the higher Q and, as expected, it exhibited a less noisy spectrum. Superimposed on the measured spectra are shown calculated curves derived from the $1/f^4$ noise distribution law assumed for Gunn oscillators in the theoretical analysis above, and which most nearly fit the observed distribution.

Thus assuming the analysis to be valid, we see that when considering the contribution to the video signal/ noise ratio of the local oscillator alone, the CL8390 noise spectra corresponds to a noise margin of 10 dB below that for a Grade 1 picture—see Table 2. We would therefore expect the noise due solely to a CL8390 local oscillator to be well below the level of perceptibility.

For the experimental varactor-tuned oscillator the noise spectra appears to correspond to a noise margin of 0 dB below that for a Grade 1 picture. Thus we would expect the noise due solely to a varactor-tuned oscillator
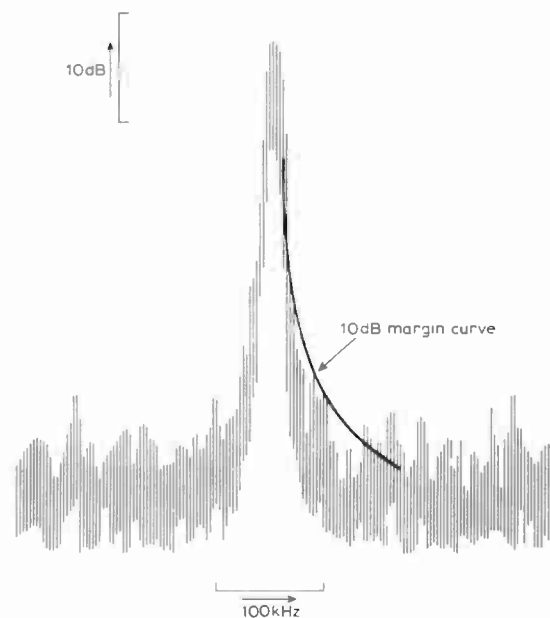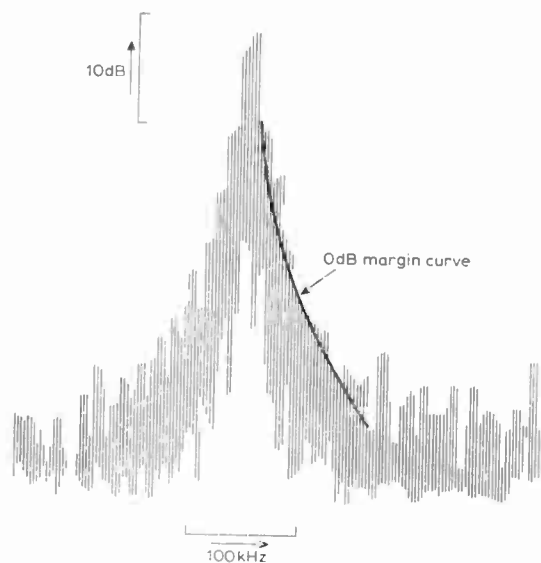

Fig. 4. Spectrum for experimental varactor tuned coaxial Gunn oscillator. Centre frequency = 11·7 GHz, spectrum analyser bandwidth = 3 kHz.

of the type used to be at or just below the limit of perceptibility.

To verify this last conclusion the second part of the observations consisted of a subjective examination of the noise produced in a working simulation of a satellite broadcasting system in which two varactor-tuned Gunn oscillators were used. The first was used to generate a source of 12 GHz f.m. signals, the input to the frequency modulating varactor being an electronically generated PAL colour bar test signal (which could therefore itself be assumed to be noise-free). The 12 GHz f.m. output was coupled directly via an attenuator to the input of an experimental single-superhet (image rejection) 12 GHz receiver[1] which used the second varactor-tuned Gunn oscillator as the local oscillator. The input was at a level sufficient to swamp any noise contribution due to the noise factor of the converter. The demodulated output from the frequency discriminator was then fed to a colour

monitor via a PAL decoder. Thus the only noise expected in the final picture was that due to the two varactor-tuned oscillators. Under these conditions no degradation of the picture by noise was observed. From this it seems reasonable to conclude that the contribution to picture noise arising solely from the use of a single varactor-tuned Gunn device—as the receiver local oscillator—is below the limit of perceptibility. This is perhaps rather better than predicted by the theoretical analysis, but as was pointed out, some of the assumptions made were known to err on the conservative side.

From these results, it would appear that the calculated target spectra for the local oscillator f.m. noise, at least in the case of Gunn devices, is a realistic, if perhaps rather conservative, one. The observations to date suggest that some degradation of performance relative to the calculated targets may well be acceptable.

7 Conclusions

The purpose of this paper has been to provide a tentative target specification for the noise spectrum of a local oscillator used in a 12 GHz f.m. satellite television broadcast receiver. For the purposes of calculation two rather arbitrary noise margins have been chosen such that the weighted noise produced by the local oscillator alone is respectively 10 dB or 20 dB below a weighted noise level due to all other sources which in the absence of oscillator noise would yield a Grade 1 picture on the EBU six-point subjective impairment scale. By making some simplifying approximations regarding the shape of the f.m. noise spectra of Gunn and avalanche devices two sets of curves have been derived, as in Figs. 1 and 2, which indicate the expected noise spectra corresponding to the above two noise margins. For Gunn devices these correspond to unweighted video signal/noise ratios due to the local oscillator alone of 56·5 dB and 66·5 dB respectively.

Which of these two criteria (or some less stringent criterion) should be adopted is not certain. Comparison of the observed noise spectrum of coaxial Gunn oscillators with the theoretical analysis indicates, as shown in Figs. 3 and 4 for example, that the f.m. noise of a varactor-tuned Gunn oscillator alone would result in an unweighted signal/noise ratio of 46·5 dB—corresponding to a Grade 1 picture with a weighted signal/noise ratio of 48 dB—i.e. a noise margin of 0 dB. If this were the case, then in a system designed to produce a Grade 1 picture assuming no local oscillator noise, the presence of such a level of noise due to the oscillator would degrade the total weighted signal/noise by 6 dB—which would mean a drop in subjective Grade of 1 to 1½. However we may note that if one starts with a system design noise equivalent to, say, Grade 2·5 (i.e. 'good' on the EBU quality scale which corresponds to a 42 dB weighted signal-to-noise ratio and equivalent to 36 dB unweighted signal-to-noise assuming flat noise) then the subjective degradation additionally due to the local oscillator becomes relatively smaller.

Actual observations in an experimental system have proved only that the signal/noise criteria chosen for the local oscillator appear to be rather conservative. More work is required to determine whether worse noise margins can be tolerated.

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9 Appendix 1

Calculation of Weighting Factor for Decreasing Noise Spectrum

In the references which quote values of Δf_{rms}, the r.m.s. frequency deviation, as a function of frequency f_m about the carrier, the results are usually taken from measurements made on an instrument of finite bandwidth. For example, Herbst and Ataman show[12] that Δf_{rms}, (or ΔV_{rms}, the demodulated voltage produced by Δf_{rms}), is proportional to $1/f_m^{\frac{1}{2}}$ when measured using a 100 Hz bandwidth. In the present analysis it is necessary to assume that as the measurement bandwidth is shrunk to 1 Hz the dependence of Δf_{rms} on $1/f_m$ is retained at all values of f_m so that a smooth curve is obtained.

The power spectrum will be given by

$$\Delta f_{rms}^2 \propto \Delta V_{rms}^2 \propto \frac{1}{f_m}.$$

In the bandwidth of interest, i.e. the luminance channel 10 kHz to 5 MHz the total power will be given by

$$P_{tot} = \int_{0·01 \times 10^6}^{5 \times 10^6} K \frac{df}{f_m} \text{ where } K \text{ is a constant}$$
$$= K(\log_e f)_{0·01 \times 10^6}^{5 \times 10^6}$$
$$= 6·215 K.$$

The signal is now passed through a weighting network, of time constant $\tau = 200$ ns, so that at each frequency, f, the power is reduced by a factor $1/[1+(2\pi f\tau)^2]$. Therefore after the filter, the total power P'_{tot} will be given by

$$P'_{tot} = \int_{0.01 \times 10^6}^{5 \times 10^6} \frac{K \, df}{f(1+(2\pi f\tau)^2)}.$$

Now

$$\frac{K}{f(1+(2\pi\tau f)^2)} = \frac{K}{f} - \frac{K(2\pi\tau)^2 f}{1+(2\pi\tau f)^2}.$$

Therefore

$$\int_{0.01 \times 10^6}^{5 \times 10^6} \frac{K \, df}{f(1+(2\pi f\tau)^2)} = \int_{0.01 \times 10^6}^{5 \times 10^6} \frac{K \, df}{f} - \int_{0.01 \times 10^6}^{5 \times 10^6} \frac{K(2\pi\tau)^2 f \, df}{1+(2\pi\tau f)^2}.$$

The first integral on the right-hand side has already been calculated. The second integral is equal to

$$\frac{K}{2} \left[\log_e (1+(2\pi\tau f)^2) \right]_{0.01 \times 10^6}^{5 \times 10^6}$$

for $\tau = 200 \times 10^{-9}$ and $f = 5 \times 10^{+6}$,

$$(2\pi f\tau)^2 = 39.5$$

for $\tau = 200 \times 10^{-9}$ and $f = 0.01 \times 10^6$,

$$(2\pi f\tau)^2 = 1.58 \times 10^{-4}.$$

Therefore

$$\text{integral} = \frac{K}{2} [\log_e 40.5] - \frac{K}{2} [\log_e 1]$$

$$= 1.857 \, K = P, \text{ say}.$$

Hence reduction in power due to network

$$= 10 \log_{10} \left[\frac{P_{tot}}{P'_{tot}} \right] = 10 \log_{10} \left[\frac{P_{tot}}{P_{tot} - P} \right]$$

$$= 10 \log_{10} \frac{6.215}{6.215 - 1.857}$$

$$= 1.52 \text{ dB}.$$

Therefore the weighting factor for noise with distribution $\Delta f_{rms}^2 \propto 1/f_m$ is 1.52 dB.

10   Appendix 2
Derivation of Spectra

For a simple frequency modulated carrier, the carrier and sidebands are given by

$$J_0(m) \sin \omega_c t +$$

$$J_1(m) \sin (\omega_c + \omega_m)t + J_1(m) \sin (\omega_c - \omega_m)t +$$

$$J_2(m) \sin (\omega_c + 2\omega_m)t + J_2(m) \sin (\omega_c - 2\omega_m)t + \ldots \text{etc.}$$

where J_0, J_1, J_2 etc. are the zero, first, second order etc. Bessel functions

ω_c	is the carrier frequency (in rad/s)
ω_m	is the modulation frequency (in rad/s)
m	is the modulation index $= \Delta f/f_m$
Δf	is the deviation (in Hz)
f_m	is the baseband frequency (Hz) $= \omega_m/2\pi$.

The values of J_0, J_1 etc. for various values of m are given as follows:

m	J_0	J_1	$m/2$	J_2	J_3
1	0.765	0.44	0.5	0.115	0.020
0.7	0.881	0.399	0.35	0.059	0.069
0.6	0.912	0.329	0.3	0.044	0.044
0.5	0.938	0.242	0.25	0.031	0.003
0.4	0.966	0.196	0.2	0.020	0.001
0.3	0.978	0.148	0.15	0.011	0.001
0.2	0.990	0.100	0.1	0.005	0.000
0.1	0.998	0.050	0.05	0.001	0.000

It is seen that for $m \leqslant 0.3$, $J_0 = 1$, $J_1 = m/2$ and J_2, J_3 etc. $= 0$ to within 2%. Therefore the relative amplitude of first sideband to carrier is given by

$$\frac{V_{sb}}{V_c} = \frac{J_1(m)}{J_0(m)} = \frac{m}{2} = \frac{\Delta f}{2f_m} \qquad m \leqslant 0.3$$

or

$$\left(\frac{V_{sb}}{V_c} \right)_{dB} = 20 \log \frac{\Delta f}{2f_m}.$$

Therefore, assuming certain values of $\Delta f, f_m$ etc. within the condition $m \leqslant 0.3$ enable the spectrum as shown on a spectrum analyser to be calculated. For $m > 0.3$, $\Delta f/2f_m$ must be replaced by $J_1(m)/J_0(m)$ and the error introduced by neglecting J_2, J_3 must be observed separately. Any contribution from higher-order Bessel functions of lower frequencies will tend to increase the spectrum at the frequency of interest and hence if these contributions are ignored then the resultant bound determined for the given noise distribution and margin will be tighter than that including all the contributions. Hence the specification calculated will be tighter than necessary, i.e. it will err in the correct direction. The spectrum analyser in general is an r.m.s. measuring device and hence Δf in this case will be the r.m.s. deviation as required by the analysis in the text.

10.1   The Spectrum for the Flat Noise Case

The calculated values of Δf_{rms} for the two safety margins were 9·0 Hz and 2·8 Hz for the 10 dB and 20 dB margins respectively as measured in a 1 Hz bandwidth.

10.1.1   *10 dB margin*

f_m (MHz)	$\Delta f/2f_m$	$20 \log_{10} \Delta f/2f_m$ (dB)
0.01	4.5×10^{-4}	-67
0.02	2.25×10^{-4}	-73
0.05	9×10^{-5}	-80
0.07	6.4×10^{-5}	-84
0.1	4.5×10^{-5}	-87
0.2	2.25×10^{-5}	-93
0.5	9×10^{-6}	-100
0.7	6.4×10^{-6}	-104
1	4.5×10^{-6}	-107
2	2.25×10^{-6}	-113
5	10^{-6}	-120

10.1.2 *20 dB margin*

These values will be 10 dB below the previous case. These curves are plotted in Fig. 1. The curves for margins other than 10 and 20 dB relative to the signal/noise ratio assumed in the absence of any local oscillator noise, (or for any other signal/noise ratio chosen) can be derived from these by the addition or subtraction of the appropriate number of decibels.

10.2 The Spectrum for the Decreasing Noise Case

The calculated values of Δf_{rms} evaluated at 10 kHz for the two safety margins were 45 Hz and 14·2 Hz for the 10 dB and 20 dB margins respectively as measured in a 1 Hz band.

10.2.1 *10 dB margin*

f_m (MHz)	Δf_{rms} (Hz)	$\Delta f/2f_m$	$20 \log_{10} \Delta f/2f_m$
0·01	45	$2·25 \times 10^{-3}$	− 53
0·02	31·8	$7·95 \times 10^{-4}$	− 62
0·05	20·1	$2·01 \times 10^{-4}$	− 74
0·07	17·1	$1·22 \times 10^{-4}$	− 78
0·1	14·23	$7·11 \times 10^{-5}$	− 85
0·2	10·1	$2·53 \times 10^{-5}$	− 92
0·5	6·4	$6·4 \ \times 10^{-6}$	− 104
0·7	5·4	$3·8 \ \times 10^{-6}$	− 108
1	4·5	$2·25 \times 10^{-6}$	− 113
2	3·2	8×10^{-7}	− 122
5	2·0	2×10^{-7}	− 134

10.2.2 *20 dB margin*

The values will be 10 dB below the previous case. These curves are plotted in Fig. 2. As before, curves for other conditions can be derived from these directly.

The Author

Dr. Trevor Linnecar read electrical sciences at Trinity College, Cambridge, obtaining his first degree in 1967, and he continued to work at the Cavendish Laboratory for his doctorate which was awarded in 1970 for research into microwave amplification using the interaction between an electron beam and a plasma. Between 1971 and 1973 Dr. Linnecar was a member of a small team at Mullard Research Laboratories studying the problems of the reception in the home of colour television signals transmitted on a 12 GHz carrier from a geostationary satellite. He is now at CERN II, Geneva, working with the R.F. Group of the 300 GeV proton accelerator project.

Applications of the Taguchi gas sensor to alarms for inflammable gases

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and

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SUMMARY

In domestic, commercial, light industrial, and some recreational environments, there exists a requirement for the economical detection of potentially explosive concentrations of inflammable gases. The recently introduced Taguchi gas sensor, or T.G.S., is an inexpensive semiconductor transducer which is sensitive to all inflammable gases, and has electrical characteristics which appear to make it suitable for incorporation into the relevant instruments. These characteristics are presented, as are the basic circuit design concepts for operating the device from both mains and battery power sources. Although brief notes on safety requirements are included, it is not intended that the paper should advocate, or otherwise, the use of the device from this point of view.

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List of Principal Symbols

V_H r.m.s. voltage applied to the T.G.S. heater filament

V_B direct voltage applied across the T.G.S. bead, between the filaments

V_{out} output voltage of operational amplifier

R_B bead resistance between the filaments

R_F feedback resistor across operational amplifier

P_B power dissipated in the T.G.S. bead

I_B bead current between the filaments

k T.G.S. sensitivity in mA per parts/10^6

C gas concentration (parts/10^6)

Z projection of I_B/C curve at the Y-axis when $V_H = V_{H(opt)}$

1 Introduction

The lowest concentration of an inflammable gas in air which will ignite is known as the l.e.l., or lower explosive limit. The converse of this is the higher explosive limit, or h.e.l., which defines the lowest concentration of air in gas which will also ignite. For normal environments, it is only the l.e.l. which is of importance, and the present paper is concerned only with applications relevant to areas called 'Division 2' by the relevant British Standard Code of Practice,[1] namely 'an area within which any flammable or explosive substance, whether gas, vapour or volatile liquid, although processed or stored, is so well under conditions of control that the production (or release) of an explosive or ignitable concentration in sufficient quantity to constitute a hazard is only likely under abnormal conditions'.

This definition clearly encompasses homes and commerical premises in which town gas, natural gas or bottled gas is in use, and can reasonably be extended to include recreational and other vehicles, including small boats and caravans wherein fuel is stored, possibly in addition to bottled gas. Although the hazard from gas or fuel vapour leakage in such premises is individually small, their sheer numbers ensure that accidents do occur frequently on a country-wide basis. Unfortunately, this very small individual hazard implies that the provision of a means of leak detection is of low priority to the relevant owners, so that any instrument developed for this purpose must be a low-cost device. It is for this reason that the Taguchi gas sensor[2] is currently the most appropriate transducer, and it will be shown that it can be incorporated into instruments which are of a cost low enough to permit widespread installation provided that both the sensor and the instrument are in due course shown to conform with safety standards. At the time of writing, such safety standards have yet to be formulated and published, but certain of the more obvious requirements can be tentatively listed as follows:

(i) In Division 2 situations, it would be expected that a gas *alarm* as opposed to a concentration-measuring instrument would be required. Since the purpose of an alarm is to draw attention to a potential hazard, an audible, rather than a visual, display would be necessary.
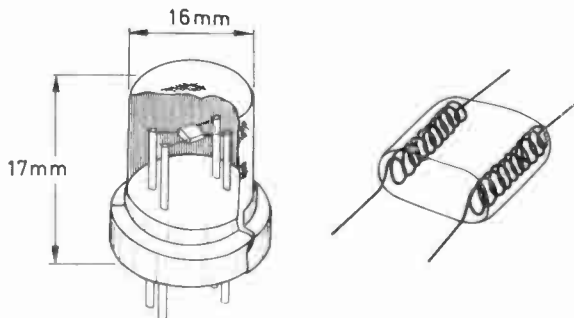
Fig. 1. Structure of complete Taguchi sensor and enlarged sectional view of detector bead and filaments.

(ii) The alarm could reasonably be expected to be set to operate at about 20% l.e.l. Such a figure is arrived at by considering the possibility of false alarms due to 'safe' sources of inflammable vapour, such as paint, solvent, or alcohol; and by bearing in mind that the siting of the alarm is all-important, it being difficult to predict where the major concentration of gas will occur.

(iii) The alarm should operate reliably (i.e. within about $\pm 20\%$ of the set level) under various specified conditions of electrical supply and environmental circumstance, including:

  (a) For mains-operated alarms, a mains variation of about $\pm 10\%$. (Note—the legally allowable mains variation is $\pm 6\%$.)

  (b) For battery-operated alarms, a variation in battery voltage of -5% to $+15\%$.

  (c) Within circumscribed limits of temperature, humidity, atmospheric pressure, and wind speed. (Limits are not suggested at this time since the considerable information necessary on light industrial environments is not immediately available.)

(iv) Electrical safety in Division 2 areas is covered by the relevant BASEEFA documents.[3] In the case of gas alarms, the important implications are that no spark or arc should be produced within the instrument, and no part should become sufficiently hot to ignite any inflammable gas. Hence, the audible alarm should not be of a type which includes vibrating contacts, and any relays should be reed types or be inherently flameproof.

2 Power Sources

For the situations in which it is envisaged that gas alarms would be useful, available power supplies may be tabulated as follows:

(i) Mains: available in houses, flats, shops, offices etc.

(ii) 12 V or 24 V d.c.: usually available in boats, land vehicles and aircraft.

(iii) Internal batteries: necessary for fully portable operation.

In order to show that the T.G.S. is capable of operation from all three sources, an introduction to its mode of operation and characteristics follows.

3 Construction and Operation of the T.G.S.

Figure 1 is a sketch of the physical configuration of the T.G.S. It consists essentially of two helical platinum-iridium filaments around which is formed a glassy bead of a bulk N-type semiconductor based on stannic oxide. The ends of the filaments are welded to nickel support pins which protrude through a header as shown, and the entire structure is protected by a two-layer stainless steel gauze envelope.

In operation, one of the filaments is used to heat the bead to a working temperature which may be up to 300°C. This greatly reduces the availability of electrons for conduction in the bead by anionic adsorption of oxygen. In the presence of a deoxidizing gas, cationic adsorption occurs which increases the apparent conductivity of the bead. After removal of the gas, adsorption of oxygen again becomes the dominant process. (Note that this implies that the sensor will not operate in the absence of oxygen, such as in vacuum conditions.) This change in resistance is the operating parameter of the device, and is measured by a second circuit, between the filaments.

It should be noted that during the cold state, some water vapour and gases are absorbed, so that upon initial heating, the resistance of the bead falls until these inclusions are expelled. This procedure can last up to a few minutes, after which the normal working resistance of the bead stabilizes at not less than 10 000 ohms.

In operation, the stainless steel gauze envelope acts as a Davy cage, so that even if an external atmosphere contaminated with an inflammable gas beyond l.e.l. is present, it should not be ignited by the T.G.S. (Mounting the T.G.S. behind a sintered bronze or stainless steel disk would further reduce the small risk of ignition.)

From the foregoing explanation, it will be clear that one filament must be continuously energized to maintain the bead at a constant temperature. (Normally, about 1 V d.c. or r.m.s. is applied, whereupon a current of about 0·6 A flows.) This implies that the resistance of the bead itself, as measured by a second circuit, will be a function of the filament temperature and the applied voltage, V_H. It also implies that the device will be somewhat sensitive to draughts in the sense that the rate of heat removal from the bead will affect its working temperature.
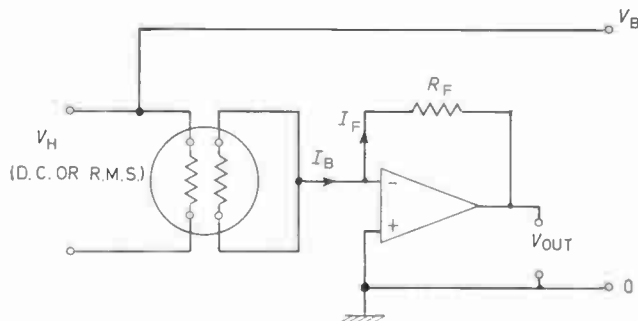

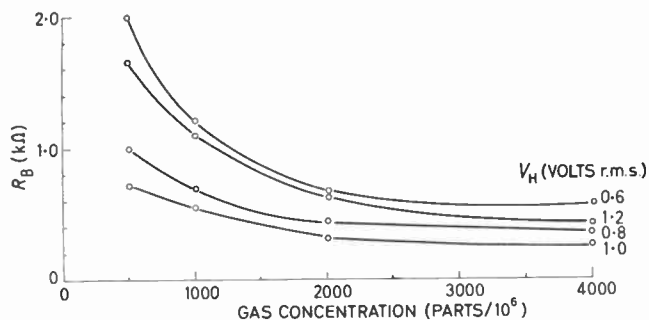Fig. 2. Measurement of I_B at constant V_B.

Fig. 3. Bead resistance versus petrol vapour concentration for an H10 T.G.S.

4 Characteristics of the T.G.S.

Measurement of the bead resistance as a function of both gas concentration and filament voltage, V_H, makes it possible to define operating conditions for most applications. Knowing that the temperature of the bead is of major importance, it is clearly necessary to establish that self-heating in the bead by the measuring circuit is avoided, and this is most easily done by applying only a low voltage. The circuit of Fig. 2 is particularly convenient for this purpose, for the voltage across the bead is held at V_B, which exists between the heater filament and the virtual earth point of the operational amplifier. Hence, the bead resistance is given by

$$R_B = \frac{V_B}{I_B} = \frac{V_B}{I_F} = \frac{V_B R_F}{V_{out}}. \tag{1}$$

Also, the power dissipated in the bead will be

$$P_B = V_B I_B = V_B I_F = \frac{V_B V_{out}}{R_F}. \tag{2}$$

(Both equations assume that the input current to the amplifier is small compared with I_B, which is the case if an f.e.t.-input module is used.)

It may be assumed that a power dissipation of 1 mW will not affect the bead temperature significantly compared with some 600 mW dissipated by the heater filament. Hence, the minimum bead resistance to give a maximum P_B of 1 mW if $V_B = 1$ volt will be:

$$R_{B(1\,mW)} = \frac{V_B^2}{P_B} = 1000 \text{ ohms.}$$

As an illustration of the performance to be expected from a T.G.S., Fig. 3 shows a family of curves of R_B versus petrol vapour concentration in air for $V_B = 1$ volt, using the circuit of Fig. 2. It is clear from these curves that the resistance change is indeed a marked function of the filament voltage, V_H; but that an alarm instrument which operates at 20% l.e.l. (as previously specified) is entirely feasible.

From the shape of the curves, it would seem more reasonable to plot the bead current, I_B, against gas concentration, rather than R_B, as derived from equation (1), and this has been done in Fig. 4. This family of curves shows that linearity of bead current with gas concentration is a function of heater voltage, and that for a particular T.G.S. a value of V_H can be chosen so that I_B is most nearly directly proportional to gas

concentration. This finding is of importance if a measuring instrument is to be designed, as opposed to an alarm instrument, but it also poses questions of stability, which have not yet been investigated in this context.

The shapes of the curves for other gases and vapours, including ethanol, methane and carbon monoxide are not dissimilar, though the magnitudes of I_B for given concentrations are dependent upon the particular gases involved. This point is best illustrated by plotting the effect of variations in V_H, and an appropriate family of curves is shown in Fig. 5. It is to be expected that similar concentrations of different gases lead to different values of I_B, having regard to the operation of the T.G.S., for each gas will have its own characteristic oxidizing capability. Hence, any instrument based on the T.G.S. must be calibrated in terms of a chosen gas or vapour, and for a general-purpose alarm, the gas to which it is least sensitive should be specified.

A second, and more important observation is that the characteristic convex-upward curves of I_B versus V_H have maxima which are marked functions of the particular gas involved, and weak functions of its concentration. For example, V_H for a peak response to 800 parts/10^6 carbon monoxide occurs at about 0·75 V, while for 400 parts/10^6 ethanol, it occurs at about 1·1 V. This immediately suggests that the T.G.S. is capable of some degree of selectivity to various gases based upon choice of filament heater voltage—a particularly important point where both carbon monoxide and light gases are intimately mixed as in motor vehicle exhausts.

A third observation—which has proved true for each T.G.S. so far tested—is that the value of V_H for greatest sensitivity, $V_{H(opt)}$, is also the value which gives the best linearity in I_B versus gas concentration plots such as those of Fig. 4.

Having established that I_B is a function of both gas concentration, C, and of heater voltage, V_H, it becomes
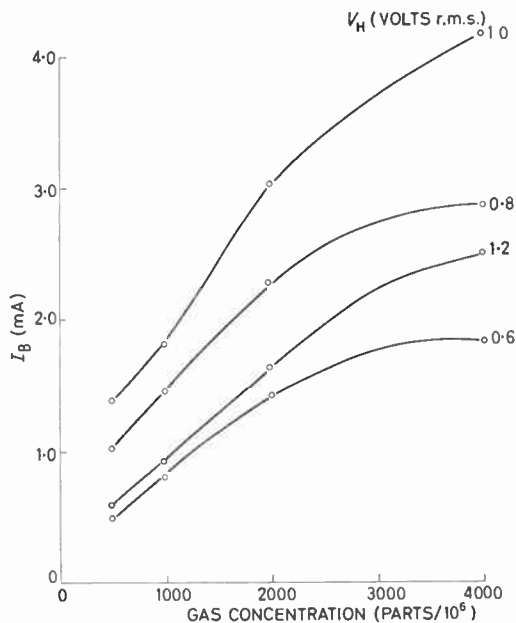

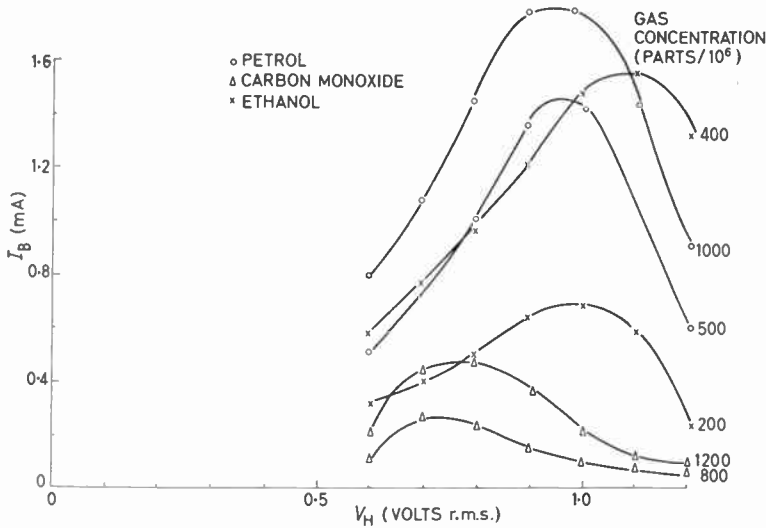Fig. 4. Bead current versus petrol vapour concentration for an H10 T.G.S.

Fig. 5.
Bead current versus heater voltage
for an H10 T.G.S.

relevant to ask whether it is necessary to stabilize V_H. Such a decision must be taken having regard to the nature of the instrument into which the T.G.S. is incorporated, and if a gas alarm is involved, then if $V_{H(opt)}$ is determined for the alarm level in question (e.g. 20% l.e.l.), this represents the 'best case' condition, because any subsequent variation in V_H will have the least effect on I_B.

From examination of curves such as those of Fig. 5, it is possible to extract an approximate expression for I_B in terms of C and V_H which is particularly valid for concentrations below about 1000 parts/10^6:

$$I_B \simeq (kC+Z)\left(\frac{V_H}{V_{H(opt)}}\right)^2 \qquad \text{for } V_H < V_{H(opt)} \qquad (3a)$$

$$I_B \simeq (kC+Z)\left(\frac{2V_{H(opt)}-V_H}{V_{H(opt)}}\right)^2$$
$$\text{for } V_{H(opt)} < V_H < 2V_{H(opt)} \qquad (3b)$$

where k = sensitivity of T.G.S. (I_B/C mA per part/10^6 at $V_{H(max)}$)

C = gas concentration in parts/10^6

Z = projection of I_B/C curves (at the Y-axis) when $V_H = V_{H(opt)}$.

For the T.G.S., the spreads in k and Z are considerable at the time of writing; further development of production techniques may improve this situation. The values given in Table 1 are relevant only to the device whose characteristics are plotted in Fig. 5, and the fit of a theoretical curve derived from equation (3a) is compared with an experimental plot in Fig. 6.

Table 1
Values of V_H, Z and k for an H10 sensor

		Petrol	Ethanol	Carbon monoxide
$V_{H(opt)}$	volts r.m.s.	0.95	1.1	0.75
Z	mA	1	0.01	0.01
k	mA/parts/10^6	0.85×10^{-3}	3.9×10^{-3}	0.32×10^{-3}

5 Basic Applications of the T.G.S.

Operation of the T.G.S. using the three basic power sources listed previously (mains, 12 or 24 V, or internal batteries) will now be considered.

Using mains, the simplest possible circuit, patented by Taguchi,[4] is shown in Fig. 7. Here, an autotransformer supplies V_H, while full mains voltage appears across the
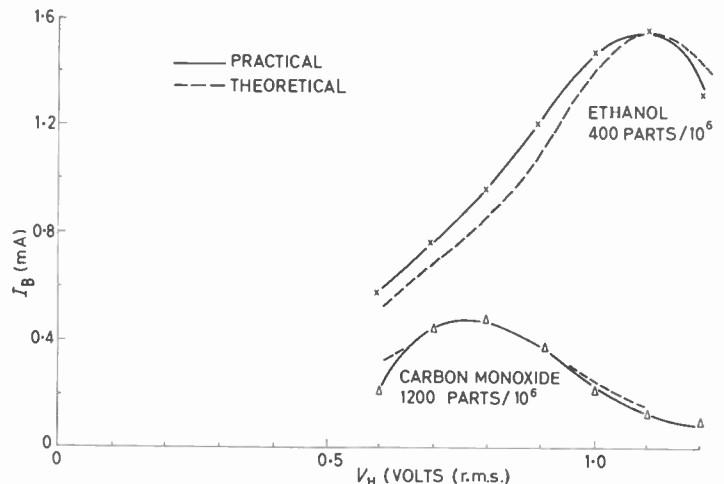

Fig. 6.
Comparison of practical and
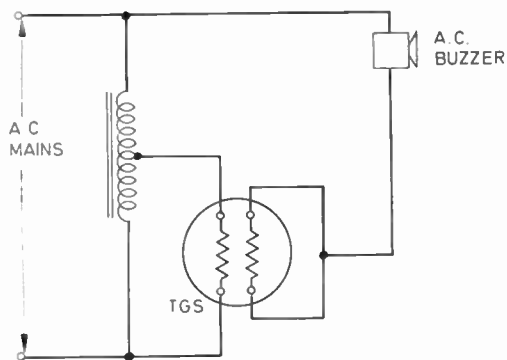theoretical I_B/C curves.

Fig. 7. Taguchi's original circuit.

bead and an audible alarm in series. With this arrangement, severe self-heating can occur if the resistance of the audible alarm is low, and if the T.G.S. is subjected to high concentrations of inflammable gas, even transiently, this can lead to avalanche breakdown, particularly if 240 V mains as opposed to 110 V is involved. Also, no sensitivity adjustment is provided.

A low bead voltage derived from a transformer can be used to provide inherent safety as shown in Fig. 8. This circuit[5] uses a thyristor which increases the sensitivity of the instrument considerably, and also provides half-wave rectification which allows a d.c. audible alarm to be used, along with a smoothing capacitor. The potentiometer is used to set the sensitivity of the instrument, which means that an alarm level of 20% l.e.l. can easily be obtained according to the specifications outlined above. Very little further circuitry is needed, but should include a means of protecting the thyristor gate from breakdown resulting from the negative-going half-cycles of the triggering waveform.

Simple circuits such as those described above have the disadvantage that the sensitivity of the T.G.S., being a function of V_H, is also dependent upon the mains voltage level. If the mains voltage is high or low by 10%, and the alarm is set for 20% l.e.l., it should nevertheless operate within $\pm 20\%$ of this nominal setting, according to the tentative gas alarm specifications suggested early in this paper. If a simple transformer of adequate size is used to supply the heater power, then V_H will vary by much the same percentage as the mains voltage, which
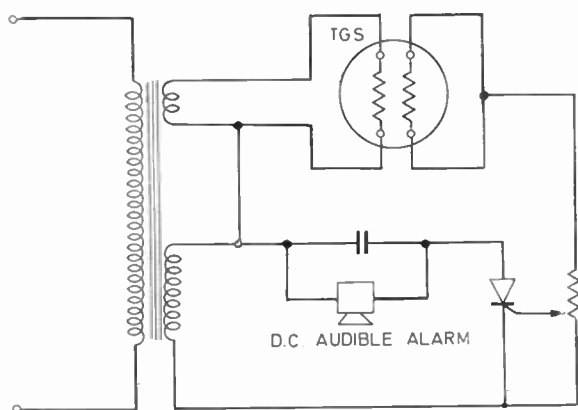

Fig. 8. Basic thyristor circuit.

allows curves such as those of Fig. 4 and 5 to be used to determine whether the alarm instrument will accommodate the $\pm 10\%$ mains variations for the T.G.S. used. Clearly, in a production situation proper account of parameter spreads for a given type of T.G.S. will be necessary, for $V_{H(opt)}$ is likely to have a unique value for each individual sensor.

If it is found necessary to stabilize V_H in order to achieve a more constant sensitivity than is possible by choosing $V_H = V_{H(opt)}$, it is unlikely that stabilization of the r.m.s. alternating voltage is the most economical method. If a nominal 6 V d.c. is produced by simple rectification and smoothing, a circuit such as that of Fig. 9 will be found to accommodate voltage variations of $\pm 20\%$ with ease. In this circuit, the Darlington pair is run as a constant-current source, and derives its base voltage from a lower value constant-current source provided by the f.e.t. current limiter, CL. Some degree
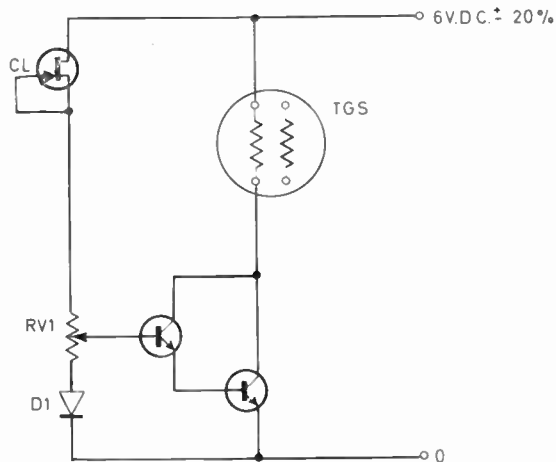

Fig. 9. A simple heater current stabilizing circuit.

of temperature compensation is provided by the diode, D1. (A constant V_B can easily be obtained using a Zener diode network.)

Where 12 V or 24 V storage batteries are to be used, the power loss in a series control element cannot usually be tolerated, so that it becomes necessary to utilize a pulsed power supply to the T.G.S. heater filament, having a true r.m.s. value corresponding to the chosen value of V_H. For battery voltages down to about 9 V, a simple unijunction oscillator operating a 'hammer driver' Darlington pair will be found to offer an appropriate duty cycle. Such a circuit is shown in Fig. 10(a). The pre-set resistor, RV1, may be adjusted so that an appropriate amount of power is dissipated in the T.G.S. heater, but it must be recognized that this power is very dependent upon the value of the supply voltage. This is not only because the p.r.f. is supply-dependent, but also because the energy per pulse injected into the filament is proportional to the square of the available voltage in the ON condition. Figure 10(b) illustrates a simple method of applying feedback to compensate for supply changes. The unilateral transfer element, u.t.e., may be either an indirectly-heated thermistor, or a calistor.[6,7] (The latter consists of a small filament lamp encapsulated along with

a photo-resistor, forming a photo-coupler having a long characteristic time-constant.) Referring to this calistor, the operation of the circuit is as follows.

If the r.m.s. value of the T.G.S. heater voltage falls, so does the light output from the calistor filament lamp. Hence, the resistance of the photo-conductor rises, which increases the current through transistor Tr1, which in turn increases the p.r.f. of the system. Thus the r.m.s. value of V_H is caused to rise.

Stability of the feedback loop is ensured by the dominant time-constant inherently provided by the calistor—an attribute possessed by the alternative indirectly-heated thermistor to an even greater degree. Further, both devices can be obtained in forms suitable for operation from filament voltages of r.m.s. values near to unity, which makes for a considerable reduction in component complexity over equivalent circuits using more conventional components.

The power level at which the filament operates may be preset using RV1, but it must be remembered that the photo-resistance presented by the calistor (or the bead resistance of the indirectly-heated thermistor) is very high when cold, so that the circuit will not start when switched on. A suitable starting network is provided by

R1 and R2, which turns Tr1 on via the diode D1. The values of R1 and R2 are such that after start-up, D1 becomes reverse-biased and only the photo-resistor (or thermistor bead) and R3 determine the base current to Tr1.

An alternative to the pulsed-heater systems described above is the blocking oscillator, or the two-transistor inverter. The stabilizing method described is easily adapatable to the latter, but unless the inverter transformer can be bought at quantity prices, the inverter instrument is likely to prove considerably more costly than the simple unijunction oscillator equivalent.
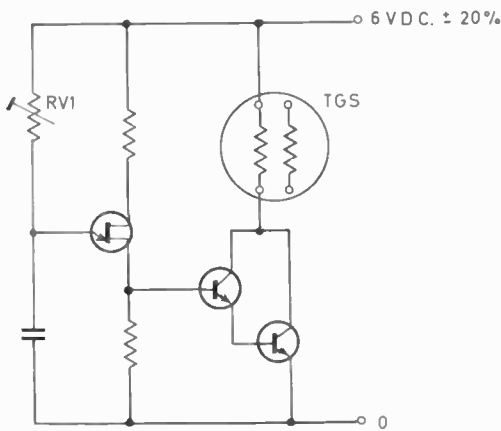
For completely portable operation, it is appropriate to consider the use of both primary and secondary cells as sources of heater filament power. The existing range of T.G.S. devices is such that any filament voltage from 0·7 V up to 1·5 V can be accommodated, depending upon the appropriate choice of both sensor and of V_H. This means that, in principle, any of the following (easily-available) cells might be employed:

Primary cells: Leclanché, mercury, zinc-air
Secondary cells: lead-acid, nickel-cadmium.

Of the primary cells, the Leclanché dry cell is the least appropriate because of its poor voltage regulation characteristics. It is however cheap, and large multiple-voltage units might be used along with constant-current circuits similar to that of Fig. 9, particularly where very short duty periods are involved. The mercury cell is much superior in terms of both voltage regulation and energy storage per unit volume capability. It is, however, expensive compared with the zinc-air cell, which will be shown later to provide a near-ideal power source for the T.G.S.

Insofar as secondary cells are concerned, the nickel-cadmium cell has advantages over the modern forms of jellied-electrolyte lead-acid cell, in that it delivers about 1·2 V compared with the 2·0 V of the lead-acid type, and is readily available in a wide variety of sizes.

An interesting direct comparison of cells is afforded if a pocket-sized gas alarm designed to take a 'D-size' cell is considered. According to the suggested specifications, a battery-operated gas alarm must operate in the

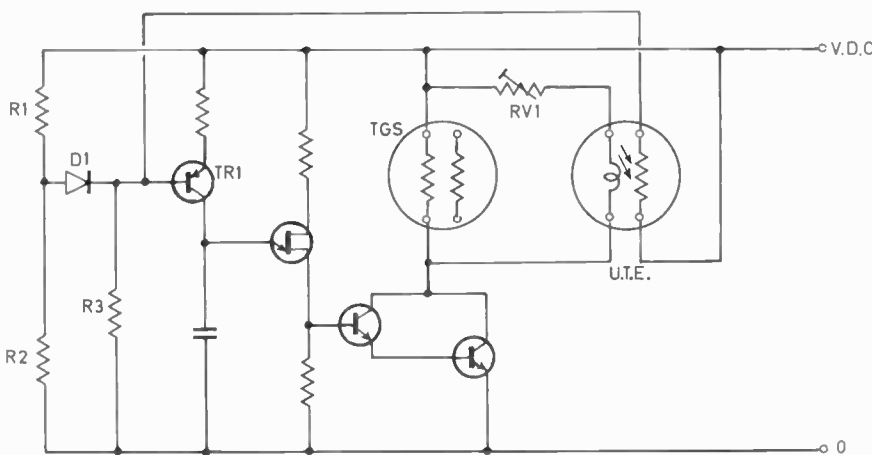(a) Unstabilized pulsed heater current supply.

Fig. 10.
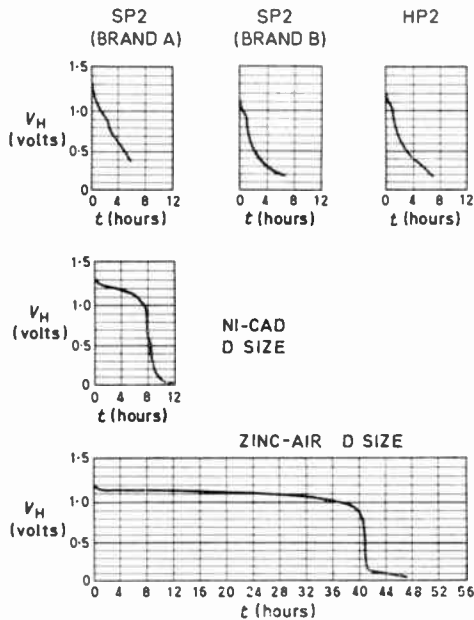(b) Stabilized pulsed heater current supply.

Fig. 11. Battery voltage across a T.G.S. filament versus time.

appropriate manner over voltages from 95% to 115% of nominal; that is, within a voltage range of 20%. Figure 11 shows that the voltage supplied by the D-size Ni-Cd cell remains within this range for some 6 hours; whereas the similarly-sized, but lighter, zinc-air primary cell will maintain its voltage for at least 36 hours. At a quantity unit-cost of about 50p, the zinc-air cell can clearly afford competition from the nickel-cadmium cell in circumstances where the establishment of charging facilities is undesirable.

In all cases, the provision of a small multiple-unit Leclanché or a mercury battery is necessary for the operation of the alarm circuit.

6 Conclusion

In that the T.G.S. is basically a highly-sensitive, rather than a highly-stable device, its chief field of application would appear to be in gas alarms, as opposed to gas concentration measuring instruments. This is in direct contrast to the Pellistor[8], which has achieved a high reputation when used in a bridge configuration in gas concentration meters. The measured parameter of the Pellistor, however, is the actual resistance of its (single) filament, which implies a comparatively low sensitivity, and the necessity of using a matched, but inactive, unit in the alternate bridge arm.

In the light of the measured linearity of the T.G.S. bead current, I_B, under the proper electrical conditions, as illustrated in Fig. 4, the device should not be dismissed in the gas concentration measurement field, however, and the results of current work on this aspect will be presented in due course.

7 Acknowledgments

The authors would like to acknowledge the co-operation of the Figaro Engineering Co. Ltd. of Shannon, Ireland; Ultramarine Instruments Ltd., Salford; and the ENMET Company Inc., of Ann Arbor, Michigan, in the preparation of this paper. They would also like to tender their thanks to the University College of Swansea, where the work was carried out, and to the Science Research Council which supported an M.Sc. project on this topic.

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Conference Report

Hybrid Microelectronics

University of Kent at Canterbury
25th—27th September 1973

The opening address by Dr. G. E. Weibel (Zenith Radio) was entitled 'Hybrid microelectronics revisited'. This was a stimulating review of the development of hybrids over the past ten years and a look into the future of the technology. Dr. Weibel highlighted the difficulties being caused in hybrid manufacture by the multiplicity of form of active devices being supplied by semiconductor manufacturers. He made a strong plea for standardization in this field and outlined the present ceramic/thick film technology being used at Zenith.

A survey of the development of hybrids in Germany was given by Dr. G. Kruger (Robert Bosch) who showed some modern applications, especially in the commercial and entertainment fields. He too expressed dismay at the cost and poor supply position of add-on active devices and suggested that new methods of encapsulation and bonding for semiconductor devices would be needed before hybrids can be accepted in the really large quantity areas, such as automotive electronics.

Mr. P. Moore (Erie Electronics) described the application of thin films in British communications equipment, particularly in portable transceivers. He emphasized that penetration of hybrids in this field was expanding rapidly but that progress was contingent upon close liaison with the semiconductor manufacturers.

In the session on deposition and materials the paper by Dr. J. T. Calow (Philips, Aachen) on $Zr B_2$ attracted considerable interest as the property data so far determined for this material imply some useful improvement over conventional materials. Lower drift and better tolerance performance than nichrome compositions seem to have been established, although the work has not yet been carried far enough to give confidence in this material under volume production conditions.

The interaction between materials in thick film components was considered in two ERA papers by B. Walton and I. G. Bowkley, covering the structure of resistor and capacitor elements and several questions were concerned with the differences in performance that the new compositions promised. It seems possible that the capacitor materials could improve printed component yields but the necessity of using special metallization for the upper electrode has to be considered as a restriction in their use and must clearly be taken into account when selecting the compositions for the rest of the circuit. The ability to achieve high values in printed components is often sufficiently attractive however to justify this constraint.

The Conference was arranged by the IERE with the association of the International Society of Hybrid Microelectronics (UK), the Institution of Electrical Engineers (Electronics Division), the Institute of Electrical and Electronics Engineers (UK and Republic of Ireland Section) and the Institute of Physics. This review of the Conference is based on reports prepared by the chairmen of the various sessions, the majority of whom were members of the Joint Organizing Committee and included Prof. J. C. Anderson, Dr. P. G. Barnwell, Professor D. S. Campbell, Mr. L. G. Cripps, Dr. R. G. Loasby, Mr. B. C. Waterfield and Dr. M. J. Wood.

The full list of titles of the papers presented at the Conference was published in the July 1973 issue of *The Radio and Electronic Engineer*. The complete set of papers of Conference Proceedings No. 27 may be obtained from the IERE Publications Department, price £8.00.

The new resistor materials based on molybdenum borides offer advantages, principally in cost, over the noble metal based ones and the discussion on this topic failed to show up any demerits in this system. Development work is continuing at ERA and if the present promise is fulfilled commercial exploitation will follow, probably through a licensing arrangement.

The merits of plasma spraying as a material deposition technique were discussed at length after the CNET paper. Although photo-etching is a necessary additional step in pattern forming the process has clear attractions in depositing unstable elements, such as copper and barium titanate, and may therefore find application as a complementary process to the established ones.

Overall, this session indicated that the ranges of new materials and techniques being developed specifically to overcome the shortcomings of present systems show considerable promise. The indications are that a wider range of materials will be usable in film circuits in the future and that some of the cost pressures on current systems will be relieved through the utilization of base metals.

The important, though highly specialized subject of adhesion and bonding was dealt with in a group of four papers, the first of which considered degradation of thick film conductor adhesion, while the other three examined methods of bonding active devices to thick film circuits using conducting.

Two papers were presented in the session on thermal design and demonstrated a much more mature approach to power hybrid techniques than the empirical methods which have largely dominated the field up to now. Both the ability to predict temperature distributions and to construct low thermal resistance assemblies will greatly affect the technology in the future.

D. J. Dean (AWRE) showed how temperature distributions in hybrid circuits could be predicted by use of the principles of superposition and reciprocity. He concluded his paper by showing examples of temperature distributions that had been obtained in the course of his work.

J. T. Hughes and M. R. Naylor (ICL) discussed various methods of obtaining high power density integrated circuit assemblies. Predicted temperature distributions obtained for alumina-based assemblies were compared with those obtained using the elegant copper substrate assembly that the authors' company are pioneering.

There is an increasing interest in the application of hybrid techniques to the microwave field and papers in this session centred around means of applying the microstrip technology.

Two papers from Brighton Polytechnic reported studies of the effect that surface topography has on transmission characteristics and somewhat unique was the examination of the underside of the metal coating at the ceramic interface. Lines etched into total coated substrates gave an overall response of 0·5 dB over the whole working range. Measurements were not made to gauge the effect of edge definition and Transcote had been found to give the overall best print results; line thickness increase did not improve the results. A number of questions were put to Dr. P. G. Barnwell regarding inductor design and the work carried out appeared to tie in with Grover's formulae. All work carried out in these papers was based on thick film technology.

The work reported by Dr. W. Schilz (Philips, Hamburg) was well appreciated and promoted considerable questioning. He stated that the resistors shown on the substrate (2·5 cm square) were 1 mm wide and overlap was not critical since the 'T' network was short compared to wavelength and gave resistor values of 3% unadjusted. There was obvious interest in the dielectric that had been used for the overlay coupler and it appeared that the glass had a matching K with the alumina and was particularly low loss. Work carried out so far on X-band was not necessarily the limit for thick film technology, the high dielectric constant of the substrate itself was a greater limit and higher frequencies therefore favoured materials such as BN_2 or BeO_2 with a dielectric constant of 4. It was also interesting to hear of success using direct metal foil screens, a technique so far found to be unacceptable in this country.

A paper from the Post Office Telecommunications Headquarters introduced thin film technology to the Session and gave reasons for the choice of this process. The questions were detailed and queried a number of the techniques employed, for example, the sputtering power for NiCr film was limited to 110 W input with a 15 cm diameter target; varying NiCr did not appear to affect the performance even at 20 GHz. They had found the best photoresist to prevent edge build up was Shipley AZ 130H, which gave a build up of up to 15 μm. The width of the 3 dB coupler lines depended on substrate thickness and a maximum plating of 6 μm. Devices were normally attached to the circuit using 60/40 SnPb solder with hot gas iron and gave no failures, but they had experienced wetting problems with AuPb. On the other hand, higher temperature gold-based solder was used for attaching to the ground plane. Much work had also been carried out to assess suitable substrate materials and a current favourite was polished hot pressed alumina from a UK source.

Surface acoustic waves are yet another new technology which was possibly foreign to a number of those present and the paper from the University of Edinburgh went some way towards describing the basic functions and fabrication techniques. The fine line working emphasized the necessity of extreme care and bonding wires from bus bar to substrate was a hazardous task, particularly since the metallization thickness of 0·2 μm precluded the use of ultrasonic bonding.

The final part of the session on high frequency and microwave contained three papers, all concerned with applications below 1 GHz. The first, read by P. Williams (AB Electronics), in the absence through illness of Dr. T. L. Harcombe (Glamorgan Polytechnic), described commercial applications in television tuners. The potential advantages of thick films in a u.h.f. television tuner justified the adoption of hybrid techniques, and the paper then discussed the alternative forms of construction available, concluding that a strip line resonant circuit with packaged discrete devices added, offers the best

compromise between performance and cost. Finally, data were presented for the overall tuner performance showing a gain of 16 dB at 860 MHz with 57 dB image rejection.

D. P. Heywood (RAE Farnborough) then described some extremely novel 150–200 MHz oscillators fabricated in thick film. By sensing the substrate temperature with a printed thermistor, temperature stabilization could be incorporated, thus improving the stability of the oscillator. By the correct choice of oscillator configuration and thermal stabilization circuit, an oscillator having a stability of ± 30 parts in 10^6 over a 50 deg C temperature range had been constructed.

The final paper of the session, by Dr. R. Chaddock (Mullard), dealt with the application of lumped element techniques to high frequency hybrid integrated circuits. A large amount of data was given on inductors operating between 300 MHz and 1 GHz and the application of these inductors was described. Particularly interesting was the use of both thick film and thin film inductors in matching networks for r.f. power transistors. Low-cost broadband transistor modules were shown to illustrate this application.

The session on failure and reliability opened with a paper from M. V. Coleman (STL) covering the analytical techniques that are available for examining the structure and composition of thick films. He illustrated the use of these techniques by reporting on the examination of Dupont 7800 and 1100 resistor series, finishing up with schematic representations of the distribution of the various components of the complete bodies. The paper was a very useful summary of the analytical techniques and the results illustrated the difficulty of analysing electrical conduction processes in the bodies.

Professor J. C. Anderson's group at Imperial College discussed the use of third harmonic index measurements on screen-printed thick film and also plasma sprayed resistors. It was shown that t.h.i. measurements revealed differences between the two types but it is not yet possible to identify these measurements with any conduction modes and hence failure mechanisms.

Mrs. J. B. McCloghrie and Dr. Peter Holmes (RAE) reported on the reliability of very small (0·5 × 0·25 mm) resistors. They found that the behaviour was sufficiently reproducible to allow empirical design rules to be drawn up for such units. Care must be taken not to pass too high a current through such devices.

The designer's need for reliability information was stressed by Miss Pat Shove (ASWE). Although no available data were discussed, the questions that must be answered were posed in her short contribution.

The migration behaviour of silver in a silver/insulant/gold structure was reported by J. Savage (AWRE). With an applied field Ag will migrate through the glass insulant as Ag_2O and when it reaches the Au will dissolve in it liberating oxygen in the form of gas bubbles. This type of interaction can obviously be a major cause of failure.

The papers in the session on hybrid circuit application dealt with an extremely varied range of topics. It started with a brief but interesting contribution from P. E. K. Donaldson and E. Sayer (MRC Neurological Prostheses Unit) describing how they are using hybrid techniques to increase the complexity of a visual prosthesis while keeping size within bounds.

Power applications were considered next by Dr. M. J. Wood (Westinghouse). His objective was to complement the falling cost of mains voltage switching semiconductors with cheaper control circuitry, by using thick film methods. Modules designed included a manually-controlled open-loop a.c. regulator, a closed-loop a.c. regulator, and a series of

voltage regulators typically containing 25 components. Power semiconductors are mounted on a copper heat spreader bonded to the ceramic (usually alumina) fixed to a mild steel base plate. Overall thermal resistance (device junction to base plate mounting screws) using soft solder for assembly was typically 2·4 deg C/W. Adhesion tests for the printed patterns and the substrate-base plate interfaces were described, and provoked some discussion.

D. S. Larner (Plessey Telecommunications Research) described the application of hybrid circuits for 120 Mb/s repeaters, where reliability is important. Flat pack transistor arrays were used, and an interesting constructional point was that two parts of the circuit were electrically identical but physically mirror images—so preparation of some of the artwork was simplified.

A high-level fast limiting switch was described by C. H. Jones (SRDE). Swings of 30V with edges of 20 ns over a wide operating temperature range were achieved with unencapsulated chips wire bonded or soldered on to a thick film substrate.

D. H. Danielsson (Datasaab, Sweden) had carried out an extensive comparison of epoxy-glass and thick film multilayer boards as substrates for d.i.l. packages. Especially, he had carried out environmental tests in accordance with IEC 52 (CO) 76. Thick-film boards showed very good results in all tests. Boards measuring 9·65 cm × 14·7 cm carrying 54 d.i.l. packages each with 16 pins were described. Production and servicing aspects, and adhesion of the d.i.l. packages to the substrate were considered. In the discussion, Dr. Danielsson explained that a special soldering iron had been developed for replacing the d.i.l. packages.

Applications described in the final afternoon session of the Conference included a guidance unit for a guided missile constructed in thick film by BAC; a hybrid module for monitoring electronic equipment comprising operational amplifier and a multi-channel switch (Dr. P. L. Kirby of Welwyn Electric); and a group of papers describing hybrid inter-connexion techniques for computer circuits presented by R. C. Bingham (ICL).

International Air Traffic Control

Part of the air-traffic control room at Eurocontrol, Maastricht. Primary and secondary radar information employing an extensive computer complex is used to maintain direct radio contact with aircraft over Northern Europe. Eurocontrol controls civil aviation air traffic only above the 6000 m (20 000 ft) limit in the air space over Belgium and Luxembourg. In the near future, this Maastricht control centre will also be in charge of air traffic over Northern Germany, Holland and part of the North Sea. Decca, Plessey and other British firms have supplied equipment for various parts of the Eurocontrol organization; air conditioning for the computers and personnel is served by cooling equipment supplied by the York-Europe Division of Borg-Warner Corporation, who have issued this photograph.

Generalized duals, generalized inverses and their applications

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SUMMARY

Generalized duals and generalized inverses are defined for any general network N consisting of linear/non-linear, time-invariant/time-varying, passive/active, lumped/distributed elements. A simple method of obtaining these directly for a general planar three-terminal network is given. Applications of these concepts in network synthesis are discussed.

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1 Introduction

The concept of dual and inverse networks plays an important role in the study of electrical networks. The technique of obtaining the dual of a planar network by the 'dot-window method' and replacing the resistors of R_i ohms by resistors of $1/R_i$ ohms, inductors of L_i henrys (capacitors of C_i farads) by capacitors of L_i farads (inductors of C_i henrys) and voltage sources of E_i volts (current sources of I_i amperes) by current sources of E_i amperes (voltage sources of I_i volts), are well known. The solution of the original network yields the solution of the dual also, since the voltages and currents in the different elements of the dual network are respectively the currents and voltages of the corresponding elements in the original. For example, the solution of a series RLC resonant circuit automatically provides the solution for the dual RLC network, namely a parallel resonant circuit. The concept of duality seems to have been mainly exploited in the design of RLC one-port planar networks by observing that in such dual networks the driving point admittance (d.p.a.) $Y_1 = 1/Z_1$ of one is the same as the driving point impedance (d.p.i.) Z_2 of the other, that is $Z_1 Z_2 = 1$.
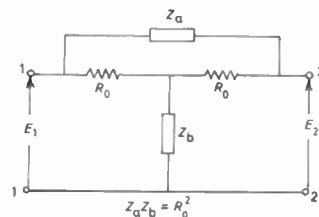

Fig. 1. Constant resistance bridged-T network.

The idea of inverse networks has been found to be also useful in the design of one-ports. Since there appears to be some confusion between dual and inverse one-ports, it is worthwhile discussing these concepts briefly.

Two one-ports are said to be inverses of each other if the d.p.i. Z_1 of one is the d.p.a. Y_2 of the other. Hence dual one-ports are always inverses while the converse is not necessarily true. For example, suppose we want to realize a one-port whose d.p.i. Z_2 is $1/Z_1$, where Z_1 is the d.p.i. of an LC-network, realized in the Foster's first form.[1] The dual of this one-port automatically realizes Z_2, the corresponding network being recognized as the Foster's second form[1] for Z_2. However, Z_2 may also be realized by Cauer's forms[1], but in this case even though the networks for Z_1 and Z_2 are inverses of each other, they are not in fact duals.

One of the main applications of the dual and inverse networks has been in the design of constant-resistance networks as equalizers and filters. For example, the network of Fig. 1 when loaded by a resistor R_0 has an input impedance R_0 and a voltage transfer function $T(s) = R_0/(R_0 + Z_a)$ provided‡ $Z_a Z_b = R_0^2$. In designing such networks, Z_a is extracted from the given transfer

‡ Two one-ports having this property are said to be inverses of each other with respect to R_0^2.

function and realized as a one-port. The impedance Z_b may now be directly realized by simply taking the dual of the network corresponding to Z_a and scaling the elements suitably.[1] However, Z_b has to be synthesized separately, if the network for Z_a is not planar.

As mentioned earlier the above concepts of dual and inverse networks are restricted to RLC-networks. These ideas are not directly applicable to active RC-networks, since the dual of a capacitive branch yields an inductive one.

For linear time invariant (l.t.i.) networks, these concepts have been recently generalized, in the frequency domain, by Mitra et al.[2,3] They define the generalized inverse of an n-port l.t.i. network N to be a network N' for which the impedance matrix $[Z]$ is $f(s)$ times the admittance matrix $[Y]$ of N. It should be mentioned that the topologies of N and N' may or may not be related. They also define the generalized dual of a network N consisting of one-port elements and controlled sources to be a network N* for which the topology is dual of that of N and each element is the generalized inverse of the corresponding element of N. Further they have considered some applications of capacitive inverses and duals† in lumped active-RC network design.

It should be emphasized that the above definitions for generalized inverses and duals are applicable only to l.t.i. networks in the frequency domain. In this paper, we first extend these definitions for a general n-port consisting of linear/non-linear, time invariant/time varying, lumped-distributed elements. We then give a simple method of obtaining the generalized dual of a general 'planar three-terminal (3-T) two-port', consisting as subnetworks of not only one-ports but also 3-T two-ports. It should be noted that for such networks the familiar technique of obtaining the dual networks is not directly applicable, since it is restricted to networks containing one-ports only as elements. In the method given here, not only the one-ports but also the two-ports are treated as blocks and replaced in the generalized dual by their corresponding generalized inverses. Finally, some applications of the generalized duals and inverses in synthesis are considered.

2 Some Definitions

Generalized Inverse: Let V_p and I_p be the vectors of port voltages and currents of a general n-port network N, consisting of linear/non-linear, time invariant/time varying, passive/active, lumped/distributed elements. Then we define the generalized inverse N' to be a n-port network, for which the port voltages V'_p and current I'_p are given by $V'_p = \lambda_1 I_p$ and $I'_p = \lambda_2 V_p$, λ_1 and λ_2 being linear operators.

Generalized Dual: Consider a general planar network N. Let G be its directed graph, where the edge e_i corresponds to the branch whose voltage and current are V_i and I_i, and the orientation of e_i corresponds to the direction I_i. Let G* be the dual of the directed graph G, where e_i and e_i^* are the corresponding edges in G and G*. Then we define the generalized dual of the network N to

† Capacitive dual (inverse) is obtained from the generalized dual (inverse) when $f(s) = k/s$.

be the network N* if the directed graph of N* is G*, and the different branch voltages and currents (V_i^*, I_i^*) represented by the directed edges e_i^* are related to the corresponding branch voltages and currents (V_i, I_i) in N by $V_i^* = \lambda_1 I_i$ and $I_i^* = \lambda_2 V_i$, where λ_1 and λ_2 are linear operators.

It should be noted that the conventional dual[4] is a special case for which $V_i^* = I_i$ and $I_i^* = V_i$.

Graph of a 3-T Two-port: Consider a 3-T two-port, as shown in Fig. 2(a). We shall define the graph of the two-port to be that shown in Fig. 2(b), the edges representing the port voltages and currents.
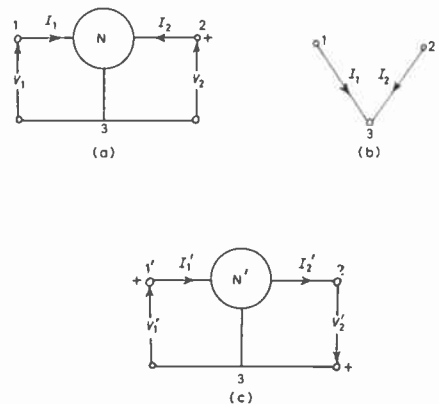

Fig. 2. (a) 3-T two-port network. (b) Graph of 3-T two-port network. (c) Generalized inverse of 3-T two-port network.

Since, according to our definition, the port-variables of the generalized inverse N' are arbitrary, we get different realizations for N' depending on our choice for the polarities of the port-variables. However, we shall choose for N', the polarities of the port-variables as shown in Fig. 2(c). Hereafter in this paper, by the generalized inverse of a 3-T two-port (Fig. 1(a)) we mean another 3-T two-port, with polarities as shown in Fig. 2(c), with $V'_p = \lambda_1 I_p$ and $I'_p = \lambda_2 V_p$ ($p = 1,2$). The reason for this choice of polarities will be clear in the next Section.

A planar 3-T Two-port: Let N be any 3-T two-port consisting of one-ports and 3-T two-ports as subnetworks. Then we shall define N to be planar, if the graph obtained after (i) replacing each of the internal one-ports and 3-T two-ports by their graphs and (ii) adding external edges corresponding to the two-ports of N, is planar.

3 Generalized Dual of a Planar 3-T Two-port

We shall now give a simple method of obtaining the generalized dual of such a planar 3-T two-port. Consider a planar 3-T two-port N consisting of one-ports and 3-T two-ports as subnetworks (for example, the network of Fig. 3(a)). Obtain the graph G of N as shown in Fig. 3(b). The direction of the edges in G represent the direction of the current through the corresponding port, whether it be internal or external: the tips of the arrow represent the negative (positive) polarities for the voltages across the internal (the two external) ports. We obtain now the dual graph G* of the graph G in the conventional way.
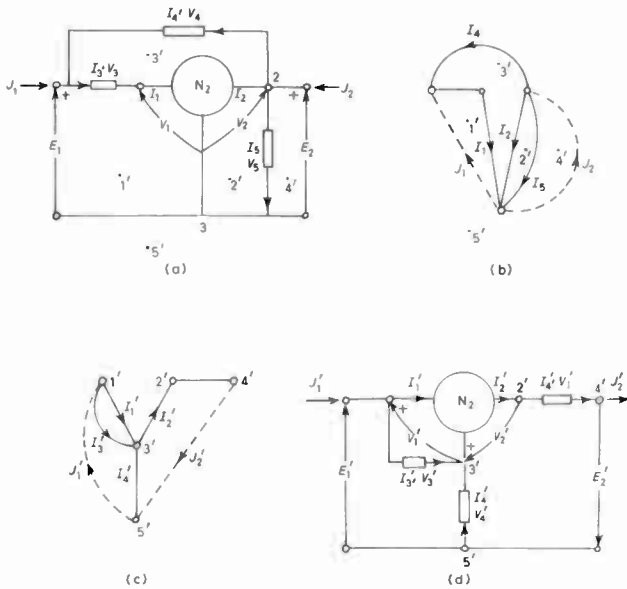
Fig. 3. (a) A planar 3-T two-port network N. (b) Graph G of the 3-T two-port network N. (c) Dual graph G* of the graph G. (d) Generalized inverse of (a).

We assume that corresponding to an internal (external) port edge e in G, the dual edge e* in G* is convergent on a node n' if the orientation of e is in an anticlockwise (clockwise) direction with respect to n', n' being the node corresponding to the window in which e appears. G* for the graph of Fig. 3(b) is shown in Fig. 3(c). Now, let

$$V_b^* = \lambda_1 I_b, \qquad I_b^* = \lambda_2 V_b \qquad (1)$$

where λ_1 and λ_2 are linear operators. Hence, it is seen that corresponding to a branch in G representing an internal one-port N_1 in N, the dual branch in G* represents a one-port which is the generalized inverse of N_1. Also, the edges 1'3' and 2'3' together may be recognized as 3-T two-port which is the generalized inverse of the corresponding 3-T two-port in N.

Let $B(B^*)$ and $A(A^*)$ be the circuit and incidence matrices of $G(G^*)$. Then we have[4]

$$A^* = B, \qquad B^* = A. \qquad (2)$$

The matrices B and A can be suitably partitioned to write the Kirchhoff's equations in the form

$$\begin{bmatrix} U & B_1 \\ 0 & B_2 \end{bmatrix} \begin{bmatrix} E \\ V_b \end{bmatrix} = 0; \qquad \begin{bmatrix} U & A_1 \\ 0 & A_2 \end{bmatrix} \begin{bmatrix} J \\ I_b \end{bmatrix} = 0 \qquad (3)$$

where $E(J)$ and $V_b(I_b)$ are the two external port voltage (current) and the internal port voltage (current) vectors. Using equations (2) and (3), the Kirchhoff's equations for G* may be written as

$$\begin{bmatrix} U & B_1 \\ 0 & B_2 \end{bmatrix} \begin{bmatrix} J^* \\ \lambda_2 V_b \end{bmatrix} = 0; \qquad \begin{bmatrix} U & A_1 \\ 0 & A_2 \end{bmatrix} \begin{bmatrix} E^* \\ \lambda_1 I_b \end{bmatrix} = 0. \qquad (4)$$

Hence,

$$E^* = \lambda_1 J, \qquad J^* = \lambda_2 E. \qquad (4)$$

Thus, the 3-T networks corresponding to G and G* are generalized duals of each other; of course they are also generalized inverses of each other.

It is now noted that the above method of obtaining the generalized dual is quite general and applicable to any planar 3-T two-port, since no assumptions are made as to the exact nature of the A and B matrices.

Also, it is not necessary to draw the graph of N to get N* and hence N'. It may be obtained directly as follows:

(i) Add external edges e_1 and e_2 corresponding to the two ports of N.

(ii) Place nodes in each of the windows. Designate by $1_i'$, $2_i'$, and $3_i'$, the nodes corresponding to the windows for which the arcs $1_i 3_i$, $2_i 3_i$, and $1_i 2_i$ are respectively parts of the boundaries. The other nodes may be designated in any manner.

(iii) For the generalized dual network N*, connect between the nodes $1_i'$, $2_i'$ and $3_i'$, 3-T two ports which are the generalized inverses of the corresponding 3-T two ports in N.

(iv) The one-ports of N* are located with respect to its nodes, in the conventional way. These one-ports are the generalized inverses of the corresponding ones in N.

(v) The two external ports of N* (and hence that of N') are located in the usual manner, bearing in mind the sign convention used in this paper.

The procedure is illustrated by taking directly the generalized dual of the network of Fig. 3(a) and is shown in Fig. 3(d).

4 Applications to Linear Time-invariant Networks

In the definition of a generalized inverse, choosing λ_1 and λ_2 to be the convolution operators for l.t.i. networks, the relationship between the currents and voltages of a given network N and its generalized inverse, N', reduces to

$$I'(s) = f_1(s) V(s), \qquad V'(s) = f_2(s) I(s). \qquad (5)$$

For a one-port network N, (5) reduces to

$$Y'(s) = [1/f(s)] Z(s) \qquad (6)$$

where

$$f(s) = f_2(s)/f_1(s). \qquad (7)$$

Z and Y' represent the impedance of N and the admittance of the corresponding generalized inverse N'.

For a 3-T two port N with a chain matrix $[A, B, C, D]$ we have

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix}. \qquad (8)$$

Substituting (5) in (8) and rearranging, we have for the generalized inverse, N'

$$\begin{bmatrix} V_1' \\ I_1' \end{bmatrix} = \begin{bmatrix} D & f(s)C \\ B/f(s) & A \end{bmatrix} \begin{bmatrix} -V_2' \\ I_2' \end{bmatrix}. \qquad (9)$$

Bearing in mind the polarities of the port voltages and currents for N' (Fig. 3(c)) we now see that the chain matrix of N' is

$$[a]' = \begin{bmatrix} D & C/f(s) \\ B/f(s) & A \end{bmatrix}. \qquad (10)$$

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Thus, for a given planar l.t.i. 3-T two-port N consisting of one-ports and 3-T two-ports as subnetworks, the generalized dual (and hence the generalized inverse, N') may be obtained by the method given previously. Also the chain matrix of N' will be related to that of N as in (10). It follows immediately that:

(i) The voltage transfer function (v.t.f.) of N becomes the current transfer function (c.t.f.) of N' and *vice-versa*. The sensitivities of these two functions are the same with respect to the corresponding parameters.

(ii) The $[z]$ of N is related to the $[y]$ of N' as

$$[y]_{N'} = \frac{1}{f(s)} \begin{bmatrix} z_{11} & -z_{12} \\ -z_{21} & z_{22} \end{bmatrix}. \tag{11}$$

Since our procedure of obtaining the generalized dual of a network N involves replacement of the internal two-ports by their generalized inverses, the generalized inverses of the different active as well as distributed elements are tabulated in Table 1. We note from Table 1 the following interesting points: (i) The capacitive inverse $(f(s) = 1/s)$ of an RC-line and the resistive inverse of an LC-line are respectively the dual lines.[5] (ii) The inverse of a voltage controlled voltage source is a current controlled current source with the same amplification factor and *vice-versa*, irrespective of the nature of $f(s)$, similar results being true for negative impedance converters (n.i.c.) and generalized impedance converters (g.i.c.).

The simple procedure of obtaining the generalized inverse, in conjunction with Table 1, can be used (i) to obtain directly realizations for a c.t.f. given that of a v.t.f. and *vice-versa*, and (ii) to obtain alternate structures for driving point functions.

Example:

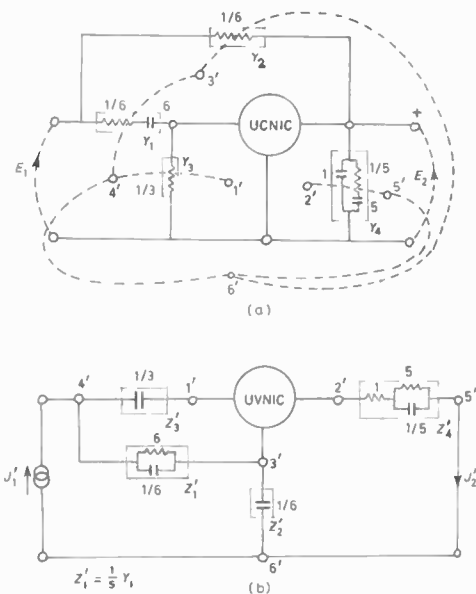Consider the network of Fig. 4(a) where the different one-ports are RC one-ports and the internal two-port

Table 1
Generalized inverses for some active and distributed elements

is a unity gain current inversion type n.i.c. (u.c.n.i.c.). This is the Yanagisawa's structure[3] realizing the voltage transfer function

$$\frac{E_2}{E_1} = \frac{6}{s^2 + 3s + 3} \tag{12}$$

and corresponds to a low-pass filter with a d.c. gain of 2, angular cut-off frequency of 1·85 rad/s and a roll-off of 12 dB/octave.

Its capacitive dual may be found directly by using the procedure given above and is shown in Fig. 4(b), where the different one-ports are the capacitive inverses $(f(s) = 1/s)$ of the corresponding ones in Fig. 4(a) and the internal two-port is a 3-T unity gain voltage inversion type n.i.c. (u.v.n.i.c.), the capacitive inverse of u.c.n.i.c. of Fig. 4(a). It is readily seen that Fig. 4(b) realizes the c.t.f.

$$\frac{J_2'}{J_1'} = \frac{6}{s^2 + 3s + 3} \tag{13}$$

which is the same as the v.t.f. realized by Fig. 4(a). It is noted that Fig. 4(b) is nothing but the Thomas's Structure[3] realizing the c.t.f. (13).

Fig. 4. (a) Yanagisawa's structure realizing the v.t.f. (12).
(b) Capacitive inverse of the structure of (a).

5 Conclusions

For the l.t.i. network, the generalized inverse defined in this paper may be used in conjunction with the network transposition[7] in not only obtaining two realizations for a c.t.f. (v.t.f.) from that of a v.t.f. (c.t.f.) but also an alternate realization for the v.t.f. (c.t.f.) itself. In addition, we can also obtain three alternate structures for a d.p.f. starting with a given one. It is hoped that these concepts would serve to unify the various RC-active structures which seemingly look different, but in fact are interrelated through these operations; for example, the two Yanagisawa and the two Thomas structures are related.

6 Acknowledgments

7 References

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STANDARD FREQUENCY TRANSMISSIONS—December 1973

(Communication from the National Physical Laboratory)

Dec. 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)		Dec. 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	+0·3	0	0	696	597·7	17	+0·1	0	−0·3	691	590·9
2	−0·1	0	0	697	597·9	18	+0·1	+0·1	−0·2	690	590·2
3	−0·1	0	0	698	597·9	19	0	0	0	690	588·8
4	+0·1	0	0	697	597·0	20	+0·1	0	0	689	588·6
5	0	0	0	697	596·7	21	−0·1	0	−0·1	690	588·9
6	0	0	0	697	596·4	22	−0·2	+0·1	−0·1	692	588·3
7	+0·1	0	0	696	596·2	23	0	0	−0·1	692	588·3
8	−0·1	+0·1	−0·1	697	595·3	24	0	+0·1	−0·1	692	590·3
9	0	0	−0·2	697	594·9	25	0	0	0	692	587·9
10	+0·3	0	−0·3	694	594·7	26	0	+0·1	—	692	587·2
11	−0·1	0	0	695	594·3	27	+0·2	0	—	690	585·6
12	−0·1	+0·1	−0·2	696	594·6	28	+0·2	0	−0·1	688	585·3
13	+0·3	+0·1	−0·2	693	593·9	29	−0·1	0	−0·1	689	585·3
14	+0·1	+0·1	−0·2	692	592·3	30	0	0	−0·1	689	584·4
15	0	0	−0·2	692	590·9	31	+0·1	0	−0·1	688	584·3
16	0	0	−0·2	692	590·5						

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.

IERE News and Commentary

New Chairman and Vice-Chairman for CEI

Major-General Sir Leonard Atkinson, K.B.E., B.Sc.(Eng.), C.Eng., and Professor J. F. Coales, O.B.E., C.Eng., F.R.S., took office as Chairman and Vice-Chairman respectively at the Council of Engineering Institutions' Annual Meeting on 24th January. Sir Leonard was President of the IERE from 1968–70, and Professor Coales, who is Professor of Engineering (Control) at Cambridge, is a Past President of the Institution of Electrical Engineers.

Sir Leonard, who succeeds the present Chairman, Sir Angus Paton, C.M.G., C.Eng., F.R.S., was commissioned into the RAOC in 1936 and in 1942 transferred to the newly-formed Royal Electrical and Mechanical Engineers and saw active service in north-west Europe, India and the Far East. From 1958 to 1963 he was Commandant, REME Training Centre and from 1963 to 1966, Director, Electrical and Mechanical Engineering (Army) in the Ministry of Defence.

On his retirement in 1966, Sir Leonard was appointed a Director of Harland Engineering Limited, Alloa, and a Director of Simon Electronics Limited, London. At the end of 1969 he became Managing Director of the newly-formed Harland Simon Limited, Bletchley, and a Director of Weir Engineering Industries Limited, Glasgow. He joined the Board of United Gas Industries Group in 1972 and also continued as a Member of the Board of Weir Engineering Industries. Sir Leonard was appointed Chairman of the DTI Committee on Terotechnology when it was set up in 1970 and is also a member of the DTI Committee on Industrial Technologies. Among his other public offices, he is a member of the Court of Bradford University and a Governor of Reading Technical College.

Professor Emrys Williams

As this issue of the Journal was being prepared for the press, news was received of the death of Professor Emrys Williams (Past President) on 13th February. A full obituary will be published in the March issue of the Journal.

Dr. von Braun to give Clark Maxwell Lecture

The eighth Clerk Maxwell Memorial Lecturer will be Dr. Wernher von Braun. His lecture will have the theme 'Our space programme after Apollo' and it will be given on Thursday, 21st March 1974, at 6 p.m. in the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1. Dr. von Braun who is now Vice-President, Engineering and Development, with Fairchild Industries was formerly Deputy Associate Administrator, National Aeronautics and Space Administration. He is a pioneer in space flight and before joining NASA held technical appointments in the US Army Guided Missile Development Division.

The Marconi Heritage

On Thursday, 25th April 1974, a one-day colloquium 'The Marconi Heritage' is being held at Savoy Place to celebrate the centenary of the birth of Guglielmo Marconi, G.C.V.O. The event is being organized jointly by the IEE and the IERE in association with the Radio Society of Great Britain.

Five lectures will be presented during the course of the day which will survey the life of Marconi and the impact of his work on technology past and present. Aspects of his early life will be presented by Professor W. P. Jolly, a recent biographer of Marconi, in a lecture entitled 'The making of the man'. The latter years of his life will be covered by Mr. G. A. Isted, Marconi's personal assistant for a number of years in the 1920s and the early 1930s, with his contribution 'A turning point in radio communications'. The three lectures 'Radio in global telecommunication' by Mr. R. G. Halsey, 'Maritime radio communications' by Mr. G. J. McDonald and 'Scientists' reactions to Marconi's transatlantic radio experiments' by Mr. J. A. Ratcliffe, F.R.S. will have a less historical emphasis and will concentrate more on the modern technology stemming from Marconi's work.

Time has been allocated after every presentation for discussion; it is particularly hoped that those who knew Marconi will be present and will speak of their personal reminiscences.

Further details and application forms for the event may be obtained by writing to the Secretary, IEE, Savoy Place, London WC2R 0BL, quoting reference LS(MA), or by telephoning 01-240 1871, ext. 319.

Bio-Medical Engineering

The Biological Engineering Society is holding its Fourth Annual Conference on 'Recent Advances in Bio-Medical Engineering' at the University of Surrey from 7th–10th April, 1974. This will be a three-day meeting on topics which have been chosen to appeal equally to those whose interests and work involve the use of electrical or mechanical techniques. One of the principal aims of these annual meetings is to review recent achievements, and this year sessions will be on four main themes, namely Telemetry, Orthopaedics, Blood Circulation and the work of a Hospital Department of Medical Engineering.

The sessions on Telemetry will cover *all* applications of radio, ultrasonic and storage telemetry. On the last day papers will be presented about the work of a Medical Engineering Department in the hospital, singling out those achievements which have become accepted as part of the routine clinical work of the hospital.

There will be a Scientific and Commercial Exhibition appropriate to the themes of the Conference, located immediately adjacent to the lecture theatre. Accommodation will be available in the University Halls of Residence. Further details can be obtained from Mr. K. Copeland, Biophysics Department, University College London, Gower Street, London, WC1E 6BT.

Conference on Noise and Vibration Control for Industrialists

This is the fourth of the series of conferences which have been sponsored by the IERE, the Society of Environmental Engineers, the University of Bath, and the University of Wales Institute of Science and Technology and held alternately at Cardiff and Bath on the subject of noise and vibration control. It will be held from Monday to Wednesday, 8th–10th April 1974 at the University of Bath and, like its predecessors, the conference seeks to increase awareness of what can be done to alleviate problems of noise and vibration in an industrial environment.

The Conference is aimed primarily at people working in industry who are not specialists in the field of noise and vibration but who nevertheless become involved from time to time with the task of trying to keep levels of noise and vibration within acceptable limits. With this in mind the first morning will be devoted to basic principles and will provide an introduction to the work covered in later sessions. It is important to assess properly the magnitude and nature of any noise or vibration problem and the whole of the second day will be devoted to measurement and analysis of the measured data. Ample time will be available on this day for visiting an associated exhibition. Consideration will also be given to the amount of noise or vibration reduction that is required. On the last day examples will be given of complete investigations and practical solutions.

The fees are £25·00 for the whole conference, £10·00 for attendance on Tuesday or Wednesday and £12·00 on Monday which includes the Conference Dinner. These charges cover meals and refreshments but not accommodation, which can be made available in single study-bedrooms at the University for the nights of Sunday–Wednesday, 7th–10th April. Applications should be made to The Courses Office, University of Bath Centre for Adult Studies, South Building Annexe, Claverton Down, Bath, BA2 7AY (Telephone: Bath 6941, ext. 622).

Spectrum Utilization in Radio Communication

An international conference on 'Spectrum Utilization in Radio Communication' will be held at the Institution of Electrical Engineers, London, from 11th–14th February 1975.

The aim of the conference is to deal with aspects of radio communication system design that will lead to a more effective use of the radio spectrum. It will be of interest to engineers involved in the research, development, operation and management of radio communication systems for both commercial and military purposes.

The conference is being organized by the Power Division of the Institution of Electrical Engineers in association with the Institution of Electronic and Radio Engineers, which is represented on the Joint Organizing Committee by Mr. L. W. Barclay (Fellow).

The areas to be covered will include:

Spectrum allocation: review of current spectrum utilization; matters relating to division of the spectrum between competing users—fixed point-to-point, mobile satellite and broadcast; real-time channel estimation techniques; and exploitation of the unused (millimetric and sub-millimetric) spectrum for communications.

Communication system design: performance specification; traffic and reliability; choice of modulation method; frequency planning; distortion; and diversity methods.

Modulation and demodulation techniques: optimum s.s.b. and i.s.b. reception; f.m. threshold extension; multilevel systems; multiplexing t.d.m., f.d.m. and spread-spectrum methods; error detection and correction; techniques to reduce selective fading; secure speech transmission; bandwidth compression for speech, facsimile and data; and digital signal processing in the design of modems.

Transmitters and receivers: frequency stability and frequency synthesis; reduction of spurious radiation; adaptive receiver filtering; and low-noise receivers.

Antennas: improved directivity; h.f. phased arrays for transmission and reception; wideband active antennas; and space diversity.

It is hoped that papers of a purely theoretical nature will show a direct relevance to future systems and their implementation.

The Organizing Committee now invites offers of contributions not exceeding 3500 words (i.e. maximum of seven A4 pages including typescript and illustrations) for consideration for inclusion in the conference programme. Those wishing to offer a contribution should submit a synopsis of approximately 250 words to the Conference Department by 17th April 1974. The full typescripts will be required for assessment not later than 23rd September 1974.

Registration forms and further programme details will be available a few months before the conference from the IEE Conference Department, Savoy Place, London WC2R 0BL.

Short Courses in Semiconductor Techniques

The Middlesex Polytechnic at Enfield is holding several short courses in the general field of semiconductor technology.

There will be a one-week course on 'Thick Film Hybrids' from 4th–8th March 1974. Designed mainly for electronics engineers and circuit designers while showing the technology of thick films in detail, the course will emphasize the design and applications of thick film circuits. Practical work will take the form of design and fabrication of typical examples. A two-day course on 'Practical Thick Films' with broadly the same scope will be held on Wednesday and Thursday, 22nd and 23rd May 1974.

'Introduction to Microelectronics' is the subject of a one-week course to be held from 10th–14th June 1974. Approximately half the course hours are to be spent in the Centre's laboratories gaining practical experience by processing of bipolar monolithic and thick film hybrid circuits.

A one-week course on 'M.O.S. Circuits and Systems' is to be held from 6th–10th May 1974. This course is intended to give a thorough grounding in m.o.s. theory, technology and circuits. It will be of interest to engineers, scientists and teachers who are conversant with bipolar devices and who wish to extend their knowledge into the rapidly expanding field of m.o.s. circuits. Lectures will cover device theory, fabrication techniques including modern methods, circuit analysis and design. The facilities of the laboratories will enable delegates to gain first-hand practical experience by processing an m.o.s. circuit from the starting wafer to the mounted and bonded chip.

Further details on all the above courses from: Mrs. D. P. Linnell, Short Course Secretary, Middlesex Polytechnic, Queensway, Enfield, Middlesex, EN3 4SF (Telephone: 01-804 8131).

Natural Resources at Risk

There was a good attendance at the Institution of Electrical Engineers on 8th January to hear Sir Kingsley Dunham, D.Sc., F.R.S., C.Eng., Director of the Institute of Geological Sciences, deliver the 19th Graham Clark Lecture, now a major event in CEI's annual calendar. The subject chosen by Sir Kingsley—'Natural Resources, the Engineer and the Environment'—could hardly have been more topical in view of the current energy and transport crises to which he alluded in departures from his set text.

The theme covered by Sir Kingsley embraced current usage and future reserves of metal resources and energy resources, together with a brief review of transport developments and the effect which these have had on the environment. Sir Kingsley started by posing some questions of a nature which he felt should be asked of engineers. 'The questions that need to be asked of the engineers,' said Sir Kingsley, 'are necessarily practical ones: what are the lower limits of grade for profitable working of metallic ores by the most sophisticated modern techniques? Is it possible to design systems for the more efficient recovery of metals from ores, wastes and in the course of recycling? What improvements may be expected in the techniques of recovering offshore petroleum? Can gas be made economically starting from low-grade solid fuels? Is the disposal of fission wastes a safe process? What are the chances of engineering a fusion reactor? Can solar energy be harnessed? What is the future of the internal combustion engine as a prime-mover in transport?'

Since World War II, Sir Kingsley said, oil had become the dominant fuel the world over and clearly it represented the most immediate problem. 'Present known reserves, according to a recent BBC programme prepared in consultation with oil companies, represent a life of 22 years at current rates of consumption; the Institute of Fuel report quoted 32–36 years at 1971 rates, but only 16–18 years if the exponential rise continues.'

The two recent discoveries in which we in this country were interested, continued Sir Kingsley, were Alaska and the northern North Sea. They had been roughly estimated to contain 1·8 and 1·5% respectively of the world's remaining oil. Both presented very interesting engineering problems; in Alaska the problem of working through thick permafrost and of transporting the oil by pipeline without either doing too much violence to the caribou and other fauna of the Alaskan wilderness or losing the pipe in melted permafrost. In the North Sea it seemed to him that if the engineering problems could be solved, extraction from well-heads situated on the sea bottom would have great advantages over the maintenance of the expensive and uncomfortable platforms which necessarily must be used in the prospecting stage.

'If the trend for oil to replace other forms of energy is allowed to continue,' Sir Kingsley said, 'it is more than likely that sources of petroleum will become exhausted in the lifetimes of our children or theirs. It must be concluded that in the long or even medium term the future of oil is in serious doubt; even in the short term, the Institute of Fuel recommends that in view of the doubtful availability of reasonably priced oil, part of this demand should be reallocated to coal, leaving oil to supply those markets where no alternative fuel is possible.'

Of natural gas Sir Kingsley said: 'This is without doubt the most satisfactory fossil fuel from an environmental point of view, easily transmitted, readily desulphurized if any sulphur is present, burning to innocuous products, mainly water. Unfortunately it suffers from limitations as to resources comparable with those applying to oil. We were assured of only 25 years supply when the first natural gas flowed into British mains six years ago and though some further discoveries will improve this figure, the quantity is strictly limited.'

On nuclear energy Sir Kingsley stated that 'at present in this country we produce nearly 10% of our energy requirements by nuclear means, from nine generating stations. Though this is more, proportionately, than in any other country, the application of nuclear generation has not developed as rapidly as was expected in the mid-1950s, when it seemed to offer an easy solution to Suez and oil politics. To quote Lord Penney from his introduction to a book on Uranium Prospecting published last year, "Nuclear energy in spite of the ups and downs of the last twenty years—and that is a short time even compared with one man's lifetime—is the only certain source of the vast quantities of electric power needed to continue living as we do into the next century and beyond."' Sir Kingsley went on to say that the whole question of safety against human error in reactor operation should be widely investigated and discussed before we were committed to major nuclear programmes. 'Doubts must be felt,' he said, 'about injecting even slightly radioactive wastes into the deeper levels of the crust, a practice now very much under discussion.'

On solar energy Sir Kingsley said: 'Surely the time has come to make a serious bid to develop this source on a large scale, though this country is obviously not the ideal place for such a development.'

Copies of the full lecture, 24 pages, are available from CEI, 2 Little Smith Street, London S.W.1, price 40p.

STUDY ON TELECOMMUNICATIONS AND EDUCATION BY POST OFFICE

The development of education in Britain up to the 21st century—and its impact on telecommunication services—is the subject of a major study commissioned by the Post Office. During the study, which is to cover the period 1983–2003, a team of leading educationists will look into the future to forecast the likely form of education Britain will have in the year 2000.

The study, which will last a year, is being undertaken by the Council for Educational Technology, and is led by Mr. Anthony Howarth. A wide range of educational services will be examined—with emphasis on the use of telecommunications, in particular in relation to closed-circuit television and computers in education. The study will seek to establish patterns of communication in education, matching its likely future needs with the services likely to be available from an advancing telecommunications technology.

The study is one of a series commissioned by the Long Range Studies Division of Post Office Telecommunications. Others look at the future of the economy, leisure, transport, health services, and the urban environment. The overall objective is to establish the long-term demands which changes and developments in these areas are likely to place on telecommunication services, so that the Post Office can plan effectively to meet them.

British Executive Service Overseas

an opportunity for retired members

BESO is a new organization, sponsored by British Industry and the Government, which aims at helping smaller industries, mainly in the private sector, in developing countries by the short-term loan of retired senior British Executives and Engineers.

Most developing countries, struggling to raise the standard of living of their people, are looking to the growth of their internal industries and a decreasing dependence on imported consumer goods to achieve this end. This means initially carrying the processing of their exportable raw materials to further stages than previously, and later introducing the manufacture of consumer goods of increasingly sophisticated types, together with the development of banking and insurance services which go hand in hand with such growth.

At some stage, generally quite early in this development, most of them come up against the problem of the lack of indigenous skills in modern industrial and commercial techniques, and the necessary management expertise to carry forward developments at the rate they desire.

The governments frequently make an attempt to meet this problem by setting up state owned enterprises with 'know-how' and technical assistance purchased from more advanced countries. However, such enterprises are inevitably confined to the production of basic materials such as fertilizers, steel, cement, etc., and leave unsolved the problems of small private industries which do not have the resources to buy foreign advice.

In an endeavour to help to overcome this problem, advanced countries are organizing schemes by which smaller private industries in the developing territories can 'borrow' for short periods the services of experienced senior executives and technicians who are no longer in regular employment, to advise and help them overcome specific difficulties. Such schemes have been in successful operation for a number of years by USA and Canada, both of which send abroad several hundred executives each year.

A similar scheme was initiated in Britain about a year ago. It is sponsored 50% by British industry and 50% by the British government through the Overseas Development Administration, and is managed jointly by the Institute of Directors and the Confederation of British Industry. It is called British Executive Service Overseas (BESO), and as is the case with the American and Canadian schemes, offers the services of managerial, professional and technical executives on short-term assignments to enterprises in the developing countries to assist them in improving the efficiency of their operations and in their expansion plans.

An executive stays in the country for a maximum of six months. He is not paid a fee for his services, but all travelling and living costs and certain other expenses such as kit allowances and insurance premiums are paid. His wife may also accompany him on the same basis.

The charges are divided in such a way that BESO pays travelling expenses of the executive and his wife to and from the territory together with kit and insurance costs, etc., while the requesting company pays the local living and travelling expenses on a scale agreed between BESO and the company. This, it is felt, ensures that any company considering using the service will make absolutely sure that there is a real need for it, and a real job for the executive to do, before incurring the expense.

BESO is run by a senior executive who functions as Director, under a council of distinguished businessmen and senior representatives of ODA and CBI. There is a small—mainly part-time—staff of retired business executives who are well aware of the difficulties of industries in developing countries, and the problems likely to be met by the executives who go to help them.

In each overseas territory in which BESO operates a resident representative is appointed, generally a prominent business personage in the territory, who acts for BESO in a part-time capacity. The function of the resident representative is to be the point of contact between industrialists in the territory and BESO headquarters in London. He will discuss an individual industrialist's problems with him and help him to decide precisely what aid he needs, and will advise BESO accordingly as to the type and experience of the executive who could be of most use in each case.

Over the next two years BESO will operate in Hong Kong, India, Malaysia, Brunei, Singapore, Jamaica, Trinidad, Barbados, Malawi and Zambia, and local representatives have already been appointed in these territories. Later the fields of operations will be extended as experience indicates.

A register of volunteer executives is being compiled, and already over 100 senior executives from a wide field of commerce and industry have joined the register to offer their experience and pool their specialized knowledge on these short-term assignments. From the range of requests already received from overseas, it is clear that a very wide range of specialist knowledge is needed and to be able promptly to answer any request needs a pool of executives with experience in the fields of production in many industries, in transport, in the handling of food, in management and financial control, in insurance and in fact in most fields of industry and commerce.

As the Minister for Overseas Development has said: 'The scheme represents a splendid opportunity for British businessmen to share their knowledge and expertise in a practical way with indigenous business overseas.'

Any retired member who is interested and would like an opportunity to exercise once more his expertise in his particular field and at the same time help a developing country, is urged to get in touch with The Director, British Executive Service Overseas, 10 Belgrave Square, London SW1.

Conference on

'The Electronics Industry and Higher Education'

Organized by THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS with the Association
of the Institution of Electrical Engineers.

Royal Holloway College, Egham, Surrey, 28th to 31st March 1974

PROGRAMME

Thursday, 28th March, 7.30 p.m.

Conference Dinner and Address by Dr. I. Maddock, C.B., O.B.E.
D.Sc., F.Inst.P., F.R.S., C. Eng., F.I.E.E., F.I.E.R.E., *President of
the Institution of Radio and Electronic Engineers and Chief Scientist,
Department of Trade and Industry.*

Friday, 29th March, 9.30 a.m. to 6.30 p.m.

THE PRESENT AND FUTURE OUTPUT OF HIGHER
EDUCATION

'Government Policy for Higher Education'.
Dr. H. W. FRENCH, *Department of Education and Science.*

'Employment and Qualifications in the Electronics Industry'.
PROFESSOR G. D. SIMS, *University of Southampton.*

Job Profiles

'A Survey of Job Profiles'.
F. R. J. LANGRIDGE, *Engineering Employers' Federation.*

'Undergraduates' Attitudes to Types of Employment'.
DR. J. R. TILLMAN, *Post Office Telecommunications Research
Department.*

'The Graduate in Industry'.
J. C. TODD, *EMI Electronics.*

Single Discipline Courses

'The Single Subject Course'.
PROFESSOR W. D. RYAN, *The Queen's University of Belfast.*

Multidiscipline Courses

'Establishing Common Ground in Multidisciplinary Courses'.
DR. P. S. STRINGER, *Lanchester Polytechnic.*

'Interdisciplinary and Multidisciplinary Courses'.
DR. N. G. MEADOWS, *Glasgow College of Technology.*

Sandwich Courses

'Sandwich Degree Courses'.
DR. J. M. IVISON, *University of Technology, Loughborough.*

Modular Courses

'Modular Degree Schemes'.
M. BRANDON-BRAVO, *City of London Polytechnic.*

Crossing the Interface and Industrial Training

'Change for the Graduate—Change from the Educational to the
Industrial Environment'.
DR. F. W. HARRISON, *Philips Industries.*

'Post Graduate Training for Industry at University'.
J. M. SHARMAN, *University of Cambridge.*

'The Work of the EITB in the Electronics Industry'.
A. ASPINALL, *Engineering Industry Training Board.*

Saturday, 30th March, 9.30 a.m. to 5.30 p.m.

Post-Graduate Activities

'An Introduction to Post-graduate Activities'.
PROFESSOR E. D. R. SHEARMAN, *University of Birmingham.*

Post-Experience Courses

'Post-experience Courses and Possible Alternatives'.
PROFESSOR H. SUTCLIFFE, *University of Salford.*

Post-Graduate Degree Courses

'Post-graduate Courses with Industrially Based Projects'.
PROFESSOR W. F. LOVERING, *University of Surrey.*

'SRC Policy and Post-graduate Degree Courses'.
PROFESSOR W. E. J. FARVIS, *University of Edinburgh.*

'Post-graduate Degree Courses'.
PROFESSOR C. W. TURNER, *King's College, University of London.*

Research

'The Role of Universities in Research and Development'.
PROFESSOR E. A. ASH, *University College London.*

'The Electronics Industry and Academic Research'.
PROFESSOR K. HOSELITZ, *Mullard Research Laboratories.*

'The Role of Polytechnics in Research'.
DR. N. G. MEADOWS, *Glasgow College of Technology.*

'SRC Policy on Electronics Research'.
PROFESSOR W. E. J. FARVIS, *University of Edinburgh.*

'Direct Industrial Participation in University Research. The Goth
Within the Gate?'
PROFESSOR W. GOSLING, *University of Bath.*

'The Exploitation of University Inventions'.
P. G. TANNER, *National Research Development Corporation.*

Sunday, 31st March, 9 a.m. to 1 p.m.

Bridging the Interface

Staff Exchange/Experience/Consultancy

'Universities—Nothing But Graduate Factories'.
J. D. E. BEYNON, *University of Southampton.*

Visiting Professors

'Some Views of a Visiting Industrial Professor on the Education of
Electrical Engineers'.
DR. P. L. KIRBY, *Welwyn Electric.*

Formalized University/Industry Collaboration

'Industrial Units and Bureaux'.
J. P. MOORE, *University of Surrey.*

'Industrial Projects in Universities—Advantages and Some Problems'.
A. P. DOREY, *University of Southampton.*

'Some of the Problems in University/Industry Collaboration'.
D. STEWART, *University of Strathclyde.*

'A Faculty Training Unit'.
D. F. DUNSTER, *N.E. London Polytechnic.*

'The Role of the Science Research Council'.
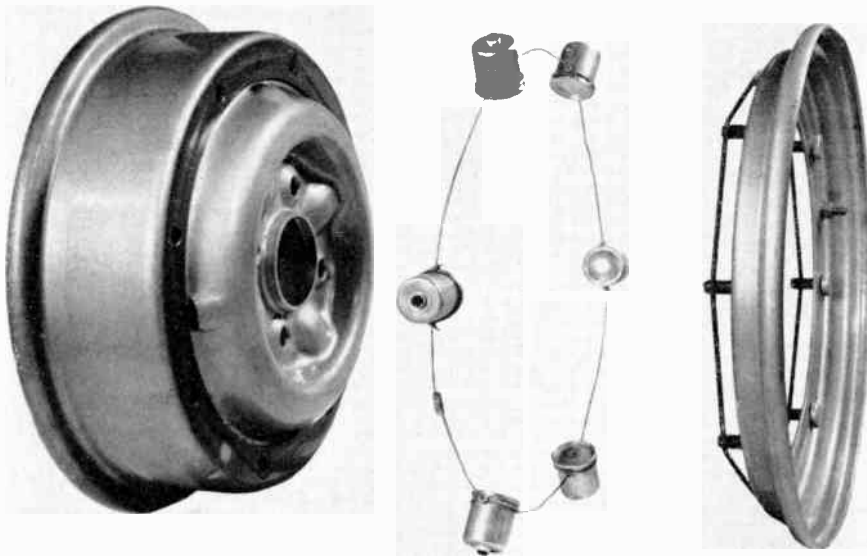A. J. EGGERTON, *Science Research Council.*

Further details and registration forms may be obtained from the
Conference Department, I.E.R.E., 8–9 Bedford Square, London,
WC1B 3RG. (01–637 2771, ext. 20).

MacRobert Award for Road Safety Development

The 1973 MacRobert Award—the major engineering award in Great Britain—has been won by three staff members of Dunlop Limited for their development of the total mobility

physical technologies or any of the physical sciences which has enhanced or will enhance the prestige and prosperity of the United Kingdom.

In his citation for the Award, Lord Hinton of Bankside, Chairman of the MacRobert Award Evaluation Committee, said that the development of Denovo tyre marked a significant improvement in motoring safety and convenience. In respect of safety it would enable a driver to maintain complete control of his vehicle following a puncture or blow out and for the vehicle itself to be driven in safety for distances of up to 100 miles at speeds of up to 50 miles/hour. Its convenience was that there would be no necessity for roadside wheel changing and it eliminated any need to carry a spare tyre.

WHEEL COMPONENTS OF DENOVO

The four components of the divided wheel of the Denovo wheel, which enable the specially-developed tyre to run safely in a deflated state. The two halves of the wheel are bolted together with eight studs and nuts, and an 'O' ring, or seal, is used between the inner surfaces to form an air-tight tubeless tyre and wheel unit. The metal cylinders attached to the wire band (centre) which is fitted over the main part of the wheel, contain the special lubricant necessary to prevent friction when the tyre is deflated after a puncture, and to help seal the hole. The lubricant, which also generates a vapour to reinflate the tyre to a sufficient pressure to allow the tyre to be driven on, is released when the containers are contacted by the inner surface of the flat tyre.

Denovo tyre and wheel system. They are Mr. Tom French, Development Manager, Tyres; Mr. Eric Mitchell, Development Manager, Rim and Wheel; and Mr. Reginald Edwards, a senior tyre designer. The award, made annually by the Council of Engineering Institutions on behalf of the MacRobert Trusts, consists of a Gold Medal and £25,000 prize money. It is presented in recognition of an outstanding contribution by way of innovation in engineering, any other

Last year's MacRobert Award was to Mr. Godfrey Hounsfield and EMI Limited for the EMI Scanner X-Ray technique for diagnosing brain disease (see *Journal* for January/February 1973). Rules and conditions for the Award, for which submissions are to be made to CEI by 30th April for 1974, can be obtained from: The MacRobert Award Office, Council of Engineering Institutions, 2 Little Smith Street, London SW1 3DL. (Telephone: 01–799 3912.)

THE ROYAL NAVY AND THE HISTORY OF RADIO

Two pioneers in radio engineering are commemorated in the names given to two new buildings in H.M.S. *Collingwood*, the Weapon and Electrical Engineering School of the Royal Navy at Fareham in Hampshire, officially opened last October by the Principessa Elettra Giovanelli, daughter of Guglielmo Marconi. The WRNS Block is called 'Electra' which is the name of the yacht from which Marconi conducted many of his later experiments, while the CPO and PO's Mess is called 'Jackson' Block to commemorate Admiral of the Fleet Sir Henry B. Jackson, G.C.B., K.C.V.O., F.R.S., another well-known pioneer in the development of wireless telegraphy and a contemporary and firm friend of Marconi.

A display of early maritime radio illustrating the work carried out by Marconi and Jackson towards the end of last century was arranged for the opening ceremony. Some examples of wireless coherers actually made by Jackson and Marconi in about 1895, together with replicas of Marconi's first beam transmitter and receiver of 1895 were loaned by the Science Museum for the occasion and also on display, on loan from H.M.S. *Mercury*, the Royal Navy Signal School, were replicas of Jackson's spark transmitter and receiver which he used to transmit messages between ships in Devonport harbour. The centenary of Marconi's birth falls this April (see page 1).

Recent BSI Publications

The following publications may be obtained from BSI Sales Department, 101 Pentonville Road, London N1 9ND. (They are not available from the IERE but copies may be consulted in the Library.)

School Television

School television is already well established, but is likely to increase considerably in scope, and the need for technical guidance on the requirements of equipment for use in schools has been felt for some time—in fact there has been nothing substantial in the field since 'School Television Broadcasting' went out of print. At the request of the Association of Education Committees, which produced this 1956 publication, the British Standards Institution has published a major specification on the subject—**BS 4958 School monochrome television receivers and stands.** (*Price £1.30*)

The new Standard is in six sections dealing with electrical signal input and output requirements, controls, construction, safety requirements and reliability. Much of the specification is also applicable to colour receivers, but techniques are not sufficiently developed to enable a complete standard to be prepared for these at present.

In a foreword to the specification it is emphasized that the specification of a school receiver should be as close as possible to that of a domestic model. The rate of growth of school television is inevitably bound up with the cost of equipment, and the nearer this equipment is in specification to domestic installations the cheaper it is likely to be. For this reason, BS 4958 is confined to those technical and safety requirements essential for school use which may differ from the needs of domestic installations.

The main points of deviation are the size of tube (the largest picture tube in volume production is specified), adequate sound, sources of signal, which may be video tape recordings and live pictures, and the need for mobility of the receiver. The Standard also deals with the provision of recording and playback facilities at video and audio frequencies.

Television and Radio Reception

During the last 10 years many changes have taken place in techniques of sound and television broadcasting. These changes, which are closely associated with the increased availability of f.m. v.h.f. sound broadcasting and with greatly increased use of u.h.f. television, are taken into account in **CP 1020 Code of Practice for the reception of sound and television broadcasting.** (*Price £2.00.*) This Code supersedes CP 327.201, published in 1960 and its supplement on u.h.f. reception.

The new Code of Practice sets out recommendations for good reception of sound and television broadcasting transmitted on all the wavebands and standards now in current use. It is particularly applicable to installations in new buildings but many of its recommendations also apply to installations in existing buildings. Guidance on the selection of appropriate receiving equipment to suit particular conditions is given.

Aerials, including communal aerials to serve more than one receiver, and the provision of wired distribution systems are covered, together with structural accommodation for equipment and wiring and protection against various hazards. There are sections on materials, appliances and components, planning, design, installation, inspection, testing and maintenance. Appendices show, in tabular and graphical form, the frequency ranges allotted to the various sound and television broadcast channels including those for wired distribution systems. Typical aerials and installation details are illustrated in diagrams.

Line Traps for Power Carrier Systems

BS 4996 : 1973 Line traps for power line carrier systems (*price £1.30*) which is technically identical with IEC Publication 353, deals with line traps intended to be inserted into high-voltage transmission lines to prevent undue loss of any carrier signal which may be superimposed for purposes of telecommunication, teleprotection or telecontrol. It specifies blocking requirements, temperature rises, ability to withstand short circuit, insulation levels and the tensile strength of the suspension system. Methods of measurement for the inductance of the main coil, temperature rise, blocking impedance and insertion loss are included. Recommended values of rated quantities are given in an appendix.

Standardization in Production

Productivity in a large sector of the British engineering industry could be increased between three-fold and five-fold by standardization, especially if the opportunity of a 'fresh start' in the metric system is seized. This is one of the main conclusions of 'The Management of Design for Economic Production', (*PD 6470, price £2.90*). Its author is Mr. L. Sumner, Deputy Director of Weapons Production (Navy), former Deputy Director Standardization, Ministry of Defence. He was assisted by a Steering Panel representative of industry, research, Government departments and BSI, and a small working party on Relative Costs.

The aim of the guide, says Mr. Sumner, is to make Management and designers aware of the extremely effective role which standardization philosophy can play in ensuring that manufacturers can meet the needs of the market place while at the same time making the most economical use of resources. Change is inevitable; if this is recognized and properly controlled, it is possible to take advantage of change in the continual optimization process. He continues:

'It is often said that standardization is inimical to high performance and progress.. Anyone holding that view has a fundamental misconception. Standardization, in its broadest sense, is the discipline of using the minimum number of parts for the maximum number of purposes, produced by the most economical manufacturing processes, of the appropriate quality to give reliable and acceptable performance at minimum (whole life) cost. It is an essential tool for use when businesses diversify in order to spread the risk over a wider sector of the market, since, if control is not applied, the increased variety of components can undo the beneficial effects.'

The guide discusses in detail the nature of design; design and standardization philosophy—the economical use of resources; production by the most economical manufacturing processes; obtaining the necessary quality; designing for minimum cost; and a check list for action by company management. Cost comparison charts cover material prices, machining costs, and the costs of accuracy, complexity and of producing various degrees of surface texture, etc. These enable the designer to appreciate production costs so that he can design to achieve a high degree of productivity, low complexity and the lowest possible production costs.

The author concludes: 'There are lessons to be learnt in industry from the SYMES (Systematic Machinery and Equipment Selection) approach. If the Royal Navy can power 14 different classes of ships with three prime movers, could not the variety of some industrial products be reduced while still meeting requirements? If this approach could also be considered at the time of change to metric designs, further economic and logistic advantages would also ensue.'

Standard Symbols for Data Processing

The new British Standard **BS 4058 : 1973 Data processing program and data flow chart symbols rules and conventions** (*price £1.30*) includes two flow chart symbols additional to the 30 symbols which were listed in BS 4058 : 1966. These two symbols are 'off page connector' and 'control document'. These symbols have been added because experience of the application of BS 4059 : 1966 has indicated the need. The other 30 symbols are exactly the same as those listed in the 1966 edition and they are in conformity with ISO/R 1028 the relevant Recommendation of the International Organization for Standardization. The Standard also lists, and illustrates

by way of examples, rules and conventions for the preparation and construction of program and data flow charts using these symbols. The rules and conventions are based on those given in ISO/DIS 2636 but extended and clarified where considered necessary.

Electronic Measuring Equipment

BS 4889 : 1973 Method for specifying the performance of electronic measuring equipment (*price £1.30*) describes a standard method of specifying the performance of electronic measuring equipment for use by the manufacturer of such equipment. Limits of operating error, which must not be exceeded under any combination of environmental conditions within the specified operating conditions, are required. Three sets of environmental conditions

are provided to suit equipment designed for use under different circumstances.

Reliability of Electronic Equipment

BS 4200 Guide on the reliability of electronic equipment and parts used therein: Part 7 : 1973. The inclusion of reliability clauses in specifications for components (or parts) for electronic equipment (*price 90p*) is intended to guide rather than limit writers of specifications and technical committees in considering factors to be taken into account in preparing reliability clauses in specifications for components (or parts) for electronic equipment. The standard is identical in technical content to the relevant Publication of the International Electrotechnical Commission except where indicated.

Power from the Telephone Line—Requirements for Attachments

As from 1st July, 1974, certain attachments will be permitted to draw their activating power from Post Office exchange lines or PBX extensions. The requirements and conditions applied to such devices in addition to existing requirements for self-powered attachments to the public switched telephone network are set out in a recent Post Office technical specification.

1. The Post Office will not undertake to provide lines to a particular standard. Line-powered attachments must function with conditions normal to the telephone installation and be independent of line polarity.

2. Suppliers and users must accept that attachments may be rendered unusable by Post Office system changes. A minimum of 2 years' notice of such changes is normally given by the Post Office. To safeguard the interests of users in the event of system changes, line-powered attachments must be supplied on rental terms.

3. Telephone exchange battery voltage limits are 46–52 V and may contain ripple components not exceeding 2 mV. PBX line voltage limits are normally 45–55 V but some PBXs have lower nominal voltages and limits. Line current may be any value between 25 and 110 mA depending upon the combination of feeding, line and terminal resistances. It should be noted that additional power components exist on lines which might affect line powered attachments. These are:

    (a) Longitudinal induced voltages up to 20 V r.m.s. at 50 Hz.

    (b) Uniform spectrum and random noise having a power not exceeding −45 dBm over the band 300–3400 Hz.

    (c) Random impulsive noise in excess of −22 dBm.

4. Power may be drawn from line only when the line is diverted to the attachment by means of a telephone push-button switch which is restored by replacing the receiver. Power may be drawn from line only during the chargeable duration of a call. The line-terminating condition consists of the following conditions presented to line by the attachment:

    (a) A direct-current circuit with resistance not exceeding 300 Ω.

    (b) An alternating-current circuit with impedance in the range of 400–900 Ω with angle not greater than 45° in the frequency range 300–3400 Hz.

    The circuit must be capable of carrying d.c. up to 120 mA without going outside the limits in 4(a) and (b) above.

5. Motors or other components liable to introduce electrical interference into the telephone line may not use line power.

6. Attachments connected to line through Post Office modems may not use line power.

7. The decision to permit line powering is most likely to affect multi-frequency data devices but is not restricted to such attachments.

8. The terms of Technical Guide No. 26 (Protection) apply if the attachment is powered partly by a private power supply or is connected to another device which is powered by a private power supply.

9. Designers and suppliers are advised to consider the need to protect the attachment from voltages which may be impressed on Post Office lines by lightning and other extraneous sources. These voltages may be unidirectional or alternating and of a transient nature. Transient and prolonged voltages are also present on many Post Office circuits as normal functional characteristics. They may reach a peak value of several hundred volts. Prolonged voltages of at least 100 V peak may be encountered (e.g. from ringing current). Circuits containing semiconductors are particularly vulnerable to damage by comparatively small excess voltages even of short duration. The Post Office will not accept any responsibility for damage caused to attachments by any of the above conditions.

10. The Post Office reserves the right to withdraw the line powering concession at reasonable notice. All applications for consideration by the Post Office should be submitted to Post Office THQ, Service Department (Sv 1.1.3), 45 Moorfields, London EC2Y 9TH.

STEREOPHONIC BROADCASTING

Test Tone Transmissions—BBC Radio 3 V.H.F. Stereophonic Transmitters

(Communication from the BBC Engineering Information Department)

After the close of BBC Radio 3 programmes each Monday and Saturday evening, the Radio 3 v.h.f. stereophonic transmitters carry test tone transmissions to assist in the setting-up and checking of stereophonic receivers. The table and notes below describe the tests and their purposes.

It is emphasized that the use of these tests for the adjustment of decoders requires test instruments, technical knowledge, and experience. Listeners lacking any of these are advised to ignore all tests after the first.

Test No.	Time* (min)	Left Channel (A)	Right Channel (B)	Purpose
1	T	250 Hz at zero level	440 Hz at zero level	Identification of left and right channels and setting of reference level
2	$T + 2'$	900 Hz at $+7$ dB	900 Hz at $+7$ dB, antiphase to left channel	Adjustment of phase of regenerated subcarrier (see Note 4) and check of distortion with signal wholly in the $(A - B)$, i.e. S, channel
3	$T + 6'$	900 Hz at $+7$ dB	900 Hz at $+7$ dB, in phase with left channel	Check of distortion with signal wholly in the $(A + B)$, i.e. M, channel
4	$T + 7'$	900 Hz at $+7$ dB	No modulation	Check of A to B cross-talk
5	$T + 8'$	No modulation	900 Hz at $+7$ dB	Check of B to A cross-talk
6	$T + 9'20''$	Tone sequence at -4 dB: 60 Hz 900 Hz 5 kHz 10 kHz This sequence is repeated	No modulation	Check of A-channel frequency response and A to B cross-talk at high and low frequencies
7	$T + 10'20''$	No modulation	Tone sequences as for left channel on Test 6	Check of B-channel frequency response and B to A cross-talk at high and low frequencies
8	$T + 11'20''$ $T + 13'$	No modulation Reversion to monophonic transmission	No modulation	Check of noise level in the presence of pilot tone

* T is approximately 4 minutes after the end of Radio 3 programmes (see *Radio Times*)

NOTES

1. The above schedule is subject to variation or cancellation to accord with programme requirements and essential transmission tests.

2. The zero level reference corresponds to 40% of the maximum level of modulation applied to either stereophonic channel before pre-emphasis. All tests are transmitted with pre-emphasis.

3. Periods of tone lasting several minutes are interrupted momentarily at one-minute intervals.

4. With receivers having separate controls of subcarrier phase and cross-talk, the correct order of alignment is to adjust first the subcarrier phase to produce maximum output from either the A or the B channel during Test 2 and then to adjust the cross-talk (or 'separation') control on Tests 4 and 5 for minimum cross-talk between channels.

5. With receivers in which the only control of cross-talk is by adjustment of subcarrier phase, this adjustment should be made on Tests 4 and 5.

6. Adjustment of the 'balance' control to produce equal loudness from the A and B loudspeakers is best carried out when listening to the announcements during a stereophonic transmission, which are always made from a centre-stage position. If this adjustment is attempted during the tone transmissions, the results may be confused because of the occurrence of standing-wave patterns in the listening room.

7. The outputs of most receivers include significant levels of the 19 kHz tone and its harmonics. These components do not interfere with normal listening but do affect most signal-level meters. It is essential, therefore, to provide filters with adequate loss at these frequencies if instruments are to be used for the above tests.

8. On Sundays, Tuesdays, Wednesdays, Thursdays and Fridays, to facilitate channel identification and adjustment of channel cross-talk, 250 Hz tone is transmitted in the left channel only from about four minutes after the end of Radio 3 for twenty minutes. This test may be interrupted from time to time.

Letters to the Editor

From: **J. R. Edwards**, B.Sc.(Eng.), A.C.G.I., C.Eng., M.I.E.E.

and **Arthur Towning** (Graduate)

A Comparison of Modulation Schemes for Binary Data Transmission[1]

Mr. D. A. H. Brown of the Royal Radar Establishment, Malvern, has kindly pointed out that equation (5) in Table 1 which gives the error probability for incoherent detection of on/off (unipolar) amplitude modulation has a term $(U_D/2\sqrt{E_B})^n$ missing, and the equation should read:

$$P_B = \tfrac{1}{2} \exp\left(-\frac{U_D{}^2}{2N_0}\right) \times$$
$$\left[1 + \exp\left(-\frac{2E_B}{N_0}\right) \sum_{n=1}^{\infty} \left(\frac{U_D}{2\sqrt{E_B}}\right)^n \cdot I_n\left(\frac{2U_D\sqrt{E_B}}{N_0}\right)\right]$$

Equation 17 of Reiger's paper (ref. 3 of my paper) also has this term missing, although his equation 16 is correct. Equation 17 may be derived from equation 16 by successive integration using[2] $\int_0^z t^v I_{v-1}(t)\,dt = z^v I_v(z)$. With suitable changes of variable, these expressions can be equated with those in Rice's classic paper[3] and in Bennett and Davey[4].

J. R. EDWARDS

Electrical Engineering Branch,
Royal Military College of Science,
Shrivenham, Swindon,
Wiltshire SN6 8LA.

24th January 1974

1. *The Radio and Electronic Engineer*, **43**, No. 9, pp. 562-8, September 1973.
2. Abramowitz, M., and Stegun, A., 'Handbook of Mathematical Functions', Equation 11.3.25 (Dover Publications, New York, 1965).
3. Rice, S. O., 'A mathematical theory of noise', *Bell Syst. Tech. J.*, **24**, p. 46ff, 1945 (Section 3.10).
4. Bennett, W. R., and Davey, J. R., 'Data Transmission', p. 155 (McGraw-Hill, New York, 1965).

Electronic Tuning of Musical Instruments

With reference to Mr. L. H. Bedford's article on Tuning Pianos and subsequent correspondence,* I am inclined to agree with him that analogue indication of tuning is superior to digital. This is perhaps because tuning a piano string for zero beat, like tuning the tank circuits of a transmitter for peak output power, is essentially an 'analogue' operation. With the transmitter, the engineer 'perceives' a relationship between tuning movement (including backlash, etc.) and meter needle movement (or oscilloscope amplitude)—a relationship which the engineer gets to know by 'feel', and once learnt, enables him to tune quite rapidly.

An analogous relationship exists between the *moving* wrest pin and aural perception, and this is essential because we need to know exactly when to bring the rotating pin to a halt in the correct position, to avoid, as much as possible, the new hardship encountered from the initial friction of every stationary pin not correctly tuned. Certainly the engineers of my acquaintance who have practised transmitter tuning with digital indication have given an unanimous verdict of 'hopeless'.

With regard to the provision of an equal temperament scale by frequency division, at least two commercial integrated circuits have appeared for this purpose (the Marconi-Elliott MATO and the Motorola dividing system whose worst error is 0·55 cents). Moreover, commercial tuners for laying bearings employing a stroboscope have been available for at least 20 years.

What, however, disappointed me about Mr. Bedford's article is that it does little about the overwhelming problem which is *not* how to tune a piano given a stable 12-note equal temperament (and I speak from 20 years' experience of piano tuning) but how to provide a cheap 12-note electronic oscillator tuned to equal temperament in the first place!

All of Mr. Bedford's schemes (except the Mk. IV and even that has a —1·9 cents change) *assume* that this 12-note oscillator is available when, in fact, it must form the most difficult and expensive part of any scheme. If Mr. Bedford will publish details of a 12-note RC transistor oscillator stable in terms of relative pitch and equally tempered then he will render a great service to piano tuners. It should be noted however that the majority of commercial designs (e.g. in electronic organs) prefer the frequency stability of the Hartley or Colpitts configurations.

My own method of piano tuning (without oscillators and using one 440 Hz tuning fork) results in a maximum error of 3 cents in any one note in the octave: C4 to B4—which represents (according to Motorola and others) a fairly accurate musical scale. It is certainly easy on the ears in all keys, and is likely to represent the limit of accuracy that is reached by piano tuners in practice.

ARTHUR TOWNING

9 Middlesex Drive,
Bletchley MK3 7JE
6th December 1973

* 'Electronic aids for the tuning of musical instruments', *The Radio and Electronic Engineer*, **42**, No. 9, pp. S.131-5, September 1972. Letters, **43**, No. 1/2, pp. 169-120, January/February 1973.

Members' Appointments

CORPORATE MEMBERS

Mr. M. J. Crawt (Fellow 1966, Member 1963) who has been Chief Engineer, Transmitters, with the BBC since 1968 is retiring in April. A member of the Corporation since 1940, Mr. Crawt has lately been particularly concerned with expansion of the u.h.f. television transmitting network, notably for setting up automatic unattended stations.

Mr. H. E. Drew, C.B. (Fellow 1948, Member 1944) has had conferred upon him the Insignia Award in Technology *honoris causa* by the City and Guilds of London Institute. The citation for this honorary award, two of which are usually made each year, recognized Mr. Drew's practical skill, knowledge and attainment in Radio and Electronic Engineering. Before his retirement from the Civil Service two years ago Mr. Drew was Executive Officer of the Defence Quality Assurance Board and he is at present an Industrial Consultant. He was for many years chairman and member of the Institution's Membership Committee and a member of the Finance Committee and he served on the Council for several years.

Professor William Gosling (Fellow 1968) has now taken up the chair of Electronic Engineering at the University of Bath where he will be extending research facilities in radio engineering. Professor Gosling was at the University College of Swansea from 1958 to the end of 1973, since 1960 as Professor of Electrical Engineering. He is Chairman of the Institution's Professional Activities Committee and was re-elected to the Council for a second year of office as a Vice President last October. He is author of numerous papers in the Journal and other publications and of several books.

Mr. C. R. Longman (Fellow 1968, Member 1955, Graduate 1954) has been appointed Chief Engineer, Television, with the BBC. He was previously head of Studio Capital Projects Department.

Mr. Norman Wheatley (Fellow 1973, Member 1970, Associate 1961) who is with Cable & Wireless Ltd., has taken up the post of Manager, Overseas Services, in Hong Kong. He was temporarily on secondment to Jamaica International Telecommunications Ltd. as Chief Engineer from April to September last year, and before that he was Regional Engineer in the Arabian Gulf. In 1970 Mr. Wheatley contributed a paper to the Journal discussing the planning and commissioning of the communications satellite Earth station at Bahrain.

Mr. L. J. Brogan (Member 1970, Graduate 1964) who was previously a Senior Engineer at Ferranti Limited at Wythenshawe, has joined Loyds Retailers Limited of Manchester as Materials Manager.

Mr. C. A. Casling (Member 1973, Graduate 1969) is now a Field Sales Engineer with Newmarket Transistors Limited. He was previously Area Account Manager with Transitron Electronics Limited.

Mr. M. J. Chaney (Member 1971) has joined Marconi Communications Systems Limited as an Installations Engineer. He was previously with Sound Reproducing Equipments (Bristol) Limited.

Mr. P. Denby (Member 1958, Graduate 1952) who has been with the Designs Department of the BBC since 1954, latterly as Head of the Sound Apparatus Section, has been appointed Head of Studio Group with responsibility for all BBC-designed equipment in radio and television studio installations.

Mr. R. V. Gates (Member 1970, Graduate 1960), formerly a Process Control Engineer with Wiggins Teape R & D Ltd., has moved to RHM Research Limited as Instrument and Control Engineer.

Mr. M. H. Green, who was appointed Quality Manager at Marconi Space & Defence Systems Limited at Dunfermline in 1972, has moved to take up a similar position at the Company's Applied Electronics Laboratories in Portsmouth.

Mr. R. E. Haggitt, B.Sc.(Eng.) (Member 1973, Graduate 1967) has joined Sperry Rand Limited as Systems Engineer. He had been with the British Aircraft Corporation at Filton, latter as Group Leader concerned with design of flying controls systems.

Mr. R. N. Jones (Member 1970, Graduate 1963) who has been with Varian Techtron in Australia since 1971, is now Senior Electronic Engineer, Engineering Services, at Mulgrave in Victoria; he was previously Senior Project Engineer (UV-Visible Products).

Mr. J. R. Kirkus (Member 1970, Graduate 1959) has been seconded from his post of lecturer with the BBC Engineering Training Department to the International Telecommunication Union as a Training Expert on Broadcasting Engineering with the United Nations Development Project in Malaysia.

Mr. R. S. Lester (Member 1968, Graduate 1961) has been promoted to Manager of the Autotest Section at Marconi Instruments Limited. He has been with the Company since 1968 and for the past four years has been Sales Manager for the division.

Mr. T. L. Martin (Member 1971) has returned to the United Kingdom as an Executive Engineer with the Scottish Telecommunications Board of the Post Office Corporation. During the past two years he has been on secondment to the Hong Kong Telephone Company.

Inst. Lt. Cdr. P. Nightingale, R.N. (Member 1972, Graduate 1967) has taken up an appointment of Programmed Learning Instructor at the Army School of Instructional Technology, Beaconsfield. For the past two years he was Radar Group Officer at the Radio Electrical School, H.M.S. *Collingwood*.

Mr. R. O. Prout (Member 1963, Associate 1962) has been appointed lecturer in the Electrical Engineering Department of the City of Gloucester College of Technology. He was previously a marketing engineer with Emihus Microcomponents Ltd. of Weybridge, and Waycom Ltd., of Bracknell.

Mr. J. Rabson (Member 1970, Graduate 1966) has joined the Post Office Advisory Group on Systems Definition as an Executive Engineer. He was previously a Research Officer in the Department of Electrical Engineering Science, at the University of Essex.

Mr. J. M. F. Spalding (Member 1973, Graduate 1963) has been appointed Chief Engineer, Zambia Broadcasting Services, on secondment from the British Broadcasting Corporation. Previous secondments abroad have been as Radio Network Project Engineer in Laos, S.E. Asia, and Television Project Engineer, Tripoli, for the Libyan Broadcasting Service. While in the United Kingdom Mr. Spalding was with the Studio Capital Projects Department from 1969 to 1973.

Lt. Col. B. Southwell, REME (Member 1967, Graduate 1960) is now a Staff Officer I (Guided Weapons and Radar) at the Headquarters DEME (Army).

NON-CORPORATE MEMBERS

Mr. John Savage (Companion 1972), formerly Managing Director of Collins Radio Company of England Limited, has joined Plessey Avionics and Communications as Commercial Director. Mr. Savage serves on the Committees of the Aerospace Maritime and Military Systems Group and of the Communications Group.

Mr. K. C. Ball, B.A. (Graduate 1972) has been appointed to the Government Communications Headquarters, Cheltenham as Professional and Technology Grade II.

Mr. J. S. Basi, B.Sc. (Graduate 1973) who was previously with the Post Office, has now joined Digital Equipment Co. Ltd., Reading, as a Lecturer on Computers.

Mr. J. Burns (Graduate 1972) is now an Assistant Engineer with W. & T. Avery Ltd., Smethwick.

Mr. D. J. Gough (Graduate 1969), who is a Professional and Technology Officer, Grade II, at the Ministry of Defence, has joined the Dockyard Production Control Study Team at Portsmouth; for the past three years he has been a Quality Assurance Officer with the Module Repair and Calibration Facility at the Clyde Submarine Base, Faslane.

Mr. M. E. Harper, B.A. (Graduate 1972) who was with Computer Technology Ltd., has joined Kienzle Apparate Gmbh, Schwarzwald, West Germany, as Development Engineer.

Mr. A. M. Jordan (Graduate 1970) is now a Senior Design Engineer with Fab-Ritek Computer Components, Blantyre. He was previously an Electronics Engineer with Honeywell Ltd.

Mr. P. R. O'Neil (Graduate 1970) has been promoted to Engineer in the Studio Capital Projects Department of the BBC.

Mr. A. H. Shaikh, B.Sc. (Graduate 1970) who was a Test Engineer with the Decca Navigator Co. Ltd. from 1970 to 1973, has joined Data 100 Systems Limited as Systems Engineer.

Mr. Gurcharan Singh, B.Sc.(Eng.) (Graduate 1968) has been appointed Senior Engineer (Commissioning) TXE4 with Standard Telephones and Cables Limited, New Southgate.

Mr. A. S. Wake, B.Sc. (Graduate 1969) who was formerly an Electronic Design Engineer with George Kent Ltd., has joined Marconi Space and Defence Systems Ltd., Portsmouth, as a Senior Engineer.

Mr. M. R. Weaver (Graduate 1969) is now Area Sales and Service Manager, for South East England with Instrumentation Laboratory (U.K.) Ltd., Altrincham; he was previously with Standard Telephones and Cables.

Mr. C. J. Wormald, B.Sc. (Graduate 1968) who joined the Marconi Co. Ltd. in 1968, has moved to Ferranti Limited, Wythenshawe as Systems Engineer.

Mr. A. W. Eves (Associate 1972) is a Technical Instructor Grade II at South Brisbane Technical College. Before emigrating to Australia in 1973, he was a Lecturer in Electronics and Electrical Engineering at the North West Kent College of Technology, Dartford.

Mr. M. J. Potter (Associate 1973) has joined IBM U.K. Ltd., as an instructor, following his retirement from the RAF.

Mr. C. Willerton (Associate 1973) who joined Cable & Wireless Ltd. in 1970 following service in the RAF as a Chief Technician, has recently taken up a position with the Electricity Supply Commission in South Africa.

Obituary

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

Peter Ronald Harrington (Graduate 1963) died on 12th September 1973 at the age of 38. He leaves a widow.

Following National Service as a Radio Mechanic in the Royal Signals, Mr. Harrington joined the Guided Weapons Division of the Plessey Company in 1955. During the next five years he studied for HNC qualifications at the South West Essex Technical School, Walthamstow. He transferred to Plessey Marine Systems

Division in 1960 as an Assistant Engineer and completed his studies for Graduate Membership at North West Essex Technical College in 1963. His career continued with Plessey Marine, during which time he was promoted to Engineer and Senior Engineer and at the time of his death he was engaged in defence contracts engineering duties.
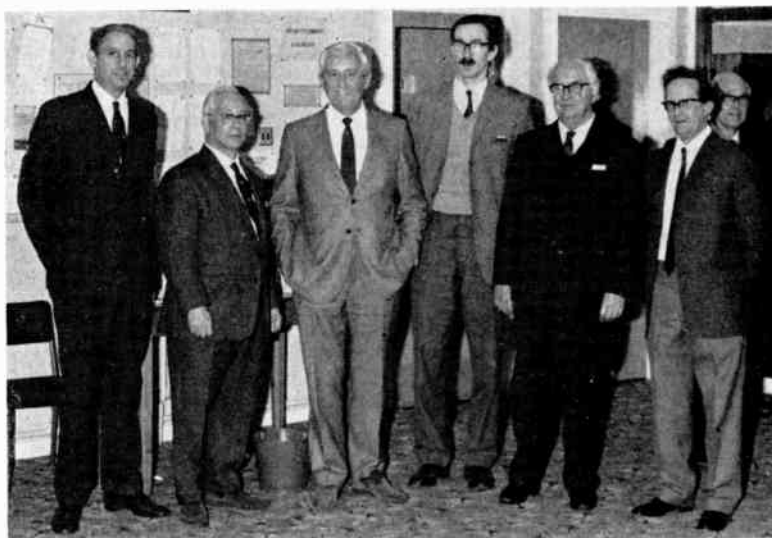
Gilbert Frank Herbert Blackwell (Graduate 1969) died October last, aged 58 years. He leaves a widow.

Mr. Blackwell obtained his technical qualifications by part-time study at Nottingham and District Technical College and at Derby and District College of Technology while working at British Thomson Houston Limited, Rugby, and with REME Technical Services, Chilwell as a Technician Engineer. In 1960 he moved to the Weapons Branch of the Technical Group REME, Woolwich where he supervised a laboratory for investigating repair limits and tolerances of electronic equipment; latterly he was particularly concerned with medical electronic equipment.

West Midland Section meets at RAF Cosford

Left to right:
Sqn. Ldr. T. R. Bartlett (Member) (RAF Cosford), Wing Cdr. D. R. McCall (Member) (RAF Cosford and a member of the Section Committee), Chief Inspector W. A. Hambrey, Mr. R. H. Whitlock, Mr. E. L. Gardiner, and Mr. B. R. Evans (members of the Section Committee).

For the second year running the West Midland Section commenced its autumn programme of meetings at RAF Cosford, by kind permission of the Commandant, Group Captain C. L. Parkinson, M.A., RAF. The 1972 lecture was on 'European Communications Satellites' and this year's was given by Chief Inspector W. A. Hambrey, of the Midland Links Motorway Police Group, on 'The Electronic Control and Communications Network Employed on the Midland Links Motorways'. It is astonishing to note the breadth of technology—including electronics—which is incorporated in the control of a modern motorway and an expressway such as at Aston. The lecture was well received and it is hoped that Cosford will be a recognized venue for the Section for many years to come. D.R.McC.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 28th December 1973 and 15th January 1974 recommended to the Council the election and transfer of 108 candidates to Corporate Membership of the Institution and the election and transfer of 11 candidates to Graduateship and Associateship.¹ In accordance with Bye-law 23, 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: 28th December 1973 (Membership Approval List No. 173)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

ABRAHAMS, Peter John. *New Ash Green, Kent.*
ADSETT, David Edward Murray. *New Ash Green, Kent.*
ARCHER, Lionel Graham, Flight Lieutenant. *Prestwood, Buckinghamshire.*
AINGE, Derek Richard, Flight Lieutenant. *Henlow, Bedfordshire.*
ATKINSON, Dennis. *Frodsham, Cheshire.*
BALDWIN, Neil Frank. *Milton Keynes, Buckinghamshire.*
BANDYOPADHYAY, Sanatkumar, B.Sc. *Croydon, Surrey.*
BARTHOLOMEW, John Francis. *Dorking, Surrey.*
BLUNSDON, Peter Charles. *Macclesfield, Cheshire.*
BOLTON, Edward Ronald. *Carshalton, Surrey.*
BRATTON, Roger Hugo. *Poynton, Cheshire.*
DICKENSON, Ronald. *Otley, Yorkshire.*
HUTCHESON, Thomas Campbell. *Larne, County Antrim, Northern Ireland.*
IVETT, Peter Robert. *Cowes, Isle of Wight.*
JAMISON, David Dempster. *Lisburn, County Antrim, Northern Ireland.*
JONES, Maurice Terrance. *Telford, Shropshire.*
LEDAMUN, David Allen. *Cheltenham, Gloucestershire.*
LINDER, William Joseph. *Barnton, Edinburgh.*
LORD, Anthony Hardman. *London, N.14.*
MAIN, William. *Welwyn Garden City, Hertfordshire.*
MASON, David Ralph. *Portsmouth, Hampshire.*
MARTIN, Geoffrey Ronald. *Poole, Dorset.*
MERRITT, David John. *Hornchurch, Essex.*
MITCHELL, Peter Stewart. *Hazel Grove, Cheshire.*
MORRIS, David Barry. *Pontardulais, Swansea.*
NEESON, Dominic Francis. *Dunmurry, County Antrim, Northern Ireland.*
NEIGHBOUR, David Charles. *Hinckley, Leicestershire.*
NEWTON, Arthur Edward. *Solihull, Warwickshire.*
NIBLOCK, John Andrew. *Northenden, Manchester.*
OATEY, Alan Henry. *Peterborough, Northants.*
OWEN, Maurice Thomas. *Clevedon, Somerset.*
PARKER, Arthur William. *Chippenham, Wiltshire.*
PARKER, Ronald. *London, W.C.1.*
PATERSON, George McAndrew. *Edinburgh.*
PAYNE, Alan James. *Biggleswade, Bedfordshire.*
PAYNE, Geoffrey Michael. *Rainham, Kent.*
PEARSON, Eric Alexander, Inst. Lieutenant, R.N. *Fareham, Hampshire.*
PEREIRA, Raphael Frank. *London, N.W.2.*

PHYALL, David Robert. *Gosport, Hampshire.*
RABONE, Norman Stanley, M.Sc. *Strood, Kent.*
RIVERS, Colin Walter. *Plymouth, Devon.*
ROBERTS, Colin Humphrey Faraday. *Hook, Hampshire.*
ROBINSON, Kenneth William. *Luton, Bedfordshire.*
RUBERY, Ian Leslie. *Ilkeston, Derbyshire.*
SHERWOOD, Kenneth Trevor. *Gt. Yarmouth, Norfolk.*
STEPHENSON, John Raymond. *Farnborough, Hampshire.*
STOCK, Alan Henry Herbert. *Brixham, Devon.*
STREETLEY, Anthony William. *Chessington, Surrey.*
SWABE, Roger Alan. *Harrow, Middlesex.*
TAYLOR, David William. *Worcester Park, Surrey.*
TAYLOR, Frederick Lorraine. *Stourport-on-Severn, Worcestershire.*
TAYLOR, Hugh Taylor. *Reading, Berkshire.*
TICKNER, Michael Edward. *Crowborough, Sussex.*
TULL, Geoffrey Reginald, Lieutenant, R.N. *Stubbington, Hampshire.*
UNDERDOWN, David William. *Woking, Surrey.*
VAUFROUARD, John Charles. *Billericay, Essex.*
VENTERS, John. *Kempsey, Worcester.*
WARREN, Michael John. *Sheffield, Yorkshire.*
WATTS, Brian Lawrence. *Duston, Northampton.*
WHEELER, Edwin Ernest. *Dorking, Surrey.*
WHITE, Thomas David. *Ditton, Kent.*
WILLERTON, James. *Woking, Surrey.*
WILLIAMS, Peter Albert. *Bolton, Lancashire.*
WILLMOTT, Peter Geoffrey. *Rochester, Kent.*
WILSON, Christopher John. *Stanmore, Middlesex.*
WILSON, Malcolm Albert. *Scarborough, Yorkshire.*
WILSON, Terrence Heath Walker, Squadron Leader. *Bushey Heath, Hertfordshire.*
WITHALL, Kenneth John. *Portsmouth, Hampshire.*
WRAY, Patrick Leon. *Westbury-on-Trym, Bristol.*
WRIGLEY, Peter Michael. *Alresford, Hampshire.*
ZIELINSKI, Kazimierz. *London, S.W.19.*

Direct Election to Member

BARKER, Cyril Glyn. *Basingstoke, Hampshire.*
BRADFIELD, Roy Oliver. *Cambridge.*
CORDON, Gerald, Major, R. Sigs. *Haywards Heath, Sussex.*
GERAGHTY, Francis. *Goldington, Bedford.*
GORDON, Baruch, B.Sc. (Eng.). *London, N.W.10.*
GREEN, John Edwin. *Chippenham, Wiltshire.*
GUILDFORD, Leslie Henry. *Haywards Heath, Sussex.*
HOPE, David Henry James. *Dereham, Norfolk.*
HUGHES, Hugh David, B.Sc. *Camberley, Surrey.*
LEE, Eric William James. *Sherborne, Dorset.*

McNALLY, Lionel Charles, Flight Lieutenant. *Henlow, Bedfordshire.*
NEWMAN, Haydn Augustus. *Stourbridge, Worcestershire.*
RAMSHAW, George Dobson, Squadron Leader, B.Sc. *Weston-super-Mare, Somerset.*
RICHARDS, Peter Harold Charles. *Bognor Regis, Sussex.*
SMITH, Malcolm. *Sheffield, Yorkshire.*
THOMPSON, Dennis. *London, W.5.*
WADHAM, David Robert, B.A. (Cantab). *London, N.3.*
WALTON, James William. *Chipping Campden, Gloucestershire.*
WHITTAKER, Alan William, Squadron Leader. *R.A.F. Benson, Oxon.*

NON CORPORATE MEMBERS

Transfer from Student to Graduate

FARROW, James Malcolm. *London, S.W.19.*
SCOTT, Roger Anthony. *Bury St. Edmunds, Suffolk.*

Direct Election to Graduate

BRANNON, Raymond James. *Leigh on Sea, Essex.*
DOSI, Dharamdass, B.Sc. *West Drayton, Middlesex.*
GARTH, Jack Edward. *Ipswich, Suffolk.*
GEORGE, Henry. *Haxby, York, Yorkshire.*
KNIGHT, Derek James. *Ruislip, Middlesex.*
LEWIS, Haydn Arthur. *St. Albans, Hertfordshire.*
STOCKHILL, Paul. *Leeds, Yorkshire.*
THILLAIRASAH, Thampu, B.Sc. *Richmond, Surrey.*
WILKINSON, Christopher John, B.Sc. *Tewkesbury, Gloucestershire.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

KULASINGAM, Kandiah, B.Sc. *Puttalam, Ceylon.*
NIRODI, Suresh Bhavanishanker. *Poona I, India.*
NIXON, George. *San Francisco, California, U.S.A.*
OGUNLEYE, Oladele Olayiwole. *Lagos, Nigeria.*
OKONGWU, Josiah Onyenagolum. *Lagos, Nigeria.*
OKUWA, Oladele. *Lagos, Nigeria.*
OXLEY, Kenneth Arthur. *Karrinyup, Western Australia.*
ROBSON, Andrew. *Gundernhausen, Germany.*
SCOTT, Henry George. *Brussels, Belgium.*
SETHI, Jaginder S. Paul, B.Sc. *Glinde, West Germany.*
SOOBARAH, Jagadish. *Quatre Barnes, Mauritius.*
SOUNDARARAJAN, Krishnaiyer. *Bangalore, India.*
TIMSON, John Arthur. *Tehran, Iran.*
WEST, Frederick Harry. *Gebenstorf, Switzerland.*

Transfer from Associate to Member

MORAN, Lawrence Wilkinson, M.B.E. *Boksburg, Transvaal, South Africa.*

Direct Election to Member

CHOO, Teck Seng, B.Sc. *Singapore 12.*
JOSEPH, Menacherry Paul, Commander. *Vasco da Gama, Goa.*
MURPHY, Peter Basil, Flight Lieutenant. *R.A.F. Rheindahlen, BFPO 40.*

Meeting: 15th January 1974 (Membership Approval List No. 174)

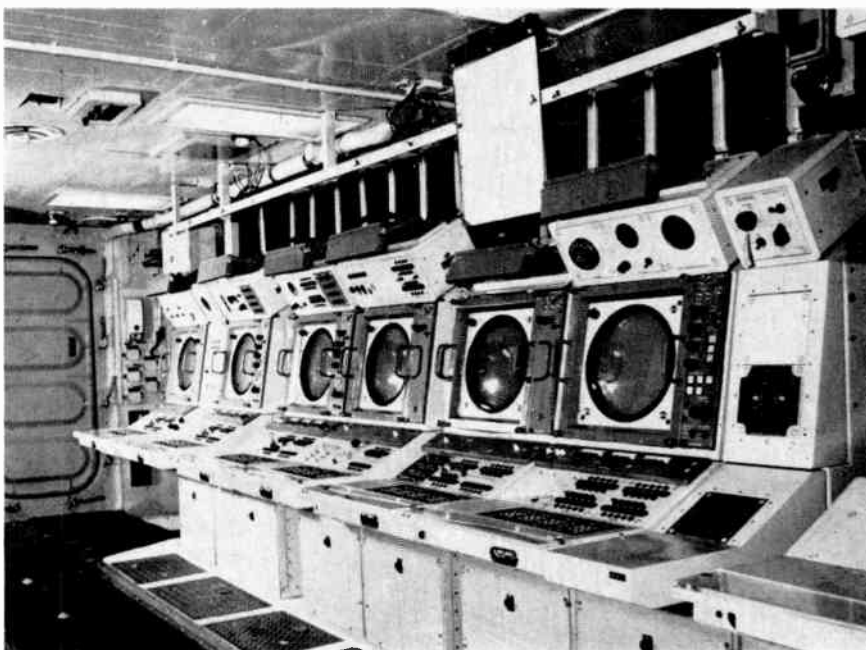### GREAT BRITAIN AND IRELAND

STUDENTS REGISTERED

ARGILE, Ian Wilmore. *Nottingham.*
BAHARUDIN, Nor Azmi. *Nottingham.*
BELLAMY, Paul Nigel. *Nottingham.*
BIRDSALL, Christopher Paul. *Nottingham.*
BRIGGS, Peter James. *Nottingham.*
CHOONG, Kok Ming. *Nottingham.*
CLEMENTS, Martin John. *Nottingham.*
COMBEN, Peter Paul. *Nottingham.*
COMLEY, John Stephen. *Nottingham.*
DAVIES, Michael John. *Nottingham.*
DEAN, Paul John. *Nottingham.*
DEY, Bidhan Ranjan. *Nottingham.*

DIMBLEBY, Barry John. *Nottingham.*
DORGAN, Michael Robert. *Nottingham.*
HALMONROYD, Anthony Ian. *Nottingham.*
HIRST, Robert. *Nottingham.*
JOHNSON, Stephen Gerard. *Nottingham.*
JOHNSON, Stephen. *Nottingham.*
KING, Derek. *Nottingham.*
LAYTON, Richard William. *Nottingham.*
LEE, John Robert. *Nottingham.*
McNULTY, Ian. *Nottingham.*
MALLET, Colin Thomas. *Bath, Somerset.*
MARVELL, Christopher John. *Nottingham.*
MOY, Anthony Jonathan Edward. *Nottingham.*
OPIE, Michael David. *Nottingham.*
PALMER, Darryl Anthony. *Nottingham.*
ROBINSON, Ian Michael. *Nottingham.*

RUSSELL, Michael Andrew. *Nottingham.*
SHENTON, Michael Paul. *Nottingham.*
SZENK, Christopher Alfons. *Nottingham.*
SHEPPERD, Elizabeth Jane (Miss). *Nottingham.*
SINGH, Saroop. *Nottingham.*
WEBB, Peter Jonathan. *Nottingham.*
WEST, Barry James. *Nottingham.*
WONG, Yip Heen. *Nottingham.*

OVERSEAS

STUDENTS REGISTERED

NJAU, Mark K. *Nairobi, Kenya.*
RAZALI, Bin Harun. *Parak, Malaysia.*

Notice is hereby given that the elections and transfers shown on Lists 171 and 172 have now been confirmed by the Council.

ADA Displays for HM Ships

Digital display equipment developed by Plessey Radar to form part of the Action Data Automation (ADA) Weapon System aboard the guided weapon destroyer HMS *Bristol*, the first operational vessel of the Royal Navy to be so fitted, is shown in the photograph.

The Plessey ADA display system comprises a number of labelled plan displays (l.p.d.s) and electronic data displays (totes) which operate in conjunction with display central equipment (d.c.e.). The displays are normally located in the operations room while the d.c.e. can be fitted remotely.

The display system operates in conjunction with a computer complex manufactured by Ferranti and based on the FM 1600 computer. The principal requirement of the display system is the provision of:

(1) The display of radar information, for the purpose of target detection and location, manual tracking and the initiation and supervision of automatic tracking.

(2) The display of synthetic data, stored in the FM 1600 computer, for the purpose of threat evaluation and weapon assignment. These data are displayed in the form of alpha-numeric and other symbols, bearing lines and dots.

Radar data are received direct from the sensors on board ship. Synthetic data are extracted from the radar data and are also received over data links from other sources.

The computer complex provides auto-tracking of selected targets and extracts target position, course and speed; these data are stored and updated. Data from other ships transmitted by link are also stored. The processed data are selectable at the displays, and are presented either as a filtered all-synthetic picture, or in association with raw radar from one of the ship's surveillance radars; the raw radar is available in the display system independent of the computer if required. All synthetic data are fed via the computer to the display system.

The display system consists of a display central equipment, usually situated alongside the computer in the computer room, and display consoles, the majority of which are located in the operations room. Five basic types of display console are used in the Plessey display systems. Each console incorporates a number of control panels and computer control and input devices. Certain controls act directly on the display system while others function indirectly. In the latter case the state of the control is made known to the computer, which interprets the information in the light of the program and modifies the associated messages.

The types of display fitted are:

(1) Labelled plan displays with full selection facilities for radar and synthetic data, tracker-ball and keyboard.

(2) Tote display (alpha-numeric read-out), tabular display with optional keyboard.

(3) Tactical displays. A combined labelled plan display and tote display having full selection facilities for radar and synthetic data and enabling three men to use these facilities whilst remaining seated.

(4) Master labelled plan display, with full selection facilities as in (1), tracker-ball and keyboard, capable of driving one or two slave displays.

(5) Slave labelled plan display, taking picture as compiled at master l.p.d. but with own tracker-ball and keyboard.

Variations of the type of presentation can be determined by the operator who adjusts the various controls at his disposal on the console. These controls act directly on the display as seen and also act as a method of communication between the operator and the computer.

Console suites can be built-up to suit operational requirements, by fitting l.p.d., tote auxiliary consoles in groups and incorporating the appropriate v.c.s. panels. By this means a tidy and uncluttered layout can be created to satisfy a particular operational task, thereby improving operator efficiency.

The tactical display console combines in a single housing a tote display and a modified form of l.p.d., having a large horizontal 22 in. display screen. It was designed as the modern solid-state replacement of the traditional projection-type plotting table. It provides access to all radar and synthetic data available in the system so that users can select the filtered data appropriate to the tactical situation. Its use solves the problem of providing a compact and efficient command area in the operations room.

Provision is made for seating three operators around the display and supplying each operator with a manual injection keyboard and tracker ball. The other controls for the l.p.d. application are mounted adjacent to the tote viewing unit.

Forthcoming Institution Meetings

London Meetings

Wednesday, 13th March

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

New Displays for Old Radar

By T. W. Welch (*T. W. Welch and Partners*)

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

In 1970, following operational and feasibility studies, it was decided to begin the development of a Civil Air Traffic Control Radar Service by the employment of an up-to-date data extraction, transmission and display system to present, for the time being, information derived from an existing secondary radar system, operated by a different agency from the ATC Authority, at a land-line distance of some 35 miles. A supply contract was effected in 1971 and the display system was installed and commissioned in the summer of 1972. This paper discusses various interference design problems encountered before, during and since the installation and commissioning of the system and the solution found effective for each. The topics discussed include protection of the existing radar system from fault conditions in the new equipment, early difficulties encountered in passing data over the existing telephone circuits, malfunctions caused by harmonic content in the selsyn azimuth data and by departures from specification of the video output of the original radar and some s.s.r. siting lessons learned from the results obtained.

Friday, 15th March

JOINT IEE/IERE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP

Some Recent Developments in Medical Ultrasonic Diagnostics

By Professor P. N. T. Wells (*Welsh National School of Medicine*)

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

Wednesday, 20th March

JOINT MEETING WITH IEE

Colloquium on TELECOMMUNICATIONS TEST—Philosophy of Test and Maintenance Equipment in the Next Five Years for use in Telephone Systems

IEE, Savoy Place, 10.30 a.m.

Apply for further details and registration forms to IEE, Savoy Place, WC2R 0BL

Wednesday, 20th March

COMMUNICATIONS GROUP

Global Communications

MEETING POSTPONED

Thursday, 21st March

THE EIGHTH CLERK MAXWELL MEMORIAL LECTURE:
'Our Space Programme after Apollo'

Dr. Wernher von Braun (*Fairchild Industries*)

London School of Hygiene and Tropical Medicine, 6 p.m. (Tea 5.30 p.m.)

Wednesday, 27th March

AUTOMATION AND CONTROL SYSTEMS GROUP

The Engineering Approach to CAD of Avionic Control Systems

By P. J. Burt (*Marconi-Elliott Avionic Systems*)

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

The methods used to design flight control systems tend to be influenced more by the engineering constraints than by the availability of sophisticated synthesis aids. Although many areas of flight control are well described by nearly linear and nearly time-invariant models, there are important exceptions. In any case the dynamics of the aircraft and hence the control laws are functions of height and speed. The engineering constraints are changing with the technology particularly due to the change from analogue to digital in-flight systems and the introduction of optically isolated consolidation of the lanes of a multi-lane defect survival system.

The CAD methods used in the author's company have tended to be either classical digital computer methods with manual trial-and-error backup by analogue simulation or automatic (hybrid) optimization by empirical means. The latter topic forms the major part of this lecture.

Friday, 29th March

JOINT IEE/IERE COMPUTER GROUP

Colloquium on INFORMATION STRUCTURES AND STORE ORGANIZATION

IEE, Savoy Place.

Further details and registration forms from IEE, Savoy Place, WC2R 0BL

Tuesday, 2nd April

AUTOMATION AND CONTROL SYSTEMS GROUP

One-day Colloquium on SIGNAL PROCESSING IN FEEDBACK CONTROL SYSTEMS

Middlesex Hospital Medical School, 10 a.m.

Determination and Realization of Compensation Filters

Computer-Aided Synthesis of Compensation Filters Required to Obtain a Specified Performance from a Closed-loop System Incorporating a Given Plant

By Dr. T. Konwerski (*University of Surrey*)

Compensation Filter Design via Low-order Modelling

By Lt. Cmdr. M. J. Ashworth (*Royal Navy*)

Digital Realization of Linear and Parametric Networks

By F. G. A. Coupe and Dr. J. I. Sewell (*University of Hull*)

Process Control Applications

Signal Adaptive Three-term Controller for Culham High Energy Laboratory Thermonuclear CLEO Experiment

By P. Atkinson and A. J. Allen (*University of Reading*)

The Design of an Adaptive Filter for Use in Process Control

By Dr. R. J. Smith-Saville (*Cambridge Consultants*)

Servomechanism Applications

Hardware Design of Adaptive Digital Filters Using L.S.I. and T.T.L. Chips

By Dr. D. R. Wilson, D. F. Neale and M. Burl (*Polytechnic of Central London*)

Adaptive Series Compensation of Stable Platforms

By Lt. Cmdr. R. Wilson (*Royal Navy*)

Transportation Applications

Design and Assessment of Signal Processing Filters for Adaptive, Automatic Control of Aircraft

By R. G. Harrison (*British Aircraft Corporation*)

The Use of a Discrete Kalman-Bucy Filter to Estimate the Misalignment Angles of a Ship's Inertial Navigation System at Sea

By Dr. M. Healey (*University College of South Wales, Cardiff*)

For further details and registration forms, apply to Meetings Secretary, IERE.

Monday, 8th April

JOINT IEE/IERE COMPUTER GROUP

Colloquium on THE MEASUREMENT OF COMPUTER PERFORMANCE

IEE, Savoy Place, 2.30 p.m.

Apply for further details and registration forms to IEE, Savoy Place, WC2R 0BL.

Wednesday, 10th April

COMPONENTS AND CIRCUITS GROUP

One-day Colloquium on SMALL POWER TRANSFORMERS AND CHOKES

Transformers—The State of the Art 1974

By J. W. McPherson (*Gardners Transformers*)

Magnetic Circuit Optimization for Miniature Power Transformers

By D. M. Fidler (*Tridem Transformers and Electronics*)

Life Testing of Chokes for Discharge Lamps

By Dr. T. R. Passmore (*Thorn Lighting*)

The Use of Matched Transformers for Synchro-Resolver Applications

By M. Pratt (*Ferranti*)

Constant Voltage Transformers

By A. Langley Morris

Designing for Production

By V. E. Cole (*Plessey*)

Ferrite Cored Transformers

By E. C. Snelling (*Mullard*)

IERE Lecture Room, 10.30 a.m.

For further details and registration forms, apply to Meetings Secretary, IERE

Wednesday, 17th April

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Colloquium on ELECTROMAGNETIC COMPATIBILITY OR CONFUSION— EMC ON LAND, AT SEA AND IN THE AIR

IERE Lecture Room, 11 a.m.

For further details and registration forms, apply to Meetings Secretary, IERE.

Tuesday, 23rd April

Please note change of date

COMMUNICATIONS GROUP

Uncertainties and Inertia in Long Range Telecommunications Planning

By A. A. L. Reid (*Post Office Long Range Studies Divisions*)

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

Wednesday, 24th April

JOINT IERE/IEE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP

Colloquium on HEARING AIDS

IERE Lecture Room, 10 a.m.

For further details and registration forms, apply to Meetings Secretary, IERE.

Kent Section

Thursday, 21st March

ANNUAL GENERAL MEETING at 7 p.m.

Followed by

Modern Railway Signalling Systems

By R. M. Bell and P. M. Flook (*British Rail*)

Lecture Theatre 18, Medway and Maidstone College of Technology, Maidstone Road, Chatham.

The feature of the lecture is an hour-long sound film, made by the authors, 'First of the Thirteen', and is a documentary of the British Rail, Southern Region North Kent Lines resignalling scheme. The introduction

will describe the events leading up to the scheme, the making of the film and Southern Region's 'Thirteen signal box plan'. After the film the present London Bridge re-signalling scheme will be introduced briefly, pointing out the similarities and variations with the North Kent scheme, together with novel features at present being developed. The presentation is the basis for a future lecture on the use of electronics on the London Bridge scheme.

Thames Valley Section

Thursday, 14th March

Recent Advances in Radio Navigation

By J. E. Viles (*Marconi Avionic Systems*)

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

A general survey will be given of new navigational aids such as Omega and aircraft microwave landing systems and the integration of aids in area navigation and mixed systems, with reference to some modern techniques which have helped these advances in the airborne equipment field.

Thursday, 4th April

ANNUAL GENERAL MEETING at 7 p.m.

Followed by **C-MOS and its Applications**
By W. G. Saich (*National Semiconductors UK*)

J. J. Thomson Physical Laboratory, University of Reading, Whiteknights Park, Reading.

Complementary m.o.s. is the ideal logic family as it approaches the perfect characteristics of no power, zero propagation delay, controlled rise and fall times, and has high noise immunity. This lecture will discuss the characteristics of the devices available, interfacing to other logic families, and non-logic applications.

Southern Section

Wednesday, 13th March

ANNUAL GENERAL MEETING at 6.30 p.m. Followed by

The History of some Major Inventions and Concepts in Electrical Communications

By Professor D. G. Tucker (*University of Birmingham*)

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

Many of the basic concepts in electrical communications today have a surprisingly early origin in the nineteenth century, although they were often not understood and were often misunderstood. The lecture will examine some of these origins with particular reference to ideas and methods of multiplex telephony for both line and radio, including frequency-division and time-division multiplex, and of amplitude and frequency modulation; it will conclude with a consideration of the influence of the thermionic valve and positive and negative feedback.

Wednesday, 20th March

JOINT MEETING WITH IEE

Colloquium—ENERGY IN THE 80's

Lectures by
G. N. Stone (*CEGB, Bristol*), Dr. I. Fells (*University of Newcastle Upon Tyne*) and P. T. Cast (*Shell Mex and BP*)

Registration not necessary; visitors welcome. Lanchester 'A' Lecture Theatre, Southampton University, 4 p.m.

Wednesday, 27th March

JOINT MEETING WITH IEE

Multi-channel U.H.F. Reception on Naval Ships

By H. P. Mason (*ASWE*)

HMS *Collingwood*, Fareham, 6.30 p.m.

Tuesday, 2nd April

Charge Coupled Devices

By J. D. E. Beynon (*University of Southampton*)

Brighton Technical College, Pelham Road, 6.30 p.m. (Tea in Refectory)

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 field. 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.ds many present and future applications will be discussed.

Yorkshire Section

Thursday, 14th March

The Use of Circuit Blocks in Digital Systems

By J. M. Reid (*South Yorkshire Automatic Control Engineering*)

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

A survey of discrete component circuit blocks available at the present time and the detailed application of one product range in an industrial system will be given. The challenge from various integrated circuit families will be discussed and probable developments outlined.

Friday, 26th April

ANNUAL GENERAL MEETING

Leeds University, 7 p.m.

West Midland Section

Monday, 25th March

IEE MEETING

Radio Telemetry Network for the Retrieval of Seismic Signals

By Z. E. Jaworski (*City of Birmingham Polytechnic*)

M.E.B. Offices, Summer Lane, Birmingham at 6 p.m.

Wednesday, 3rd April

ANNUAL GENERAL MEETING at 7 p.m. Followed by

Recognizing Musical Instruments

By K. A. Macfadyen (*University of Birmingham*)

The Polytechnic, Wolverhampton.

East Midland Section

Wednesday, 27th March

ANNUAL GENERAL MEETING

Leicester Polytechnic, 7.30 p.m.

The development of an electronic performer-instrument (as distinct from a synthesizer) for players in amateur ensembles has demonstrated that the recognition of musical instruments from extended passages played on them depends on a number of rather diverse physical attributes of the sounds produced. Considerations of complexity and cost of the instrument have led to an awareness of the most important attributes.

South Midland Section

Thursday, 21st March

Fibre Optic Communications

By Professor W. A. Gambling (*Southampton University*)

G.C.H.Q., Oakley, Cheltenham, 7.30 p.m.

Wednesday, 24th April

Developments in Electronics for Motor Vehicles

Speaker to be announced

To be followed by

ANNUAL GENERAL MEETING

The Foley Arms, Malvern, 7 p.m.

South Western Section

Wednesday, 20th March

JOINT MEETING WITH IEE

Aphid Numerical Control for Printed Circuit Card Layouts

By G. Skidmore (*British Aircraft Corporation*)

Lecture Room 2E3.1, University of Bath, 7 p.m. (Tea 6.45 p.m.)

Wednesday, 3rd April

Electronics in Support of the Police

Speaker to be announced

Lecture Room, 2E3.1, University of Bath, 7 p.m. (Tea 6.45 p.m.)

Monday, 6th May

ANNUAL GENERAL MEETING

The Royal Hotel, College Green, Bristol, 7 p.m.

North Eastern Section

Wednesday, 13th March

Electrical Filters

By Professor A. G. J. Holt (*University of Newcastle Upon Tyne*)

The Main Lecture Theatre, Ellison Building, Newcastle Upon Tyne Polytechnic, Ellison Place, Newcastle Upon Tyne, 6 p.m. (Refreshments in Staff Refectory 5.30 p.m.)

Wednesday, 10th April

Communication with Light

By Professor W. A. Gambling (*University of Southampton*)

Followed by ANNUAL GENERAL MEETING

The Main Lecture Theatre, Ellison Building, Newcastle Upon Tyne Polytechnic, Ellison Place, Newcastle Upon Tyne, 6 p.m. (Refreshments in Staff Refectory 5.30 p.m.)

North Western Section

Thursday, 21st March

Machine Tool Control: The Interfacing of NC to the Machine Tool

By R. W. Prew (*Plessey Numerical Controls*)

Lecture Theatre R/H10, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.)

The paper will cover the evaluation of the interface from the Ferranti Mark IV equipments up to present soft wired developments.

Thursday, 9th May

ANNUAL GENERAL MEETING

Lecture Theatre R/H10, Renold Building, UMIST, 6.15 p.m. (Tea 5.45 p.m.)

Merseyside Section

Wednesday, 13th March

Magnetic Bubbles and their Applications

By P. J. Banks (*Liverpool Polytechnic*)

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

Magnetic bubbles are cylindrical magnetic domains which are formed in films of suitable materials by applying a bias field normal to the plane of the film. The

bubble domains can represent a bit of information, and thus can be used both for the storage and the manipulation of digital data. Work has been done on them by several manufacturers interested in computer applications.

An explanation of the basic requirements for the formation of bubbles will be given as well as descriptions of some of the methods proposed for generating, moving and detecting them.

Wednesday, 24th April

ANNUAL GENERAL MEETING at 7 p.m. Followed by

Wide Ranging Applications of Metal Oxide Semiconductors

By J. A. Everist (*ITT Semiconductors*)

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

Northern Ireland Section

Wednesday, 13th March

Megaw Memorial Lectures

The Main Lecture Theatre, Ashby Institute, 6.30 p.m.

The Megaw Memorial Award is given annually for the most outstanding paper presented at a meeting of the Institution's Northern Ireland Section by a final-year student of Electrical Engineering at the Queen's University of Belfast.

Wednesday, 3rd April

Please Note Change of Date

Making Good Television

By W. Woods (*B.B.C.*)

Cregagh Technical College, 7 p.m.

South Wales Section

Monday, 18th March

JOINT MEETING WITH IEE

Automobile Electronics

By C. S. Rayner (*Lucas Electrical*)

Department of Applied Physics, UWIST, Cardiff, 6 p.m. (Tea in College Refectory from 5.30 p.m.)

Thursday, 11th April

JOINT MEETING WITH IEE

Quadrophonics

By K. Barker (*University of Sheffield*)

University College, Swansea, 6.15 p.m.