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Communication between People

AMONG the many greater and lesser anniversaries which fall during 1976, there are three of special importance, each one of which has had an enormous, almost incalculable effect on mankind. The 'man-in-the-street' will certainly be aware of the Declaration of Independence by the 13 American Colonies in 1776, which represented the birth of the United States of America.

But two less well-known and, at the time, less spectacular events took place respectively 300 years earlier and 100 years later. In 1476 William Caxton set up his printing press in the precincts of Westminster Abbey in London and printed his first works in English—not classical texts for classical readers, as Gutenberg's original machine had been producing for several years in Cologne, but books intended for the 15th century man-in-the-street. Caxton's enterprise in popular publishing entitles him to be regarded as a creator of modern communications.

So, too, may the inventor linked with our third anniversary—Alexander Graham Bell—be regarded as a creator of modern communications. It was on 10th March 1876 that Bell, a Scotsman working in Boston, Massachusetts, achieved the first successful transmission of intelligible speech, and this event has been widely commemorated by telephone organizations throughout the world and by a meeting at the IEE supported by this Institution. Today there are estimated to be nearly 400 million telephones throughout the world and over 20 million in the United Kingdom—the rate of expansion in recent years has been such that the first half of the British telephone system took ninety years to build! Nevertheless it must be agreed that the immediate exploitation of Bell's work was rapid, for in 1879 the first telephone exchange in England was opened in London by the Telephone Company Limited. A service with France was established in 1891 and in 1896 the Post Office took over the National Telephone Company's trunk service and in 1912 all that Company's exchanges. In the same year the first automatic exchange in the UK was opened at Epsom.

International landmarks in the history of the telephone have been the London-New York service in 1927 and services with India in 1933 and Japan two years later. These developments called in the skills of the radio engineer and the subsequent history of virtually every telephone advance has depended on electronics. Thus the first submerged repeater was laid in the Irish Sea in 1943 and in 1956 the first Trans-Atlantic telephone cable came into operation, followed by other Atlantic and, later, Pacific links. Telephony in the 'sixties saw major electronic developments in two areas—the electronic exchange and, even more striking, the advent of the communications satellite, first proposed in 1945 by the Englishman, Arthur C. Clarke. After the launching of *Telstar* in 1962 steady progress was made towards greater and greater capacity and complexity and since 1970 the *Intelsat* satellites have served all the continents, complementing the still expanding submarine cable network.

Worldwide communications represent an impressive modern technological achievement as significant as the printed word, with which in teleprinter, facsimile, and, now, teletext, it is becoming closely linked. Many important innovations have originated in the United States but communications are, almost by definition, universal and contributions to our present stage of development have come from many countries. For the second century of the telephone the Managing Director of Telecommunications at the British Post Office, Sir Edward Fennessy, forecast at the commemoration meeting the growth of wholly automatic networks with a host of ancillary services. He continued:

'The global system will place at the disposal of mankind vast resources of information and advice on the conduct of its affairs. How mankind will employ this gigantic global communication system is less easy to predict. We—the engineers and system planners—can only follow the objectives of Alexander Graham Bell and seek to improve communication between people.'

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* See also page 120.

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Stationary patterns in radar sea clutter

R. G. TAYLOR, B.Sc.*

SUMMARY

Experimental observations with coastal radars at 3 and 10 cm wavelengths have revealed patterns in the sea returns which are stationary with respect to the radar site. The patterns take the form of a family of curves which exhibit a linear dependence upon wavelength and is characteristic of interference fringes. The mechanism proposed in the paper is forward scatter over the sea surface and subsequent combination with the direct beam. The effect upon radar detection performance is briefly discussed.

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1 Introduction

References 1 and 2 describe work which is concerned with establishing the degree of improvement of detectability of small targets in sea clutter by the use of high rotation rate antennas (up to 1000 rev/min) to decorrelate the sea clutter. The technique employed to establish the improvements for various sea states uses camera integration of the p.p.i. displays to compare the normal radar with the decorrelated radar. The integration times may be of the order of several minutes.

It was noted quite consistently that when camera integration times of about half a minute upwards were used, giving considerable smoothing of the moving clutter, a stationary pattern appeared in the smoothed clutter, and the longer the integration time the more sharply defined the pattern became. The radar was in X band (3 cm) with its antenna (variously 2° or 1° azimuth) mounted some 14 m above sea level overlooking the beach at Southsea, the water's edge being about 140 m distant. The glancing angle on the sea was therefore less than one degree at half a nautical mile. A pulse length of 0·1 μs was used throughout the work.

2 Observed Patterns

With a normal 20 rev/min, 1° beam antenna, where most of the clutter pulses within the beamwidth are correlated, it takes several minutes of camera exposure to smooth the clutter from the travelling waves and see the stationary pattern. However, with a 400 rev/min antenna giving 2 pulses beamwidth, successive antenna scans 0·15 s apart are decorrelated, and a sufficiently smooth clutter picture builds up more rapidly.

Figure 1 shows a picture of sea clutter from 1·5 m waves recorded on the 20 rev/min antenna (repetition rate 5000); a single revolution of the display has been photographed by a 3 s camera exposure and this is very close to what the eye sees when observing a normal display.

If the picture is photographically integrated for 4 min (recording 80 revolutions of the antenna while the camera shutter remains open) a very clear pattern is seen in the smoothed clutter. In fact, using camera integration with a 400 rev/min antenna (decorrelated clutter) a similarly vivid pattern can be built up in only 32 s of exposure (Fig. 2). This pattern looks like a family of hyperbolae about an axis through the radar origin and normal to the shore-line. As would be expected, several targets can now be seen in the smoothed clutter.

The radar has no precision range marker facility, nor are there any rigidly fixed navigation marks in the sea (the buoys can swing about 12 m with tide) but within such limits of observation, the pattern seemed invariant with tide and wind direction, and could invariably be observed given (a) sufficient clutter to provide a record, (b) sufficient exposure time.

First thoughts were that the picture might represent a true set of highly regular standing waves of this shape in the sea upon which the random sea variations were superimposed. The area of observation is roughly

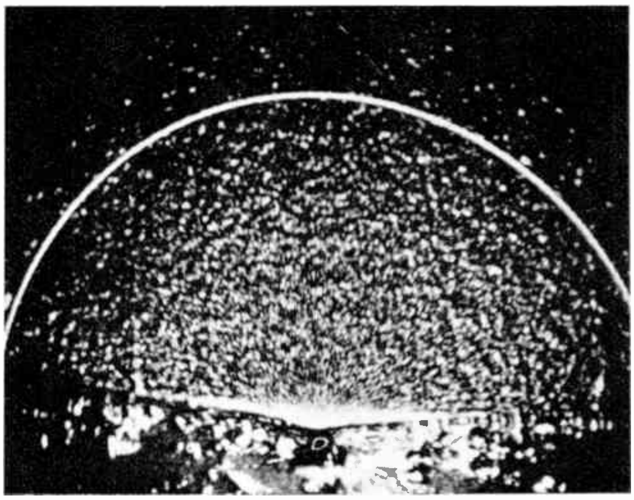

Fig. 1. Clutter from 1·5 m waves. 20 rev/min antenna (1 rev, 3 s), range ring 1·0 nautical mile.

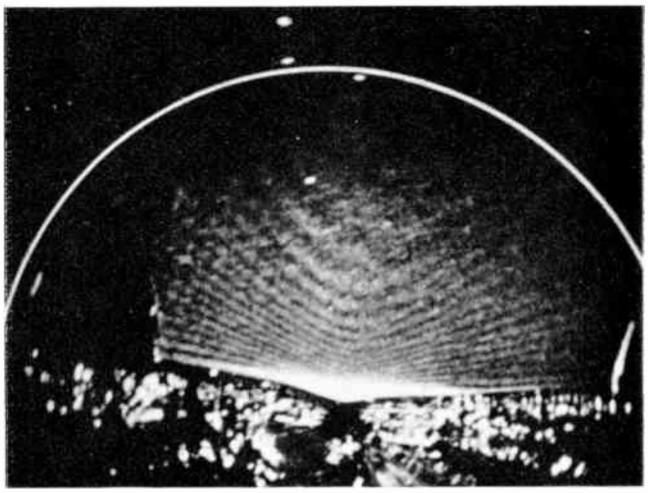

Fig. 2. Clutter from 1 m waves. 400 rev/min antenna, 32 s exposure.

equidistant between the dredged tideways of Portsmouth and Langstone Harbours (one mile to either side).

Examination of the navigation chart of the area, however, showed no features likely to produce so regular a pattern and other causes were considered.

It seemed worth eliminating any possible instrumental effect in the radar itself, and since another X-band radar was available, with an antenna at similar height, some 50 m further along the coast, records were taken from this. This radar was only capable of operating in the clutter-correlated condition (20 rev/min antenna), and did not have the same constant-false-alarm-rate facilities as the first receiver, but an exposure time of 8 min clearly revealed a pattern of closely identical shape and spacing to that derived from the other radar.

The vertical beamwidth of the radar antennas in use was 20° and this was large enough to give strong illumination of the foreshore land area as well as the line of surf breaking on this foreshore. If the pattern is a set of interference fringes arising from a forward scatter multipath mechanism a significant increase of wavelength should open out the pattern correspondingly. When, for other work, a 10 cm radar (of similar vertical and horizontal beamwidth) was temporarily set up on the same site, it was only available for any pattern investigations over a period of three weeks. On only one day of this three-week period was the sea marginally rough enough (1 m waves) to show adequate clutter, but on that day the result shown in Fig. 3 was obtained using a 4-min exposure.

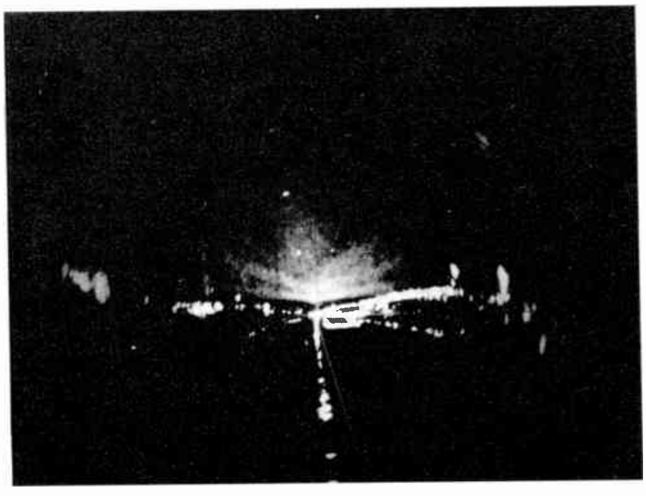

Fig. 3. Clutter from 1 m waves. 10 cm marine radar, 20 rev/min antenna, 4 mn exposure.

This radar again was a simple 20 rev/min correlated system. Clearly the scale of the pattern has changed by a factor of about three (wavelength ratio) and it looks as if a multipath interference mechanism is at work.

3 Loci of the Fringes

The model which is suggested as the cause of the experimental observations is illustrated by Fig. 4. OX and OY are Cartesian axes with OY taken along the axis of the family of curves.

The radar beam from the antenna at A illuminates the sea at point P by two paths, the direct beam AP and the indirect beam ADP. The latter beam is the result of diffraction from a point D situated at a distance L from the foot of the radar and at an elevation s. The

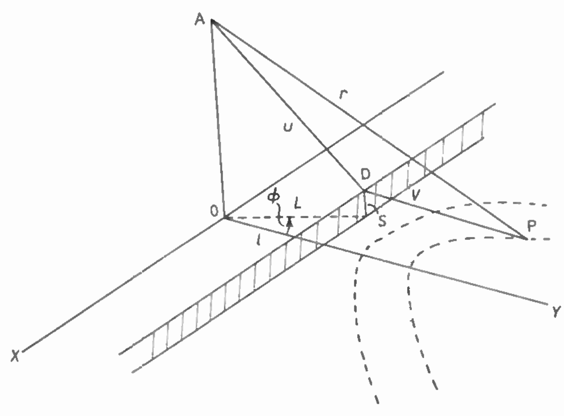

Fig. 4. Experimental configuration.

antenna height and grazing angles, γ, to be considered satisfy the inequality[3]

$$\gamma > \tan^{-1}(0{\cdot}0018\sqrt{h})$$

where h is the antenna height in metres, and an assumption of a flat earth model is justified.

From the figure we may write the path difference for a dark fringe of order n as

$$u+v_n-r_n = n\lambda \qquad (1)$$

i.e.:

$$[L^2+(h-s)^2]^{\frac{1}{2}}+ \\ +[r_n^2-h^2-2L(r^2-h^2)^{\frac{1}{2}}+L^2+s^2]^{\frac{1}{2}}-r_n = n\lambda \qquad (2)$$

A phase change of π has been assumed on diffraction.

A simple and accurate method of solving equation (2) for the parameter s is by induction. Values of r_n along the curve axis were measured from the p.p.i. photographs for two fringes separated in order by $k = 4$. Using a computer program the corresponding equations (2) were solved for preset values of k, the order difference, with incremental values of s and L. The parameter s was incremented up to a maximum value determined by the intercept at the corresponding value of L with the line joining the antenna and the waterline. At each step (s, L) the order difference was calculated and compared with the preset value for equality. The solutions are plotted in Fig. 5 as a function of k and it is evident that the observed curves correspond to a value of $s = 0$ for all L. This result implies that the diffraction point D is on or very close to the sea surface and equation (2) is re-written

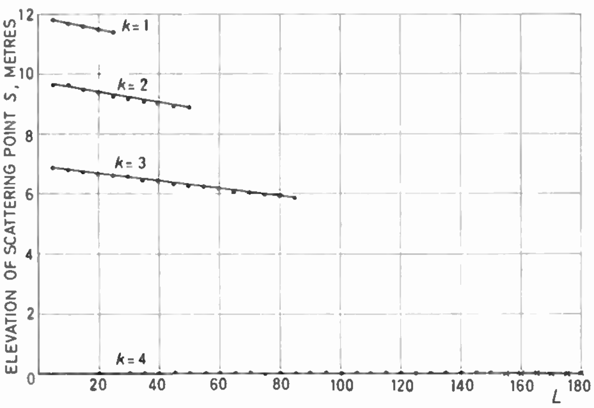

Fig. 5. Fringe order difference as a function of scattering point location.

$$(L^2+h^2)^{\frac{1}{2}}+(r_n-h^2)^{\frac{1}{2}}-L-r_n = n\lambda \qquad (3)$$

This equation allows the height of the aerial above the point D to be estimated. Subtraction of the corresponding equations (3) for fringes of order n and $n+k$ yields the difference equation

$$(r_{n+k}^2-h^2)^{\frac{1}{2}}-(r_n^2-h^2)^{\frac{1}{2}} = k\lambda+(r_{n+k}-r_n) \qquad (4)$$

Substitution into this equation of the measured values

$$r_{n+k} = 859 \text{ metres}, \qquad r_n = 222 \text{ metres}, \\ k = 9, \qquad \lambda = 0{\cdot}03 \text{ metres}$$

reveals that the effective height of the aerial above the scattering point is $h = 12{\cdot}7$ m. From a survey of the terrain elevation, Fig. 6, between the radar and the water it is found that the distance to the scattering point along the curve axis, $(L = l)$, is

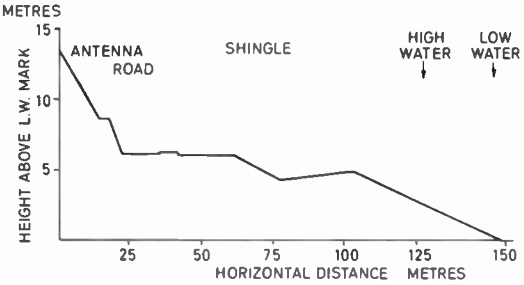

Fig. 6. Survey of land area at close range.

$$l = 139{\cdot}6 \text{ metres.}$$

The point D is therefore placed in the water along the shoreline. The low and high water marks occur at distances of 156 m and 130 m and these ranges agree well with the estimated value of l. It seems reasonable to assume that waves breaking near the beach create a region of rough water from which the radar beam is diffracted to give a surface wave. Interference between this and the direct beam forms the observed family of curves. The ranges to the vertices of the curves as a function of order may be obtained by expanding equation (3) according to the Binomial Theorem. We find, to 1st order,

$$r_n = \frac{h^2}{2\{(L^2+h^2)^{\frac{1}{2}}-L-n\lambda\}} \qquad (5)$$

Calculations of the predicted ranges have been made with this equation fitted to the observations ($\lambda = 0{\cdot}03$ m, $L = 139{\cdot}6$ m) and a comparison with the measured values is given in Table 1. The agreement is good although with discrepancies which may be attributed to

Table 1
Fringe positions along the axis
Normal distance to diffracting region, $l = 139{\cdot}6$ m

Order	Range	
	Theoretical	Experimental
5	189	—
6	203	—
7	220	222
8	240	244
9	263	275
10	292	303
11	327	346
12	373	408
13	432	483
14	515	564
15	638	669
16	836	859
17	1213	—
18	2210	—

Maximum order = 19·22

small errors introduced by approximation and experimental factors. It is interesting to note that the interference effect is predicted to extend to infinity. The form of the individual curves is given by writing

$$L = l \sec \phi \qquad (6)$$

in equation (5) and calculating r_n as a function of ϕ. A plot of the curves is given in Fig. 7 where it is seen that the general nature corresponds to the p.p.i. photographs.

The detailed form of the curves depends upon the shape of the diffracting region. In general this will not be a straight line and there will be variations in the

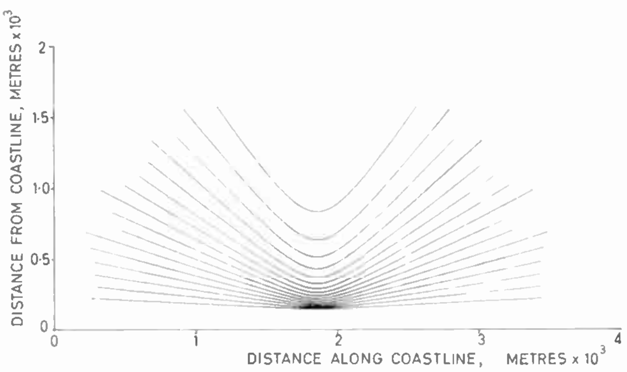

Fig. 7. Predicted loci of the interference fringes.

normal distance (l) to a line through the diffracting region at the point D parallel to the axis OX. Values of l were estimated as a function of ϕ using equations (5) and (6) for the fringe of order 10 with measured values of range. The vertex of this fringe is at a range of 408 m and the results are listed in Table 2. It will be noticed that the distances decrease away from an angle of $\phi = 15$ deg and this suggests that the shape of the diffracting region along the coast is concave towards the radar although with a large radius of curvature. Examination of a chart[4] of the area shows that this behaviour corresponds with that of the drying line and it may be that this has an influence on the wave structure near the shore. Measurements of greater accuracy than those under discussion would be required to support this hypothesis.

Table 2
Normal distance to the diffracting region
Range of curve along axis = 408 m

Angle (deg)	Range (m)	Normal distance
0	408	139·6
5	423	140·6
10	445	141·2
15	477	141·4
20	515	140·5
25	578	139·6
30	719	140·3
35	810	135·9
40	1260	136·5

For comparison purposes it is of interest to deduce the dependence upon the height of the aerial. From equation (3) on application of the Binomial Theorem, we obtain the approximate expression

$$\frac{h^2}{2L} - \frac{h^2}{2r_n} = n\lambda \qquad (7)$$

The number of orders between two ranges r_n and r_{n+k} is

$$k = \frac{h^2}{2\lambda}\left(\frac{1}{r_n} - \frac{1}{r_{n+k}}\right) \qquad (8)$$

and we see that the spacing of the fringes varies as the square of the aerial height. It follows that the fringe pattern closes or opens as the aerial is raised or lowered. The absence of the factor L shows that the change in shape of the curves is of second order or less. A plot of the relationship between k and h based on accurate calculations made from equation (3) is given in Fig. 8.

4 Quantitative Estimates of Signal Strength

It is readily shown that the scattering areas may be classed as rough surfaces by the substitution of experimental values into the Rayleigh phase criterion for diffuse scatter

$$\Phi = \frac{4\pi\sigma \sin\gamma}{\lambda} < \frac{\pi}{2} \qquad (9)$$

The parameter σ in this expression is the r.m.s. height of the surface. The best estimate of this is $\sigma \simeq 1{\cdot}5$ m and it follows that diffuse scatter is obtained for grazing angles satisfying the inequality

$$\gamma > 0{\cdot}14 \text{ deg.} \qquad (10)$$

This value of γ does not occur until a range of 5 km from the radar. In addition, the scale of roughness is greater at the sea–land boundary where the surface beam is formed than it is over the remainder of the experimental area. Consider a model for the sea surface which is defined by a normal distribution of wave height, with standard deviation σ, and an autocorrelation function

$$c(\tau) = \exp\left(-\frac{\tau^2}{T^2}\right) \qquad (11)$$

where T is the isotropic horizontal autocorrelation distance. The reflexion coefficient of a smooth, plane

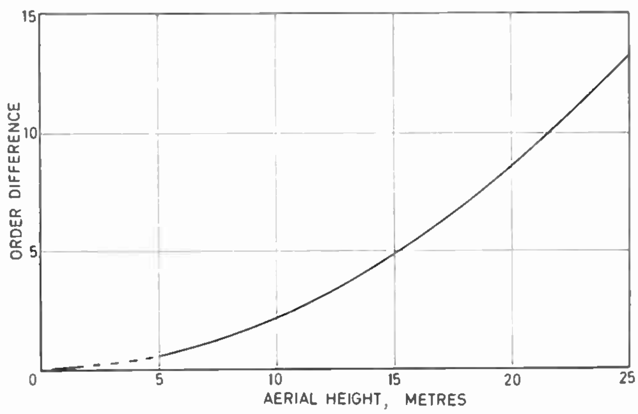

Fig. 8. Variation of order difference with aerial height.

sea surface of infinite extent is taken to be -1 and, since specular reflexion is negligible, the mean square reflexion coefficient δR_1 of a surface element δS may be written

$$\langle|\delta R_1|\rangle^2 = \frac{\delta S}{\pi r_2^2}\frac{r^2}{r_1^2}\frac{1}{\cos^4\beta}\cot^2\beta_0 \exp\left(-\frac{\tan^2\beta}{\tan^2\beta_0}\right) \quad (12)$$

The ratio of the direct and indirect beams illuminating the surface is the quantity of interest and (12) has been written in these terms.

The meaning of the terms in these equations is illustrated by Fig. 9. The angle between the bisector of the angle between the incident and scattered rays and the vertical axis is β and β_0 is given by

$$\tan\beta_0 = \frac{2\sigma}{T} \quad (13)$$

This equation describes the mean slope of the surface features. The distances r, r_1 and r_2 are respectively the distances from the radar antenna to the sea surface, to the localized scattering region at the water's edge and from the localized region to the resolution cell on the sea surface.

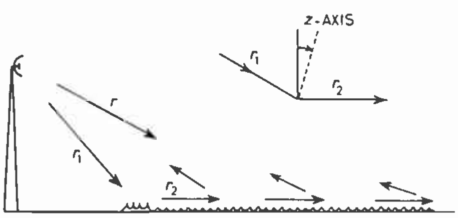

Fig. 9. Diffuse scatter and surface illumination.

The relationship at various sea states between the r.m.s. wave heights and the corresponding wavelengths[6] is ~ 20 and we may assume that the angle β_0 is small. In the experiments which have been considered the angle β is also small and equation (13) may be simplified to

$$\langle|\delta R|\rangle^2 \simeq \frac{\delta S}{\pi}\frac{r^2}{r_1^2 r_2^2}\cot^2\beta_0 \exp\left(-\frac{\tan^2\beta}{\tan^2\beta_0}\right) \quad (14)$$

The extent in azimuth of the area illuminated by the radar is given by $r\theta$, where θ is the beamwidth of the antenna taken by convention to the -3 dB points for the purposes of calculation. At the position near the waterline from which the quasi-specular beam is scattered we have

$$r\theta \simeq 1{\cdot}2 \text{ m}$$

A further application of the Rayleigh criterion for specular reflexion reveals that the phase difference between the scatter from the centre of the beam and from the edges is approximately $0{\cdot}7\pi$.

The Rayleigh criterion is not exact and it is reasonable to assume that the phase of the scatter across the illuminated area is constant. This simplifies the integration of δS with small error and enables the mean square reflexion coefficient of the quasi-specular scatter to be written as

$$\langle|R|\rangle^2 = \frac{r^2\theta c\tau}{2\pi r_1 r_2^2}\cot^2\beta_0 \exp\left(-\frac{\tan^2\beta}{\tan^2\beta_0}\right) \quad (15)$$

It may be mentioned here that the result (of effective specular reflexion) was implied by the qualitative prediction that an interference mechanism was present. The expression given in equation (15) provides the means of estimating the relative magnitude of the indirect beam. At a range of 400 m for example the value of

$$\langle|R|\rangle = 0{\cdot}24$$

is obtained for the following experimental values

$$r = 400\text{ m},\quad r_1 = 140\text{ m},\quad r_2 = 260\text{ m},$$
$$c = 3\times10^8\text{ m s}^{-1},\quad \theta = 1\text{ deg},\quad \tau = 0{\cdot}1\ \mu\text{s},$$
$$\tan\beta_0 = 0{\cdot}15,\quad \tan\beta = 0{\cdot}045$$

The angular difference in elevation between the two illuminating beams at the resolution cell of interest is small and differences in the reflexion coefficient of the sea backscattered echo may be neglected. At the maxima of the interference pattern, therefore, we may estimate that the surface is illuminated by a field of relative magnitude $1{\cdot}24$ ($+2$ dB) and at the minima by a relative magnitude of $0{\cdot}76$ ($-2{\cdot}5$ dB). These small values are consistent with the need for a considerable reduction in the variance of the normal backscattered echo before the fringes become apparent.

5 Implications of Quasi-specular Reflexion

The significant factor in this work is that it establishes the existence of an effectively specular reflexion from a natural feature of the sea surface. It has previously been generally assumed that, since the wave structure is described by a statistical function and thus averages to a mean level, the radar scatter may be similarly treated. Variations of the function over the extent of the surface illumination have not been included. A consequence of this reasoning is that the indirect illumination of an elevated target by scatter from the surface contains two components, specular and diffuse, which are respectively coherent and incoherent with respect to the direct beam and which disappear when the target area lies in the surface. In contrast to this view the present theory suggests that a third component may be present, characterized in magnitude by diffuse scatter and in phase by specular reflexion according to the geometry of the experimental configuration. This component combines coherently with the direct illumination and does not disappear with surface targets.

The relatively small magnitude of the interference effect shows that it is not important in sea clutter studies with systems in general use, that is, those that integrate correlated returns. Experimental errors would normally obscure the phenomenon. However, these conventional radars are severely limited in their ability to detect targets of small radar cross-section in sea clutter and it is likely that improvements will be required in the future. Integration of the decorrelated sea echo offers substantial potential gains which quasi-specular scattering reduces by approximately 2 dB, a figure that would not be negligible in predicting system performance.

The limitation occurs in two ways. Firstly, the target is illuminated at a weaker signal strength in the minima and secondly, an automatic system would need to place its threshold detector at a higher level, to avoid a high false alarm rate caused by the maxima.

Most of the targets illuminated by radar systems do not lie in the surface since they are either surface vessels or airborne objects. The finite vertical dimension means that estimates of the radar performance need to include all three components and the possibility arises of modification to the simple multi-path pattern. In the direction of specular reflexion the magnitude of the quasi-specular component has a maximum value[5,7] of about 0·35 and the simple specular reflexion has a magnitude which is related to the system geometry and surface roughness by the expression

$$\langle \rho \rangle = \rho_0 \exp\left(-\frac{\Phi^2}{2}\right) \qquad (16)$$

ρ_0 is the smooth surface reflexion coefficient, previously assumed to be -1. The nett relative signal intensity, E, at the target is the resultant of the three components. That is

$$\langle E \rangle = 1 + \langle \rho \rangle \cos(\varepsilon) + \langle R \rangle \cos\eta \qquad (17)$$

The parameters ε and η in this expression are relative phases and it is clearly possible for the signal strength to be increased or decreased in any given situation. In particular the model predicts that minima will be produced in the propagation pattern that are not expected on the basis of simple multi-path theory.

The discussion in this paper has been in the context of a coastal radar illuminating an area of quasi-specular scattering near the waterline. The radar site is not important to this theory and the quasi-specular scatterer may be generalized to one of significant radar cross-section with restricted linear dimensions. The bulk movement of water against underwater obstructions, for example, causes areas of surface roughness which would be expected to behave in a similar manner and influence the operation of shipborne radars. Also, the possibility should not be overlooked that the superstructure could provide a suitable localized scattering area, resulting in a ship-referenced interference pattern.

6 Conclusion

The stationary patterns which have been observed in radar sea returns have been shown to be due to an interference mechanism arising from diffuse scatter from a localized area. Coherent interference of the narrow phase distribution of the diffracted beam takes place with the direct illumination of the surface to form an interference fringe pattern. The loci of the family of curves predicted by the proposed model are in good agreement with those found in the experimental work and a quantitative estimation of signal strength is consistent with the apparent magnitude of the effect. The visibility of the fringes is determined by the amount of radar backscatter and this suggests that a critical sea state exists for the onset of the phenomenon.

The implications of the model for the detection of targets which are raised above the surface have been discussed. Illumination of the target area is by the resultant of three beams and as a consequence it is predicted that additional nulls appear in the normal two-beam multi-path propagation pattern. These minima may lead to a loss of target detection.

Localized surface disturbances may be caused by underwater obstructions and produce a quasi-specular reflexion within the coverage of a shipborne radar. The backscatter from these areas would exceed that from surrounding areas and this may be of interest in navigation and oceanography for the mapping of hazards and sea bed profiles.

7 Acknowledgments

The author is indebted to Professor J. Croney for making available his experimental results shown in Figs. 1, 2 and 3.

Appreciation is expressed of the discussions held with Professor Croney and Mr. B. Gladman and of the survey carried out by Mr. M. Waller.

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Analysis of the frequency and sensitivity of crystal-controlled oscillators

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SUMMARY

The equations for the frequency of a crystal oscillator, and the sensitivity of frequency to the circuit parameters are here extended to consider: (1) the effects of motional capacitance, tuning capacitance, circuit capacitors, and the active element impedance; and (2) the Z-type oscillator.

A precise normalized frequency difference between the frequency of oscillation and the series resonant frequency of the crystal is defined and derived for both the Y-oscillator and Z-oscillator. A precise normalized frequency difference is likewise derived for the parallel resonant frequency of the crystal. From these, the sensitivity of oscillation frequency to various parameters of the crystal and the circuit is derived. The conditions of validity and practical significance of the expressions are discussed. Experimental results obtained with the Pierce oscillator verify the applicable equations.

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1 Introduction

In stable crystal oscillators, it is essential to know the influence of the various circuit elements on the frequency of oscillation. It is also essential to know the influence of the crystal parameters (in addition to resonant frequency of oscillation). The formulation should be available in two forms—an equation relating the frequency of oscillation to the various parameters, and a set of equations for the sensitivity of this frequency to changes in each of the parameters.

Baxandall[1] has done this for the Pierce oscillator, under the assumption of a crystal of finite Q, in an ideal circuit. He considered the sensitivity to the crystal's static capacitance, the tuning capacitor, and the crystal's equivalent resistance. His results have been extended here in two ways:

1. To consider the combined effects of motional capacitance, shunt capacitance, other circuit capacitors, and the active element impedances;
2. To consider the Z-oscillator, as well as the Y-oscillator.

Expressions are derived for the normalized difference between the oscillation frequency and crystal series-resonant frequency, as well as the normalized difference between the oscillation frequency and crystal parallel resonant frequency. From these, the sensitivity of oscillation frequency to various parameters of the crystal and the circuit are derived. The conditions for which the expressions are valid are cited, as is the practical significance of the expressions. Experimental results are presented which verify the most useful equations.

2. Equivalent Crystal Network

We have defined a 'crystal network' by Fig. 1. The crystal itself is simulated by the conventional fundamental-mode model. In series with the crystal is C_3 (but for the Z-oscillator, $C_3 = \infty$, a short circuit). In parallel with this combination is C_c, which may be stray capacitance (e.g. C_{cb} of a transistor) for the Y-oscillator, or the tuning capacitor for the Z-oscillator.

The admittance of this crystal network is:

$$Y = s(C_c + C_0 \int C_3) \times \frac{s^2 + s\dfrac{R}{L} + \dfrac{1}{LC_m}\left[1 + \dfrac{C_m}{C_0 + C_c \int C_3}\right]}{s^2 + s\dfrac{R}{L} + \dfrac{1}{LC_m}\left[1 + \dfrac{C_m}{C_0 + C_3}\right]} \qquad (1)$$

where the symbol $\int$ denotes capacitors in series, so that, e.g.:

$$C_a \int C_b = \frac{C_a C_b}{C_a + C_b}$$

The roots of the numerator, in equation (1), locate the parallel resonant frequency of the network. The roots of the denominator locate the series resonant frequency.

These roots are:

$$s = -\frac{R}{2L} \pm \sqrt{\left(\frac{R}{2L}\right)^2 - \frac{1}{LC_m}\left[1 + \frac{C_m}{C_0 + C_c \int C_3}\right]} \qquad (2a)$$

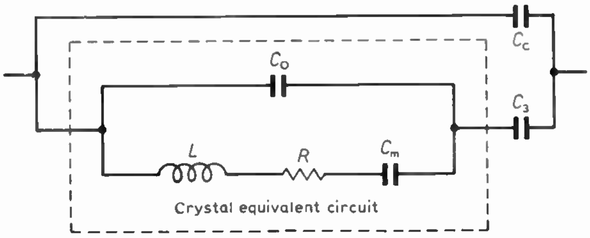

Fig. 1. Equivalent circuit of crystal network.

for the numerator, and

$$s = -\frac{R}{2L} \pm \sqrt{\left(\frac{R}{2L}\right)^2 - \frac{1}{LC_m}\left[1 + \frac{C_m}{C_0+C_3}\right]} \quad (2b)$$

for the denominator.

Therefore, for an infinite crystal Q (i.e. for $R \ll \sqrt{L/C_m}$) the effective series resonant angular frequency of the network (from equation (2b)) is:

$$\omega_s' = \frac{1}{\sqrt{LC_m}}\sqrt{1 + \frac{C_m}{C_0+C_3}} = \omega_s\sqrt{1 + \frac{C_m}{C_0+C_3}} \quad (3)$$

where ω_s is the series resonant angular frequency of the crystal itself (i.e. ω_s' for $C_3 = \infty$). For $C_m \ll (C_0+C_3)$, it is well known that the normalized frequency difference is

$$NFD_s = \frac{\omega_s' - \omega_s}{\omega_s} \simeq \frac{1}{2}\cdot\frac{C_m}{C_3+C_0} \quad (4)$$

Below is presented a more precise expression, which includes also the effects of the remainder of the oscillator circuit.

Similarly, with infinite Q, the effective parallel resonant angular frequency of the network is:

$$\omega_p' = \omega_s\left[1 + \frac{C_m}{C_0 + C_c \int C_3}\right] \quad (5)$$

Baxandall[1] (Table 1) has derived a more exact expression for (4) with finite Q, and with C_c absent. However, he did not include the secondary effects C_m, C_0 and C_3, nor the effects of other circuit elements, which are frequently critical. These are considered below.

3 Oscillation Frequency of Pierce Oscillator

The Pierce oscillator, such as the one shown in Fig. 2, is a popular special case of the Colpitts oscillator, in which the inductor is replaced by a crystal. The Colpitts oscillator is itself a special case of the type-Y oscillator[2] shown in Fig. 3, in which Y is the inductor, and Z_1 and Z_2 are capacitors C_1 and C_2, respectively.

A necessary condition for oscillation of the Y-oscillator is cited in equation (9.11a) of Ref. 2. From this, the normalized frequency differences are derived in Appendix 1. The results are as follows, where ω_0 is the angular frequency of oscillation, and

$$NFD_s = \frac{\omega_0 - \omega_s}{\omega_s}$$

$$NFD_s \simeq \frac{1}{2}\cdot\frac{C_m}{C_3+C_0}\left[1 + \left(1 + \frac{1}{a}\right)\frac{C_3}{C_0} \times \frac{C_0 \int C_3}{C_1 \int C_2 + C_c + C_0 \int C_3} - \frac{1}{4}\cdot\frac{C_m}{C_3+C_0}\right] \quad (6)$$

$$NFD_p = \frac{\omega_p - \omega_0}{\omega_p} \simeq \frac{1}{2}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0} \times \left[1 - \frac{C_0 \int C_3\left(1 - \frac{1}{2}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0}\left[1+2\frac{C_0}{C_3}\right] + \frac{1}{a}\right)}{C_1 \int C_2 + C_c + C_0 \int C_3} - \frac{1}{2}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0}\left(\frac{3}{2} + 2\frac{C_0}{C_3}\right)\right] \quad (7)$$

where

$$a = \frac{\omega_s^2}{\Delta_y}\left[C_1C_2 + (C_1+C_2)(C_c + C_0 \int C_3)\right] \gg 1 \text{ (usually)}$$

For the equations (6) and (7) to be valid, the following approximations must be applicable:

$Q \gg 1$ (i.e. $R \ll L/C_m$);

$C_1, C_2 > C_0 \gg C_m$;

$a \gg 1$ (i.e. $\Delta_y \ll f_0^2 C_1 C_2$);

$\mathrm{Re}\,(\Delta_y) \gg \mathrm{Im}\,(\Delta_y)$

From a practical point of view, the conditions are easily realizable for the great majority of applications. For example, from Baxandall we see that a typical CR-27 crystal with a Q of 20 000 will add about 0·25 parts per 10^6 to NFD_s. Also, typical values of C_1 and C_2 are in the hundreds of pF; C_0 is typically between 2 and 20 pF; and C_m is usually less than 1/100 of C_0. The condition for a is easily satisfied above about 1 MHz; at lower frequencies the choice of C_1, C_2, and the active element influences this approximation. In any case, good oscillator design dictates a large value for a, and a Δ_y which is approximately real at the frequency of operation (e.g. an r.f. transistor should be used for an r.f. oscillator).

These approximations introduce errors which are usually much smaller than 1 in 10^6. This is smaller than the usual calibrating tolerance of the crystal, and smaller than the errors introduced by differences in the assumed and the actual values of C_m, C_0 and C_3. Therefore, equations (6) and (7) are generally applicable with high Q crystals. For design of oscillators of relaxed precision, the equations may be simplified further (e.g. by neglecting the third term in the brackets [ ] in each

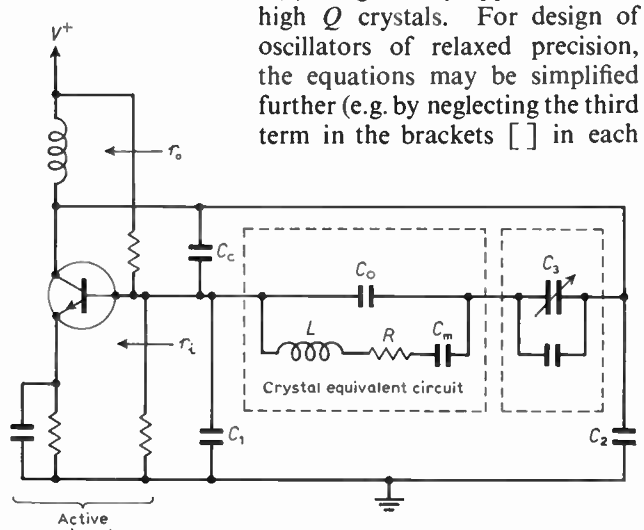

Fig. 2. Typical Pierce oscillator, with a transistor shown as the active element.

equation, by neglecting $1/a$ in the second term, by neglecting C_c when it is much less than C_2).

4. Frequency Sensitivity of Pierce Oscillator

A precise calculation of the frequency shift induced by a change in the value of any parameter is possible using one of the above equations. However, this procedure is often tedious, and offers no insight into the design of the oscillator. Consequently, we have derived sensitivity expressions for each parameter. These are valid for incremental changes in the parameter value, with the other parameters fixed.

A popular definition of sensitivity is:

$$S_u^{\omega_0} = \frac{u}{\omega_0}\frac{\omega d_0}{du} \tag{8}$$

Using this equation, the expressions below are derived (see Appendix 2) to correspond to equation (6).

$$S_{C_3}^{\omega_0} \simeq -\frac{1}{2}\cdot\frac{C_3C_m}{(C_3+C_0)^2}\times \left[1-2\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}\right] \tag{9a}$$

$$S_{C_m}^{\omega_0} \simeq -\frac{1}{2}\cdot\frac{C_0C_m}{(C_3+C_0)^2}\times \left[1-\left(\frac{C_3}{C_0}\cdot\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}\right)^2\right] \tag{10a}$$

$$S_{C_m}^{\omega_0} = \frac{NFD_s}{1+2(NFD_s)} \simeq NFD_s-2(NFD_s)^2 \tag{11a}$$

$$S_{C_m}^{\omega_0} = \frac{1}{2}\cdot\frac{C_m}{C_3+C_0}\left[1+\left(1+\frac{1}{a}\right)\frac{C_3}{C_0}\times \frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}-\frac{5}{4}\cdot\frac{C_m}{C_3+C_0}\right] \tag{11b}$$

$$S_{C_1\int C_2}^{\omega_0} = \frac{1}{2}\cdot\frac{C_3^2C_m(C_1\int C_2)}{(C_3+C_0)^2(C_1\int C_2+C_c+C_0\int C_3)^2} \tag{12}$$

$$S_{C_c}^{\omega_0} = -\frac{1}{2}\cdot\frac{C_3^2C_mC_c}{(C_3+C_0)^2(C_1\int C_2+C_c+C_0\int C_3)^2} \tag{13}$$

It is often sufficient to use equations (9a), (10a) and (11a and b) in the following simplified forms:

$$S_{C_3}^{\omega_0} = -\frac{1}{2}\cdot\frac{C_3C_m}{(C_3+C_0)^2} \tag{9b}$$

$$S_{C_0}^{\omega_0} = -\frac{1}{2}\cdot\frac{C_0C_m}{(C_3+C_0)^2} \tag{10b}$$

$$S_{C_m}^{\omega_0} = NFD_s = \frac{1}{2}\cdot\frac{C_m}{C_3+C_0} \tag{11c}$$

What are the implications of equations (9) through (13)? C_3 usually includes a variable capacitor, to permit adjustment of the oscillator frequency. Therefore, it is desired that $-S_{C3}^{\omega_0}$ (equation (9)) will be sufficiently *large* so that the frequency adjustment range will cover the possible variations in ω_0 due to variations in the crystal frequency and in the other parameters. At the same time, the other sensitivity magnitudes should be as *small* as possible, to reduce their contributions to initial frequency deviation and to frequency instability. This leads to the following conclusions:

1. A smaller value of C_3 increases $-S_{C3}^{\omega_0}$, but it also increases all other sensitivities. Therefore, C_3 cannot be made arbitrarily small. Instead, it should be chosen only as small as is required for the adjustment range—after the values and error contributions of the other parameters have been determined.
2. The crystal capacitances, C_m and C_0, should be as small as possible. (C_m and C_0 are usually in a fixed approximate ratio, which is different for each crystal type).
3. C_1 and C_2 should be as large as possible, without overdriving the crystal.
4. C_c should be as small as possible.
5. For many designs, it is possible to make the sensitivities to $C_1\int C_2$ and C_2 negligible with respect to the sensitivities to C_3, C_0 and C_m. This reduces the adjustment and stability considerations to those of the crystal network only.

From equation (8), the effective normalized change in frequency is:

$$\left.\frac{\Delta\omega_0}{\omega_0}\right|_u = S_u^{\omega_0}\left(\frac{\Delta u}{u}\right) \tag{14}$$

Then, for example:

$$\left.\frac{\Delta\omega_0}{\omega_0}\right|_{C_3} \simeq -\frac{1}{2}\cdot\frac{C_m}{(C_3+C_0)^2}\cdot\Delta C_3 \tag{15}$$

$$\left.\frac{\Delta\omega_0}{\omega_0}\right|_{C_0} \simeq -\frac{1}{2}\cdot\frac{C_m}{(C_3+C_0)^2}\cdot\Delta C_0 \tag{16}$$

$$\left.\frac{\Delta\omega_0}{\omega_0}\right|_{C_m} \simeq \frac{1}{2}\cdot\frac{\Delta C_m}{C_3+C_0} \tag{17}$$

$$\left.\frac{\Delta\omega_0}{\omega_0}\right|_{C_3,\,C_0,\,C_m} \simeq \frac{1}{2}\cdot\frac{C_m}{C_3+C_0}\left[\frac{\Delta C_m}{C_m}-\frac{\Delta C_3-\Delta C_0}{C_3+C_0}\right] \tag{18}$$

Equations (15) and (16) prove that the frequency change induced by ΔC_3 is the same as frequency change induced by an equal value of ΔC_0.

Equation (15) indicates the required temperature coefficient and long-term stability for C_3—for a minimum contribution of the trimmer capacitor to frequency instability.

Equations (16) and (17) show the desirability of having values of C_m and C_0 which deviate as little as possible, so as to reduce the adjustment range which C_3 must provide. Indeed, careful preparation and maintenance of the crystal manufacturing masks are desired to achieve minimum values of

$$\frac{\Delta C_m}{C_m} \quad\text{and}\quad \frac{\Delta C_0}{C_0}$$

Equation (18) may also be used in determining the range of C_3 which compensates for ΔC_m and ΔC_0.

5 Oscillation Frequency ot Z-Oscillator

Compare the Y-type oscillator of Fig. 3, discussed above with the Z-type oscillator[2] of Fig. 4. If Z is

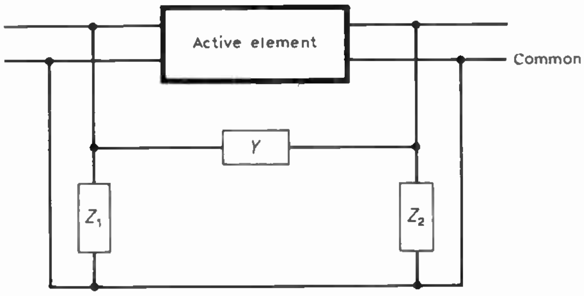

Fig. 3. Y-type oscillator.

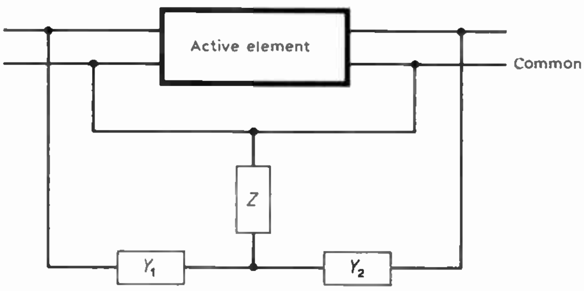

Fig. 4. Z-type oscillator.

replaced by a crystal network, and Y_1 and Y_2 are capacitors C_1 and C_2, respectively, an oscillator such as that shown in Fig. 5 results. The Miller oscillator[1] is an example of this configuration.

From the condition for oscillation of equation (9.13a) of Ref. 2, the normalized frequency differences are derived, as described in Appendix 3.

The results are as follows:

$$NFD_s \simeq \frac{1}{2}\cdot\frac{C_m}{C_4+C_0}\times \left[1+\frac{1}{\omega_s^2\Delta_z(C_1 \int C_2)([C_4+C_0]\int[C_1+C_2])}-\frac{1}{4}\cdot\frac{C_m}{C_4+C_0}\right] \quad (19)$$

$$NFD_p \simeq \frac{1}{2}\cdot\frac{C_4}{C_0}\cdot\frac{C_m}{C_4+C_0}\left[1-\left(\frac{C_m}{C_4+C_0}\right)\times \times\frac{1-C_m/C_0}{\omega_s^2\Delta_z(C_1 \int C_2)([C_4+C_0]\int[C_1+C_2])}-\frac{1}{2}\cdot\frac{C_4}{C_0}\cdot\frac{C_m}{C_4+C_0}\left(\frac{3}{2}+2\frac{C_0}{C_4}\right)\right] \quad (20)$$

where Δ_z is the impedance matrix of the active element (see Ref. 2).

The approximations in this case are as follows:

$$Q \gg 1$$

$$C_1, C_2 > C_0 \gg C_m$$

$$\Delta_z \gg \frac{1}{f_0^2 C_1 C_2}\left(1+\frac{C_1+C_2}{C_0+C_4}\right)$$

$$\mathrm{Re}(\Delta_z) \gg \mathrm{Im}(\Delta_z)$$

The form of equations (19) and (20) is similar to the form of equations (6) and (7), except that C_4 fulfills the adjustment function for the Z-oscillator which was fulfilled by C_3 in the Y-oscillator. The discussion which follows equations (6) and (7) is also generally applicable to the Z-oscillator, and equations (19) and (20).

6 Frequency Sensitivity of Z-oscillator

Using the definition of sensitivity of equation (8), the sensitivity equations corresponding to equations (19) are derived as follows:

$$S^{\omega_0}_{(C_4+C_0)} \simeq -\frac{NFD_s+\dfrac{C_m}{\omega_s^2\Delta_z(C_1\int C_2)(C_4+C_0)^2}}{1+2(NFD_s)} \simeq -\frac{1}{2}\cdot\frac{C_m}{C_4+C_0}\times \times\left[1+\frac{1}{\omega_s^2\Delta_z(C_1\int C_2)([\frac{1}{2}(C_4+C_0)]\int[C_1+C_2])}\right] \quad (21)$$

$$S^{\omega_0}_{C_4} = \frac{C_4}{C_4+C_0}\cdot S^{\omega_0}_{(C_4+C_0)} \quad (22a)$$

$$S^{\omega_0}_{C_0} = \frac{C_0}{C_4+C_0}\cdot S^{\omega_0}_{(C_4+C_0)} \quad (23a)$$

$$S^{\omega_0}_{Cm} = \frac{NFD_s}{1+2(NFD_s)} \simeq NFD_s-2(NFD_s)^2 \quad (24a)$$

$$S^{\omega_0}_{C_m} \simeq \frac{1}{2}\cdot\frac{C_m}{C_4+C_0}\times \left[1+\frac{1}{\omega_s^2\Delta_z(C_1\int C_2)([C_4+C_0]\int[C_1+C_2])}-\frac{5}{4}\cdot\frac{C_m}{C_4+C_0}\right] \quad (24b)$$

$$S^{\omega_0}_{C_1} \simeq -\frac{1}{2}\cdot\frac{C_m}{C_4+C_0}\cdot\frac{1}{\omega_s^2\Delta_z C_1([C_4+C_0]\int C_2)} \quad (25)$$

$$\omega^{\omega_0}_{C_2} \simeq -\frac{1}{2}\cdot\frac{C_m}{C_4+C_0}\cdot\frac{1}{\omega_s^2\Delta_z C_2([C_4+C_0]\int C_1)} \quad (26)$$

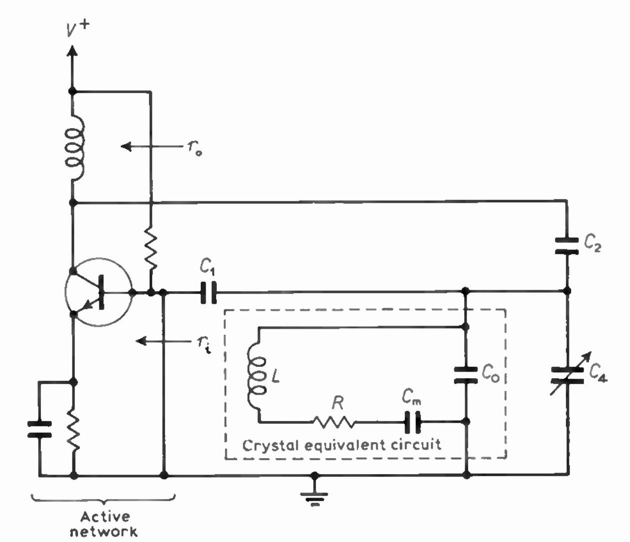

Fig. 5. A realization of the Z-oscillator, with a transistor shown as the active element.

Usually equations (22a), (23a) and (24a) may be approximated in the following simplified forms:

$$S_{C_4}^{\omega_0} \simeq \frac{C_4}{C_4+C_0}(NFD_s) \simeq -\frac{1}{2}\cdot\frac{C_4C_m}{(C_4+C_0)^2} \quad \text{(22b)}$$

$$S_{C_0}^{\omega_0} \simeq \frac{C_0}{C_4+C_0}(NFD_s) \simeq -\frac{1}{2}\cdot\frac{C_0C_m}{(C_4+C_0)^2} \quad \text{(23b)}$$

$$S_{C_m}^{\omega_0} \simeq NFD_s \simeq \frac{1}{2}\cdot\frac{C_m}{C_4+C_0} \quad \text{(24c)}$$

The implications of equations (21) through (26) are the same as those of equations (9) through (13), except that C_4 should be chosen according to the same criteria by which C_3 is chosen for the Pierce oscillator. In this case, the 'crystal network' consists of the crystal and C_4, in parallel.

7 Experimental Results

Experimental measurements verified the most useful expressions for the frequency and sensitivity of the Pierce oscillator. Examples of the results are cited below.

Five crystals of different frequencies were tested on the CI-meter, where the influence of C_1 and C_2 are negligible. The crystal series resonant frequency and the oscillation frequency with a 32-pF series capacitor were measured at 25°C. The results are shown in Table 1.

Table 1

Normalized frequency difference (NFD_s)

Measured (with C_0 = 32 pF) and calculated (from equation (6))

f_s (MHz) $C_3 = 0$	$f_0 - f_s$ (Hz)	C_m (pF×10^{-3})	C_0 (pF)	NFD_s (parts in 10^6) Measured	Calculated
3·598 988	930	18·6	3·95	258	258·5
3·599 160	798	15·8	3·7	222	221·5
6·395 830	2220	26	5·3	347	348
10·245 660	3670	27	5·5	358	360
19·595 205	5470	20·2	4·35	279	278

There is excellent agreement between the measured and calculated values of NFD_s. The first two crystals, being of about the same frequency, can be used also to verify equation (11c) for the sensitivity to C_m. The value of

$$\left.\frac{\Delta\omega_0}{\omega_0}\right|_{C_m}$$

calculated from equation (17), is 39 parts in 10^6. The equivalent measured shift in oscillating frequency (after removing the initial frequency difference) is

$$(f_0-f_s)_1-(f_0-f_s)_2 = 930-798 = 132 \text{ Hz}$$

from which the normalized shift is

$$\frac{132 \text{ Hz}}{3\cdot6 \text{ MHz}} = 36\cdot7 \text{ parts in } 10^6$$

These two results are within the experimental accuracy. Note also that the difference in f_s of these two crystals is 172 Hz (or 47·8 in 10^6)—so that the influence of C_m on f_0 is of the same order of magnitude as the influence of f_s!

It was determined in an oscillator that the addition (or deletion) of a fixed value of ΔC to C_0 has the same effect as adding this ΔC to C_3, thus confirming the relationship between equations (15) and (16).

To determine whether the C_3 (or C_0) sensitivity is valid, even for large variations in C_3 (or C_0), a trimmer capacitor with a value of 0·8 pF to 10 pF (i.e. 5·4 pF ±4·6 pF) was varied over its entire range. Table 2 shows the results (with the crystal at 75°C), where f_0 corresponds to the condition of minimum C_3, and Δf is the frequency shift introduced by adjusting C_3 to its maximum value.

Table 2

Frequency sensitivity to C_3, over trim range

f_0 Initial frequency (MHz)	Δf Frequency shift (Hz)	Crystal capacitance C_m (nF)	C_0 (pF)	$\frac{\Delta f}{f_0}$ (parts in 10^6) Measured	Calculated
3·600 068	−202	18·6	3·95	−56·1	−58·6
3·600 124	−178	15·8	3·7	−49·5	−50·4
6·398 298	−480	26	5·3	−75·0	−76·7
10·249 590	−790	27	5·5	−77·2	−78·6
19·600 888	−1275	20·2	4·35	−65·0	−62·4

In calculating $\Delta f/f_0$, using equation (15), the initial value of C_3 is taken as 29·6 pF—consisting of a fixed 26·8-pF capacitor, the 0·8-pF minimum trimmer value, and 2-pF stray capacitance. The measured and calculated results agree, within the experimental accuracy—even considering the large variation in C_3.

The sensitivities to $C_1 \int C_2$ (equation (12)) and C_c (equation (13)) were examined in a 20-MHz oscillator, with C_1 = 560 pF, C_2 = 240 pF, and C_c estimated at about 1 pF. Within measurement accuracy, the results (e.g. in Table 3) agree with the calculated values.

Table 3

Sensitivity to $C_1 \int C_2$ and C_c

Capacitor variations ΔC_1 (pF)	ΔC_2 (pF)	ΔC_3 (pF)	Frequency shift (Hz)	$\frac{\Delta f}{f_0}$ (parts in 10^6) Measured	Calculated
5·6	2·2	0	−1·5	−0·375	−0·43
0	0	0·3	−7·5	−0·075	−0·078

8 Conclusions

Equations have been presented for the frequency of oscillation of a crystal oscillator as a function of both the crystal parameters and the circuit parameters. For maximum resolution and ease of calculation, these were in the form of normalized frequency difference between the oscillator frequency and the crystal series (or parallel) resonant frequency. The expressions for the frequency sensitivity, as a function of each parameter, were presented in a more precise form than has previously been published, and for some parameters that were previously not treated. This was done for the Y-oscillator (i.e. Pierce oscillator) and the Z-oscillator. Experimental results, for the Pierce oscillator, were cited to verify the theoretical expressions.

9 References

1. Baxandall, P. J., 'Transistor crystal oscillators and the design of a 1-Mc/s oscillator circuit capable of good frequency stability', *The Radio and Electronic Engineer*, 29, No. 4, pp. 229–46, April 1965.
2. Cote, A. J. and Oakes, J. B., 'Linear Vacuum-Tube and Transistor Circuits', Chap. 9 (McGraw-Hill, New York, 1961).

10 Appendix 1: Derivation of *NFD*

A condition for oscillations, in the circuit of Fig. 3, is (equation (9.11a) of Reference 2):

$$\frac{X}{B}\Delta_y = \frac{1}{nXB} - \frac{1+n}{n} \qquad (27)$$

where X is the reactive part of Z_1, B is the susceptance of Y, $\Delta_y = y_i y_0 - y_r y_f$, and $n = Z_2/Z_1$. For the Pierce oscillator (see Fig. 2):

$$Z = jX = -\frac{j}{\omega C_1}, \quad Y = jB \quad \text{and} \quad n = \frac{C_1}{C_2}$$

Taking equation (1) of this paper for Y, with infinite Q (i.e. $R = 0$), we find that

$$B = \omega(C_c + C_0 \int C_3)\cdot\frac{\omega^2 - \dfrac{1}{LC_m}\left[1 + \dfrac{C_m}{C_3 + C_c \int C_3}\right]}{\omega^2 - \dfrac{1}{LC_m}\left[1 + \dfrac{C_m}{C_0 + C_3}\right]}$$

From equations (3) and (5):

$$B \simeq \omega(C_c + C_0 \int C_3)\frac{\omega^2 - \omega_p'^2}{\omega^2 - \omega_s'^2} \qquad (28)$$

Rearranging (27):

$$B + \frac{nX}{1+n}\Delta_y - \frac{1}{(1+n)X} = 0 \qquad (29)$$

But,

$$nX = -\frac{1}{\omega C_2}, \quad \frac{nX}{1+n} = -\frac{1}{\omega(C_1 + C_2)}$$

and

$$\frac{1}{(1+n)X} = -\omega(C_1 \int C_2) \qquad (30)$$

Substituting (28) and (30) into (29):

$$\omega(C_c + C_0 \int C_3)\frac{\omega^2 - \omega_p'^2}{\omega^2 - \omega_s'^2} - \frac{\Delta_y}{\omega(C_1 + C_2)} + \omega(C_1 \int C_2) \simeq 0$$

Multiplying by $\omega(\omega - \omega_s'^2)$, if $\omega \neq \omega_s'$:

$$\omega^2(\omega^2 - \omega_p'^2)(C_c + C_0 \int C_3) + \omega^2(\omega^2 - \omega_s'^2)(C_1 \int C_2) - \frac{\Delta_y}{C_1 + C_2}(\omega^2 - \omega_s'^2) \simeq 0$$

$$\omega^4[C_c + C_0 \int C_3 + C_1 \int C_2] - \omega^2\left[\omega_p'^2(C_c + C_0 \int C_3) + \omega_s'^2(C_1 \int C_2) + \frac{\Delta_y}{C_1 + C_2}\right] + \frac{\Delta_y}{C_1 + C_2}\omega_s'^2 \simeq 0 \quad (31)$$

From (5):

$$\omega_p'^2 \simeq \omega_s^2\left[1 + \frac{C_m}{C_0 + C_c \int C_3}\right] = \omega_s^2\left[\frac{C_c + C_0 \int C_3 + C_m\cdot(C_3 + C_c)/(C_3 + C_0)}{C_c + C_0 \int C_3}\right]$$

Substituting this and (3) into (31):

$$\omega^4[C_1 \int C_2 + C_c + C_0 \int C_3] - \omega^2\omega_s^2\left[\left(C_c + C_0 \int C_3 + C_m\cdot\frac{C_3 + C_c}{C_3 + C_0}\right) + \left(C_1 \int C_2 + C_m\cdot\frac{C_1 \int C_2}{C_3 + C_0}\right) + \frac{\Delta_y}{\omega_s^2(C_1 + C_2)}\right] + \omega_s^2\cdot\frac{\Delta_y}{C_1 + C_2}\left[1 + \frac{C_m}{C_3 + C_0}\right] \simeq 0$$

Let $x = (\omega/\omega_s)^2$. Then, dividing by

$$\omega_s^4[C_1 \int C_2 + C_c + C_0 \int C_3]$$

yields:

$$x^2 - x\left[1 + \frac{C_m}{C_3 + C_0}\cdot\frac{C_1 \int C_2 + C_c + C_3}{C_1 \int C_2 + C_c + C_0 \int C_3} + \frac{1}{a}\right] + \frac{1}{a}\left[1 + \frac{C_m}{C_3 + C_0}\right] \simeq 0$$

where

$$a = \frac{\omega_s^2}{\Delta_y}(C_1 + C_2)(C_1 \int C_2 + C_c + C_0 \int C_3).$$

We solve by using the biquadratic root formula and the approximation:

$$\sqrt{1+k} \simeq 1 + \frac{k}{2} - \frac{k^2}{8}, \quad \text{for } k \ll 1 \qquad (32)$$

Then:

$$x_0 \simeq 1 + \frac{C_m}{C_3 + C_0} \times \left[1 + \left(1 + \frac{1}{a}\right)\frac{C_3^2}{(C_3 + C_0)(C_1 \int C_2 + C_c + C_0 \int C_3)}\right] \qquad (33)$$

if the following approximations are valid:

$$C_1, C_2 > C_0 \gg C_m$$

$$a \gg 1$$

$$\mathrm{Re}\,(\Delta_y) \gg \mathrm{Im}\,(\Delta_y)$$

Then:

$$NFD_s = \frac{\omega_0 - \omega_s}{\omega_s} = \sqrt{x_0} - 1$$

Using equation (33):

$$NFD_s \simeq \frac{1}{2}\cdot\frac{C_m}{C_3 + C_0}\left[1 + \left(1 + \frac{1}{a}\right)\frac{C_3}{C_0} \times \frac{C_0 \int C_3}{C_1 \int C_2 + C_c + C_0 \int C_3} - \frac{1}{4}\cdot\frac{C_m}{C_3 + C_0}\right]$$

which is equation (6).

To determine NFD_p, we use the relationship:

$$\left(\frac{\omega_p}{\omega_s}\right)^2 = 1 + \frac{C_m}{C_0} = \frac{C_0 + C_m}{C_0}$$

Then,

$$\left(\frac{\omega_0}{\omega_p}\right)^2 = \left(\frac{\omega_0}{\omega_s}\right)^2\left(\frac{\omega_s}{\omega_p}\right)^2 = x_0\cdot\frac{C_0}{C_0 + C_m} = \frac{x_0}{1 + C_m/C_0}$$

But,

$$\frac{1}{1+k} \simeq 1 - k + k^2, \quad \text{for } k \ll 1$$

Therefore, when the above approximations apply,

$$\left(\frac{\omega_0}{\omega_p}\right)^2 \simeq \left[1-\frac{C_m}{C_0}+\left(\frac{C_m}{C_0}\right)^2\right]\cdot\left[1+\frac{C_m}{C_3+C_0}+\left(1+\frac{1}{a}\right)\left(\frac{C_3}{C_3+C_0}\right)^2\cdot\frac{C_m}{C_1\int C_2+C_c+C_0\int C_3}\right]$$

$$\simeq 1-\left[\frac{C_m}{C_0}-\frac{C_m}{C_3+C_0}\right]+\left[\left(\frac{C_m}{C_0}\right)^2-\frac{C_m^2}{C_0(C_3+C_0)}\right]+$$

$$+\left[1-\frac{C_m}{C_0}\right]\left(1+\frac{1}{a}\right)\left(\frac{C_3}{C_3+C_0}\right)^2\cdot\frac{C_m}{C_1\int C_2+C_c+C_0\int C_3}+\ldots$$

$$\simeq 1-\frac{C_3C_m}{C_0(C_3+C_0)}\cdot\left[1-\frac{C_m}{C_0}\right]\cdot\left[1-\left(1+\frac{1}{a}\right)\cdot\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}\right]$$

But:

$$NFD_p=\frac{\omega_p-\omega_0}{\omega_p}=1-\sqrt{\left(\frac{\omega_0}{\omega_p}\right)^2}$$

Using equation (32):

$$NFD_p\simeq\frac{1}{2}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0}\left[\left(1-\frac{C_m}{C_0}\right)\left\{1-\left(1+\frac{1}{a}\right)\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}\right\}+\right.$$

$$\left.+\frac{1}{4}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0}\left(1-2\frac{C_m}{C_0}+\ldots\right)\left\{1-2\left(1+\frac{1}{a}\right)\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}+\ldots\right\}\right]$$

$$\simeq\frac{1}{2}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0}\left[1-\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}\cdot\left(1+\frac{1}{a}\right)\left(1-\frac{C_m}{C_0}+\frac{1}{2}\ \frac{C_3C_m}{C_0[C_3+C_0]}\right)-\right.$$

$$\left.-\frac{C_m}{C_0}\left(1-\frac{1}{4}\cdot\frac{C_3}{C_3+C_0}+\ldots\right)\right] \quad (34)$$

Since

$$\left(1+\frac{1}{a}\right)\left(1-\frac{C_m}{C_0}+\frac{1}{2}\cdot\frac{C_3C_m}{C_0[C_3+C_0]}\right)\simeq 1-\frac{1}{2}\cdot\frac{C_3}{C_0}\cdot\frac{C_m}{C_3+C_0}\left[1+2\frac{C_0}{C_3}\right]+\frac{1}{a}$$

therefore (34) is essentially equation (7).

11 Appendix 2: Derivation of Sensitivity Expressions

From equation (8), it can be shown that:

$$S_u^{\omega_0^2}=2S_u^{\omega_0}\quad\text{and}\quad S_u^{\omega_0/\omega_s}=S_u^{\omega_0}$$

Therefore:

$$S_u^{\omega_0}=\frac{1}{2}S_u^{x_0}=\frac{1}{2}\frac{u}{x_0}\frac{dx_0}{du}$$

For example, from equation (33):

$$\frac{dx_0}{dC_m}\simeq\frac{1}{C_3+C_0}\left[1+\left(1+\frac{1}{a}\right)\cdot\frac{C_3^2}{(C_3+C_0)(C_1\int C_2+C_c+C_0\int C_3)}\right]$$

and

$$S_{C_m}^{\omega_0}\simeq\frac{\dfrac{1}{2}\cdot\dfrac{C_m}{C_3+C_0}\left[1+\left(1+\dfrac{1}{a}\right)\cdot\dfrac{C_3^2}{(C_3+C_0)(C_1\int C_2+C_c+C_0\int C_3)}\right]}{1\quad\dfrac{C_m}{C_3+C_0}\left[1+\left(1+\dfrac{1}{a}\right)\cdot\dfrac{C_3^2}{(C_3+C_0)(C_1\int C_2+C_c+C_0\int C_3)}\right]}$$

$$\simeq\frac{NFD_s}{1+2(NFD_s)}\simeq NFD_s-2(NFD_s)^2\quad\text{(see 11a)}$$

$$S_{C_m}^{\omega_0}\simeq\frac{1}{2}\cdot\frac{C_m}{C_3+C_0}\left[1+\left(1+\frac{1}{a}\right)\frac{C_3}{C_0}\cdot\frac{C_0\int C_3}{C_1\int C_2+C_c+C_0\int C_3}-\frac{5}{4}\cdot\frac{C_m}{C_3+C_0}\right]\quad\text{(see 11b)}$$

Equations (9), (10), (12) and (13) are derived, in a similar manner, from (33).

12 Appendix 3: Derivations for Z-type Oscillator

The derivation of NFD_s and NFD_p for the Z-oscillator—equations (19) and (20)—are similar to the derivations of Appendix 1, except for the following differences:

1. The circuit is represented by Fig. 4.
2. C_3 is ∞ (i.e. a short-circuit).
3. C_c is replaced by C_4.
4. The condition for oscillation is taken from (9.13a) of Ref. 2:

$$\frac{B}{X}\Delta_z = \frac{1}{mXB} - \frac{1+m}{m}$$

5. $$Y = jB = j\omega C_1, \qquad m = \frac{Y_2}{Y_1} = \frac{C_2}{C_1}$$

$$Z = jX \simeq -j\frac{1}{\omega(C_4+C_0)}\cdot\frac{\omega^2-\omega_s^2}{\omega^2-\omega_p'^2}$$

$$\omega_p'^2 \simeq \omega_s^2\left[1+\frac{C_m}{C_4+C_0}\right]$$

The sensitivity expressions—(21) to (26)—are then derived from $(\omega_0/\omega_s)^2$, as in Appendix 2.

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Standard Frequency Transmissions—January 1976

(Communication from the National Physical Laboratory)

January 1976	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)	
	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	−0·2	697·9	612·1
2	−0·2	698·0	611·9
3	−0·2	698·7	611·9
4	−0·1	698·1	611·7
5	−0·1	698·3	612·1
6	−0·1	698·1	612·2
7	−0·2	699·5	612·2
8	−0·2	699·8	612·2
9	−0·2	699·9	612·0
10	−0·2	699·3	612.0
11	−0·2	697·9	612·0
12	−0·2	698·5	611·7
13	−0·2	699·1	612·0
14	−0·2	698·9	612·4
15	−0·2	697·2	612·2
16	−0·2	700·2	612·2
17	−0·2	700·2	612·2
18	−0·3	699·8	612·3
19	−0·2	698·1	612·3
20	0	698·7	612·7
21	0	698·3	612·6
22	0	698·5	612·8
23	0	698·3	612·8
24	0	698·4	612·8
25	0	697·6	613·0
26	−0·1	700·1	613·1
27	−0·2	699·5	613·0
28	−0·2	700·3	613·2
29	−0·2	700·1	613·0
30	−0·2	701·1	613.1
31	−0·2	700·4	612·9

All measurements in terms of H-P Caesium Standard No. 344, agrees with the NPL Caesium Standard to 1 part in 10^{11}.

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

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

UDC 621.365.55.029.55:676.677.3

Radio-frequency drying of non-metallic materials

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Based on a paper presented at the Components and Circuits Group Colloquium on 'H.F. Heating—Circuits and Techniques' held in London on 9th October 1974.

SUMMARY

Hot air drying suffers from two serious drawbacks. On the one hand, it is very non-uniform even under well-controlled conditions, on the other, the rate falls off sharply towards the end of the drying period. Often the net result is a product which is unevenly dried at poor energy efficiency. Radio-frequency power can remedy both of these shortcomings. The advantages of radio-frequency power in drying from the point of view of product quality and power costs are illustrated by reference to results obtained when drying paper and wool.

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1 Introduction

It is obvious that drying is a very important industrial process of very long standing. What is perhaps not so obvious is that, in spite of the long period over which man has been drying materials, there are still serious problems to be overcome. There are essentially two stages to any drying process. Initially, when the material is wet, evaporation will occur from its surface at a constant rate for conditions of constant air flow. Eventually, however, a point is reached when water migration from the internal structure of the material begins to limit the evaporation rate, which progressively drops with reducing moisture content.

How problems can arise in both the constant and falling rate periods is best seen by taking a specific example. Wool is usually squeezed or spun to a moisture content of about 50%, from which point it is dried in a hot air dryer. The first problem occurs at the wet end, where channelling of the air occurs through the drier, less matted areas. This results in reduced drying in the wet regions and, therefore, uneven drying. Secondly, because of the mechanism described above, the drying rate falls off rapidly below about 35% (the desired final value being about 20%) which leads to very lengthy, inefficient dry end sections.

The problems of unevenness and lengthy dry end sections are common to a wide range of materials, and can result in considerable costs. The contribution of radio-frequency power lies in its ability to heat rapidly throughout the volume independently of the vagaries of convective and conductive heat transfer, and in its property of coupling selectively into the wetter areas thereby promoting uniformity.

Studies on the radio-frequency drying of paper[1–6] and wool[7] illustrate the magnitude of the advantages to be gained.

2 Paper Drying

The manufacture of paper starts with a wet slurry which, after draining and pressing, emerges as a web containing two parts of water to one of solid. From this point on, the web is dried by evaporation on huge steam-heated cylinders, maybe a hundred, each $1\frac{1}{2}$ m in diameter. Ideally the product leaves the machine at $7\% \pm \frac{1}{2}\%$; 7% being the equilibrium moisture content in contact with the atmosphere and the close tolerance being required to prevent distortion during storage or use. In practice, this specification is never achieved directly. Uniformity is the most important problem and this is usually achieved by drying down to 4%—this is very costly in energy because of the extreme inefficiency of the dry end and it loses revenue because paper is sold by weight. It is possible to rewet the paper to 7% but this is rarely done because wetting is not much more controllable than drying.

The rate at which energy is released by the mechanism of dielectric loss is given by

$$P = \varepsilon_0 \varepsilon_r'' \omega E^2 \quad \mathrm{W/m^3} \qquad (1)$$

where ε_0 = permittivity of free space,

ε'' = loss factor,

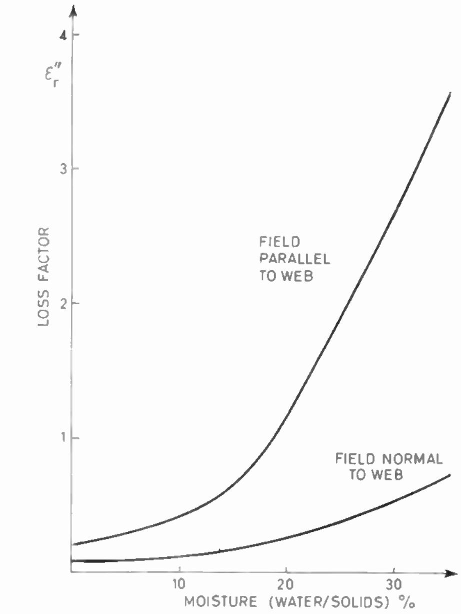

Fig. 1. Loss factor of paper at 27·12 MHz.

ω = angular velocity of field, s^{-1}

E = volt/m.

Figure 1, taken from the work of Driscoll,[8] shows the loss factor of paper as a function of moisture content at 27·12 MHz with the field orientated both perpendicular to and parallel to the plane of the web. Clearly, the absolute value of loss factor and its variation with moisture content are greater for the case where the field is in the plane of the paper—consequently any applicator aiming at producing efficient coupling and good moisture levelling should employ this field orientation. Looking at the shape of the loss factor curve more closely, the loss factor falls very rapidly with moisture content, the loss factor at the desired outlet value of 7% being

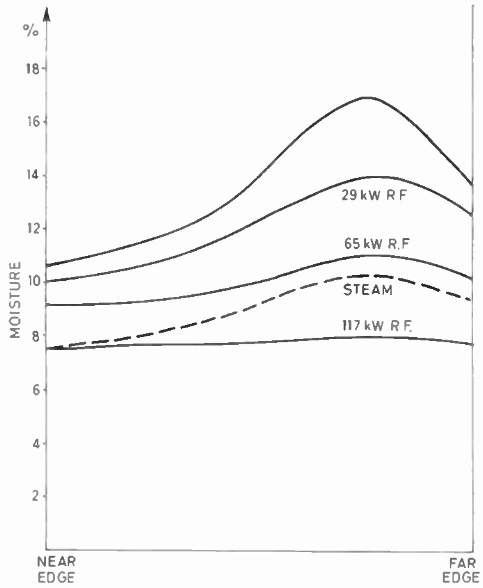

Fig. 2. Comparison of r.f. and steam cylinder drying—moisture profile across web of paper on machine (1·5 tonne/h).

only about one-third the value at 15%. In other words, as can be seen from equation (1), three times the power will be liberated per unit volume in regions containing 15% than in those containing 7%. One would anticipate, therefore, that r.f. would have a very powerful levelling effect.

Characteristically, paper machines run with the centre wetter than the edges; for example, at the point along the machine where the mean value is 14% the centre may be 17% and the edge 11%. Figure 2 compares r.f. and steam heated cylinder drying. The top moisture profile taken transversely to the motion of the web is the typical hump shape encountered in practice. As the amount of r.f. power (quoted for $1\frac{1}{2}$ tonne/h solids) is increased the moisture profile improves considerably. At an edge value of $7\frac{1}{2}$% the deviation in the centre is only about 0·2%, whereas the steam cylinders exhibit a difference of 2% between peak and trough. The type of applicator used to obtain the results from which these results were deduced is shown in Fig. 3. The electric field is contrived to be in the plane of the web in order to exploit the anisotropy of the loss factor. It is estimated that about 8% of the product of paper mills is wasted because of uneven drying—this could be entirely eliminated by the use of radio-frequency power. Assuming that product losses are only halved by the use of radio-frequency fields, one calculates a pay-back time

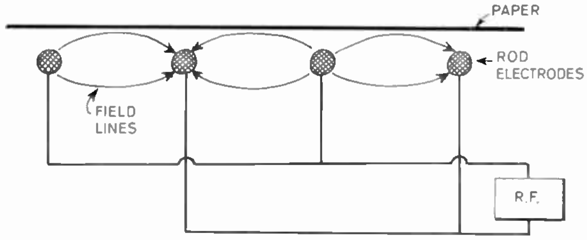

Fig. 3. Electrode configuration for paper drying at radio frequency.

of the order of one year, allowing for electrical energy consumption. In fact, the savings are very much greater because drying is so inefficient at the dry end of a paper machine that about 30% of the length of the machine is used to dry from about 14% (i.e. to remove only about 5% of the total water present)—replacement of the last 10 m of the machine by 2 m of r.f. would save a great deal of energy in terms of motive power and wasted steam heating. It is estimated[9] that, in terms of equivalent coal burnt at the power station and in the paper mill's boiler, r.f. can save about 25% of the total primary energy requirements for drying paper.

A number of installations have already been made and new orders are being placed as the paper industry begins to appreciate the advantages that are to be gained.

3 Wool Sliver Drying

The loss factor of wet wool at 27·12 MHz is shown in Fig. 4.[10] As in the case of paper, there is a strong anisotropy, the absolute value and its rate of change with moisture content being greatest when the field is

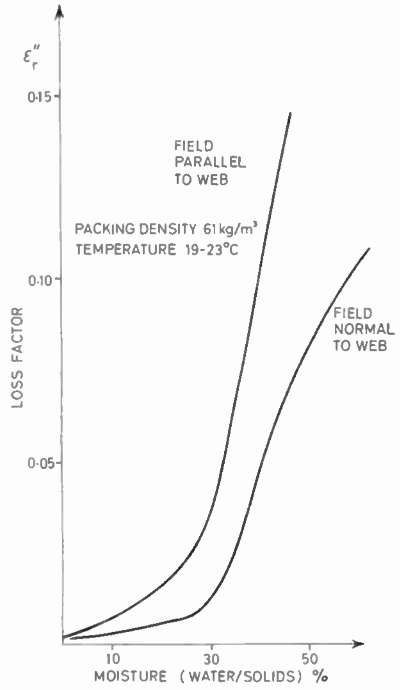

Fig. 4. Loss factor of white wool at 27·12 MHz.

aligned with the fibre—this indicates that the applicator must be designed to produce this optimum field configuration. Work at ECRC[7] was directed at the drying of wool in the form of sliver, which is a loose rope comprised of combed wool with the fibre orientation tending to be axial. Fig. 5 shows a schematic representation of the system used. The sliver was passed both inside and outside electrodes in the form of elongated loops at alternately positive and negative potentials. In this way the field was orientated along the fibre and the loaded Q of the circuit made small to increase efficiency. As would be expected, drying rates were considerably greater than those obtained using hot-air dryers, moreover, the levelling achieved was outstanding. Characteristically hot-air drying can give outlet moisture contents of $18 \pm 4\%$, plus some awkward, very wet patches that lead to problems in dyeing and combing operations. Figure 6 shows the effect of putting a radio-frequency dryer after a conventional, hot-air drum dryer. The spread out of the spin dryer was between 40 and 44%. The spread leaving

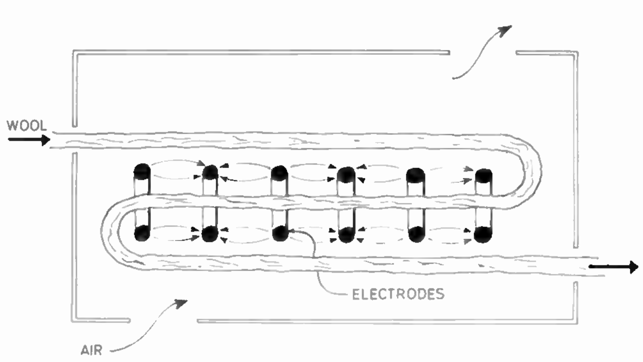

Fig. 5. R.f. wool sliver dryer.

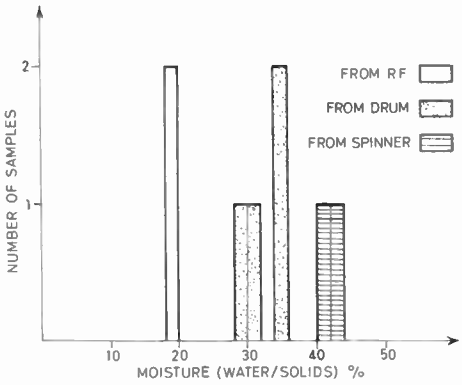

Fig. 6. Moisture levelling by r.f. after drying by spinning followed by drum dryer.

the hot-air drum dryer was between 28 and 42%. On leaving the r.f. dryer this was reduced to $18 \pm 1\%$ with no wet patches at all. Moreover, transferring the last part of the drying to an r.f. system can increase throughput by at least 40%.

Another important aspect is the ability of the r.f. system to tune in power as it is required. Keeping the tuning inductors and capacitors fixed, the power drawn rises and falls with the load introduced between the electrodes. This is shown very clearly in Fig. 7. The power drawn by the wool dryer was directly in proportion to the number of slivers introduced. This has enormous advantages from the point of view of

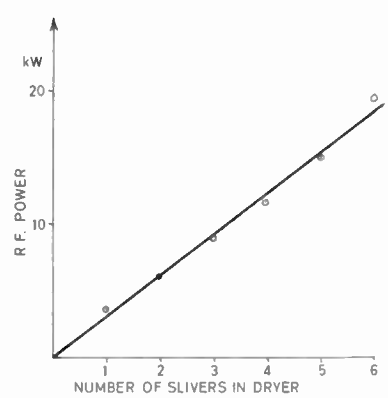

Fig. 7. Power drawn at fixed tuning as a function of number of slivers in r.f. dryer.

both control and energy conservation. For example, if for a period wetter material enters, the r.f. dryer will adjust automatically by delivering more power—obviously a hot-air system has no such response. Equally, during periods where either small amounts of material or no material is being dried the rigid operating conditions of steam-heated hot-air systems ensure that a good deal of energy is wasted. Unlikely as it may appear at first sight, an r.f. system running with a net utilization of primary energy of 17% can be considerably more efficient than hot-air dryers as used in the textile industry because of the poor efficiency of the hot-air drying operation and the temporal inflexibility of boiler-heated systems.[9]

4 Conclusion

It is clear that r.f. can improve both the speed, quality and efficiency of drying operations, thereby reducing costs through product waste, energy waste, insufficient use of capital equipment, and down time in subsequent processing. However, it is essential that the applicator be tailored to the process and that full account be taken of any special characteristics of the material, e.g. its anisotropy or inflammability. So long as reliable product-orientated equipment is put on the market r.f. drying is set fair for a good future.

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Admittance of a dipole antenna driven by a two-wire line

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and

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SUMMARY

A new theoretical approach to the problem of the symmetric dipole antenna, driven by a two-wire line, is presented. The dipole and the line are treated as a unique boundary-value problem, and, as a consequence, a new, conveniently defined, apparent driving-point admittance is introduced.

The analysis of the problem leads to a system of two integral equations with the dipole and line currents as unknown subintegral functions. The integral equations are solved approximately by using the so-called point-matching method. The current on the dipole is approximated by a polynomial with unknown coefficients and that on the line by a sum of incident and reflected travelling waves and a polynomial decreasing rapidly with the distance from the end of the line.

Theoretical results, obtained by the present theory, reveal a strong dependence of the admittance on the distance between the line conductors. Excellent agreement between the theoretical and available experimental results is found.

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1 Introduction

Most of the existing theories of the centre-driven dipole antennas, as well as the base-driven monopole antennas (erected vertically on an infinite perfectly conducting ground), are based on the simplified model of the excitation zone and feeding conditions in general. Whereas in practical transmitting systems the generator and the antenna are usually completely separate structures, connected electrically by a transmission line, many theoretical analyses disregard the transmission line and treat antennas as if they were in themselves complete transmitting systems, combining the antenna and generator into a single, simple unit. Usually, this generator is conceived in such a manner that its impressed field can easily be described mathematically, enabling more or less simple solution to the antenna problem.

Since the pioneer works of Hallén,[1] King and Middleton[2] and others, the delta-function generator has been the most widely used theoretical model of the feeding system. In spite of the fact that this model and the corresponding (Hallén's integral equation have been the basis of many fruitful investigations in the linear antenna theory, they have two well-known inherent shortcomings. First, the integral equation does not have an integrable solution in a rigorous theoretical sense.[3] Secondly, the approximate theoretical susceptance of the antenna diverges and steadily increases with the order of the approximation used. The latter can be readily understood taking into account the fact that a delta-function generator is equivalent to a discontinuity in scalar potential in parallel to a knife-edge capacitance with infinitesimally small separation.[4] However, almost accidentally, low-order approximations lead to the more or less reasonable values of the susceptance.

In order to overcome the difficulties involved with the delta-function generator, some different theoretical approaches have been attempted. One of them consists in introducing a more realistic, but still idealized theoretical model of the real feeding system. For instance, an impressed radial electric field on the annular surface of the coaxial opening[5] has been proved to be a very suitable model in the case of a monopole antenna fed by a coaxial line through the ground plane.

Due to the complexity of the feeding zone, the problem of the symmetrical dipole, fed by a two-wire line, is not suitable for similar mathematical treatment. For line spacings that are not too large a fraction of wavelength, a reasonably satisfactory, but rather artificial method has been devised by R. W. P. King and others.[6,7] This method introduces a terminal-zone network in conjunction with the ideal admittance, determined with a delta-function generator, the infinite susceptance due to the knife edges being subtracted.[8,9]

The purpose of the present paper is to present a theoretically simple but correct approach to the problem of the symmetric dipole antenna driven by a two-wire line (Fig. 1). In the present theoretical analysis the dipole antenna and the feeder are considered as a whole, as a unique boundary-value problem, so that the coupling between antenna and the line, as well as the transmission line end-effect and finite line-conductor spacing are taken into account. A rigorous one-dimensional analysis of the circuit on Fig. 1 requires the solution of simultaneous integral equations for the current-distribution functions on both the antenna and the transmission line.

Since antenna and the transmission line are mutually coupled along the lengths which are of the order of several multiples of the separation between the conductors of the line, the definition of an input admittance Y_{AB} which is the characteristic of the antenna alone and independent of the feeding line is not possible. An admittance in the sense of conventional transmission-line theory can be defined only in a transversal section of the transmission line, A′B′, that is sufficiently far removed from the terminal zone. In defining the input admittance $Y_{A'B'}$ (the input admittance of the part of the line on the right of A′B′) the assumption is implicit that the uniform line continues in both directions from the point of definition.

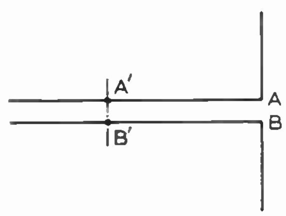


Fig. 1. Antenna as end load for symmetric transmission line.

Since the admittance $Y_{A'B'}$ depends on the position of of the cross-section A′B′, it is useful to define an apparent driving-point admittance Y referred to the end of the line. This admittance can be obtained from $Y_{A'B'}$ by transforming the latter to the end of the line by means of the well-known transmission line formulas. Though the admittance Y may seem to be an artificial quantity (it differs from the admittance Y_{AB}, which is defined as the ratio of the current entering the antenna terminals to the scalar potential difference across the antenna terminals A–B), it is a very useful technical parameter: it prescribes the voltage and current distribution along the line outside the terminal zone and corresponds to the value of the terminal impedance determined by measuring the voltage or current standing-wave ratio (outside the terminal zone).

In order to evaluate $Y_{A'B'}$ and the apparent admittance Y it is necessary first to solve the system of simultaneous integral equations for the circuit on Fig. 1. The system is solved approximately by the so-called point-matching method.[10] The unknown subintegral functions, representing the current distributions on the dipole and on the line, are approximated by the polynomials with unknown coefficients.

The theoretical results, obtained by the present method, are compared with the experimental results of Angelakos (Ref. 6, p. 211). As will be shown, the admittances resulting from the present theory are in excellent agreement with available experimental results. On the other side, the comparison of the theoretical results of the present method with those obtained by using an idealized delta-function generator shows very significant discrepancies.

2 Description of the Method

2.1 The Apparent Admittance and Currents in the System

Consider the circuit consisting of a balanced two-wire line terminated in a symmetric centre-driven dipole antenna. The geometrical arrangement is shown in Fig. 2. The dipole antenna consists of two straight, collinear cylindrical conductors, each of length h' and small radius a, separated in the middle by a gap of half length d. The distance from the centre of the dipole to each of its extremities is denoted by h, so $h = h'+d$.

The transmission line consists of two parallel conductors of radius b, the axes of which are separated by a distance $2d$. The antenna and line conductors are assumed to be perfect.

As shown in Fig. 2 the axis of the dipole coincides with the z-axis of the coordinate system, the y-axis of which is parallel to the axes of the two-wire line and bisects the distance between them.

In order to simplify the analysis it is assumed that the currents on the antenna and on the transmission line are localized on conductor axes. The positive directions of the currents are denoted on the Figure. Due to symmetry the current on the dipole arms, $I_a(z)$, must satisfy the condition:

$$I_a(-z) = I_a(z), \qquad -h \leqslant z \leqslant -d, \quad d \leqslant z \leqslant h. \quad (1)$$

In addition, the current $I_a(z)$ must fulfil the condition

$$I_a(h) = 0. \quad (2)$$

In writing the expression for the current on the transmission line, it is convenient to distinguish two parts of the line:

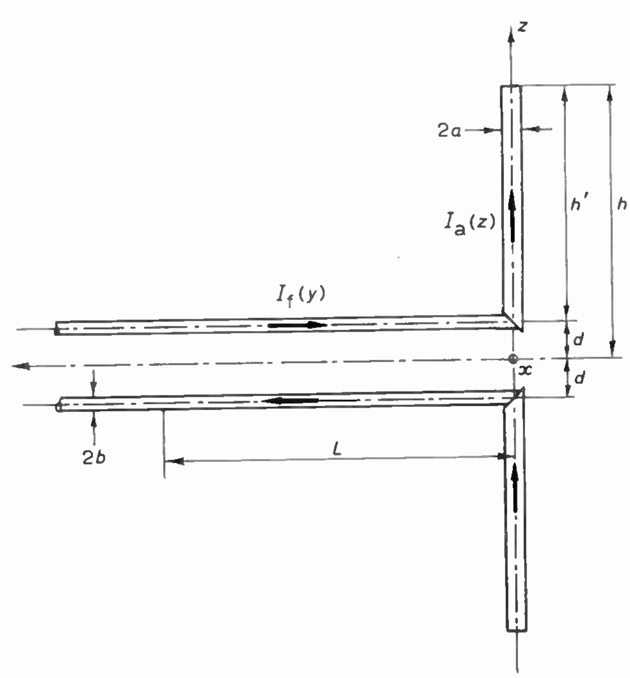


Fig. 2. Notation for symmetric dipole antenna driven by a two-wire line.

(1) the part $y \geqslant L$, where the direct influence of the dipole and of the end-discontinuity of the line can be neglected, and

(2) the part $0 \leqslant y < L$, where the coupling between the dipole and the line, as well as the line end effect, must be taken into account.

As will be seen later, the choice of the length L is not critical. It is quite satisfactory to take it equal to a small multiple of the separation $2d$ between the line conductors.

The current distribution function along the first part of the line, $y \geqslant L$, can be represented in a conventional manner as the sum of the incident and reflected waves,

$$I_f(y) = I_i\, e^{jky} - I_r\, e^{-jky}, \qquad y \geqslant L, \tag{3}$$

where I_i and I_r are unknown complex amplitudes of the two waves and $k = \omega(\varepsilon_0\mu_0)^{1/2} = 2\pi/\lambda$ is the free space propagation constant.

To account for the influence of the dipole and of the line-end on the current distribution on the second part of the line, $0 \leqslant y \leqslant L$, an additional term $I_p(y)$ should be added to (3), so that

$$I_f(y) = I_i\, e^{jky} - I_r\, e^{-jky} + I_p(y), \quad 0 \leqslant y \leqslant L. \tag{4}$$

In order to preserve the continuity of the current at $y = L$, the following conditions must be satisfied:

$$I_p(L) = 0, \qquad \mathrm{d}I_p/\mathrm{d}y\big|_{y=L} = 0. \tag{5}$$

In addition, the current at the end of the line must be equal to the current entering the dipole:

$$I_f(0) = I_a(d). \tag{6}$$

The form of equation (3) implies that on the part of the line $y \geqslant L$ a TEM field exists, so that the conventional transmission-line theory is fully applicable. Consequently, the input admittance, $Y(y)$, in each cross-section of the part of the line $y \geqslant L$ can be defined as

$$\begin{aligned} Y(y) &= I_f(y)/V_f(y) \\ &= (I_i\, e^{jky} - I_r\, e^{-jky})/Z_c(I_i\, e^{jky} + I_r\, e^{-jky}) \\ &= Y_c(e^{j2ky} - R)/(e^{j2ky} + R), \end{aligned} \tag{7}$$

where $Y_c = 1/Z_c$ is the characteristic admittance of the line, and

$$R = I_r/I_i \tag{8}$$

the reflection coefficient.

The admittance given by (7) is just the quantity that can be determined by measuring the v.s.w.r. and the position of the voltage minimum in respect of the cross-section at $y \geqslant L$.

The above-mentioned apparent driving-point admittance Y can be obtained from (7) by putting $y = 0$:

$$Y = Y_c(1-R)/(1+R). \tag{9}$$

In order to evaluate Y, the ratio I_r/I_i, and hence all the currents in the circuit, must be determined first.

2.2 The Components of the Magnetic Vector-potential

The magnetic vector-potential due to the dipole current, $I_a(z)$, has only z-component, which in the field point $M(x, y, z)$ has the following form:

$$A_z = \frac{\mu_0}{4\pi}\int_d^h I_a(z') \times \times [\exp(-jkR_3)/R_3 + \exp(-jkR_4)/R_4]\,\mathrm{d}z' \tag{10}$$

where

$$\begin{aligned} R_3 &= \sqrt{x^2+y^2+(z-z')^2}, \\ R_4 &= \sqrt{x^2+y^2+(z+z')^2}. \end{aligned} \tag{11}$$

The vector-potential due to the current in the two-wire line has a y-component, whose incomplete expression, corresponding to the part of current given by (3), is derived in the Appendix. Taking into account the additional term of the current, $I_p(y)$, the complete expression for the vector-potential can be written as follows:

$$\begin{aligned} A_y = {} & \frac{\mu_0}{4\pi} I_i\, e^{jky}\{-2\ln(r_2/r_1) + \mathrm{Ci}\,(Y_{02}) - \mathrm{Ci}\,(Y_{01}) + \\ & + j[\mathrm{Si}\,(Y_{02}) - \mathrm{Si}\,(Y_{01})]\} + \\ & + \frac{\mu_0}{4\pi} I_r\, e^{-jky} \times \\ & \times\{\mathrm{Ci}\,(Y_{04}) - \mathrm{Ci}\,(Y_{03}) - j[\mathrm{Si}\,(Y_{04}) - \mathrm{Si}\,(Y_{03})]\} - \\ & - \frac{\mu_0}{4\pi}\int_0^L I_p(y') \times \\ & \times [\exp(-jkR_1)/R_1 - \exp(-jkR_2)/R_2]\,\mathrm{d}y', \end{aligned} \tag{12}$$

where

$$r_1 = \sqrt{x^2+(z-d)^2}, \qquad r_2 = \sqrt{x^2+(z+d)^2}, \tag{13}$$

$$R_1 = \sqrt{r_1^2+(y-y')^2}, \qquad R_2 = \sqrt{r_2^2+(y-y')^2}. \tag{14}$$

The $Y_{01}, \ldots, Y_{04}$ are defined in the Appendix.

2.3 Integral Equation Derived from the Boundary Condition on the Surface of the Dipole Conductor

Since the conductor is assumed to be perfect, the tangential component of the electric field strength must vanish on the surface of the dipole. Applying this boundary condition to the z-component of the field strength along the line $x = a$, $y = 0$, $d \leqslant z \leqslant h$ on the surface of the dipole, we can write

$$E_z = -j\frac{\omega}{k^2}\frac{\partial}{\partial z}(\mathrm{div}\,\mathbf{A} + k^2 A_z) = 0\Big|_{\substack{x=a\\y=0\\d\leqslant z\leqslant h}}, \tag{15}$$

or

$$\frac{\partial^2 A_z}{\partial z^2} + k^2 A_z = -\frac{\partial^2 A_y}{\partial y\,\partial z}\Big|_{\substack{x=a\\y=0\\d\leqslant z\leqslant h}}. \tag{16}$$

The solution of the non-homogeneous differential equation (16) consists of the integral of the homogeneous equation (without the term on the right side of (16)),

$$A_{z(\mathrm{hom})} = C_1 \cos k(z-d) + C_2' \sin k(z-d), \tag{17}$$

and a particular integral of the non-homogeneous differential equation

$$A_{z(\mathrm{part})} = -\frac{1}{k}\int_{s=d}^{z} \frac{\partial^2 A_y}{\partial y\,\partial z}\Big|_{\substack{x=a\\y=0\\z=s}} \sin k(z-s)\,\mathrm{d}s. \tag{18}$$

After a partial integration of the particular integral (18) has been performed we find

$$A_z = A_{z(\mathrm{hom})} + A_{z(\mathrm{part})}$$
$$= C_1 \cos k(z-d) + C_2 \sin k(z-d) - \int_{s=d}^{z} \left.\frac{\partial A_y}{\partial y}\right|_{\substack{x=a\\y=0\\z=s}} \cos k(z-s)\,\mathrm{d}s, \quad (19)$$

where

$$C_2 = C_2' + \frac{1}{k}\left.\frac{\partial A_y}{\partial y}\right|_{\substack{x=a\\y=0\\z=d}}.$$

The partial derivative $\partial A_y/\partial y$ in (19) can be obtained from (12), so that we have

$$\left.\frac{\partial A_y}{\partial y}\right|_{\substack{x=a\\y=0\\z=s}} = \mathrm{j}\frac{\mu_0}{4\pi}k\left\{-I_i 2\ln(r_2/r_1) + (I_i - I_r)\left[\int_{kr_1}^{kr_2}\frac{\mathrm{e}^{-\mathrm{j}t}}{t}\,\mathrm{d}t + \exp(-\mathrm{j}kr_2)/\mathrm{j}kr_2 - \exp(-\mathrm{j}kr_1)/\mathrm{j}kr_1\right]\right\}\Bigg|_{\substack{x=a\\z=s}} -$$
$$-\frac{\mu_0}{4\pi}I_p(0)[\exp(-\mathrm{j}kr_1)/r_1 - \exp(-\mathrm{j}kr_2)/r_2]\Big|_{\substack{x=a\\z=s}} -$$
$$-\frac{\mu_0}{4\pi}\int_0^L I_p'(y')[\exp(-\mathrm{j}kR_1)/R_1 - \exp(-\mathrm{j}kR_2)/R_2]\Big|_{\substack{x=a\\y=0\\z=s}}\mathrm{d}y'. \quad (20)$$

Introducing (10) in the left-hand side of (19), we obtain the first of the two integral equations from which the unknown currents should be determined:

$$\frac{\mu_0}{4\pi}\int_d^h I_a(z')[\exp(-\mathrm{j}kR_3)/R_3 + \exp(-\mathrm{j}kR_4)/R_4]\Big|_{\substack{x=a\\y=0}}\mathrm{d}z'$$
$$= C_1\cos k(z-d) + C_2\sin k(z-d) - \int_{s=d}^{z}\left.\frac{\partial A_y}{\partial y}\right|_{\substack{x=a\\y=0\\z=s}}\cos k(z-s)\,\mathrm{d}s. \quad (21)$$

2.4 Integral Equation Derived from the Boundary Condition on the Surface of the Line Conductor

From the boundary condition $E_y = 0$, along the line $x = b$, $z = d$, $y \geqslant 0$ on the surface of the line conductor, we have

$$\frac{\partial^2 A_y}{\partial y^2} + k^2 A_y = -\left.\frac{\partial^2 A_z}{\partial y\,\partial z}\right|_{\substack{x=b\\z=d\\y\geqslant 0}}. \quad (22)$$

Analogously to the previous case, the integral of (22) is found as a sum of the integral of the homogeneous equation

$$A_{y(\mathrm{hom})} = C_3\cos ky + C_4'\sin ky, \quad (23)$$

and a particular integral of the non-homogeneous differential equation

$$A_{y(\mathrm{part})} = -\frac{1}{k}\int_{s=0}^{y}\left.\frac{\partial^2 A_z}{\partial y\,\partial z}\right|_{\substack{x=b\\y=s\\z=d}}\sin k(y-s)\,\mathrm{d}s. \quad (24)$$

By use of partial integration of (24), we can write

$$A_y = A_{y(\mathrm{hom})} + A_{y(\mathrm{part})}$$
$$= C_3\cos ky + C_4\sin ky - \int_{s=0}^{y}\left.\frac{\partial A_a}{\partial z}\right|_{\substack{x=b\\y=s\\z=d}}\cos k(y-s)\,\mathrm{d}s, \quad (25)$$

where

$$C_4 = C_4' + \frac{1}{k}\left.\frac{\partial A_z}{\partial z}\right|_{\substack{x=b\\y=0\\z=d}}.$$

The term $\partial A_z/\partial z$ can be obtained from (10) by partial differentiation. Since

$$\partial R_3/\partial z = -\partial R_3/\partial z' \quad \text{and} \quad \partial R_4/\partial z = \partial R_4/\partial z',$$

after some manipulations involving partial integration and transformation of $\partial A_z/\partial z$, we obtain a form which is suitable for numerical calculations on a digital computer:

$$\left.\frac{\partial A_z}{\partial z}\right|_{\substack{x=b\\z=d}} = \frac{\mu_0}{4\pi}I_a(0)\times$$
$$\times[\exp(-\mathrm{j}kR_3)/R_3 - \exp(-\mathrm{j}kR_4)/R_4]\Big|_{\substack{x=b\\z=d\\z'=d}} +$$
$$+\frac{\mu_0}{4\pi}\int_d^h I_a'(z')\times$$
$$\times[\exp(-\mathrm{j}kR_3)/R_3 - \exp(-\mathrm{j}kR_4)/R_4]\Big|_{\substack{x=b\\z=d}}\mathrm{d}z', \quad (26)$$

where I' means the derivative of I_a.

If (12) is substituted into the left-hand side of (25) and if the account is taken of (26), equation (25) becomes the second integral equation. In order to save the space this integral equation will not be written explicitly and in what follows it will be referred to as equation (25).

One of the constants C_2 and C_4 in equations (19) and (25) can be determined from the condition that the scalar-potentials on the dipole and the line conductor should be equal at the joint of the two conductors. Denoting by φ the scalar potential, the above condition reads

$$\varphi_a(x = a, y = 0, z = d) = \varphi_f(x = b, y = 0, z = d). \quad (27)$$

The scalar-potential can be determined from the vector-potential by means of Lorentz's continuity condition for potentials:

$$\varphi = \frac{\mathrm{j}}{\omega\varepsilon_0\mu_0}\left[\frac{\partial A_y}{\partial y}+\frac{\partial A_z}{\partial z}\right]. \quad (28)$$

From (27) we get

$$C_2 = C_4. \quad (29)$$

3 Approximate Solution to the Integral Equations

The exact solution to the system of the simultaneous integral equations (19) and (25) is not known, but an approximate solution can be obtained by the so-called point-matching method.[10] We assume the currents in the form of finite functional series with unknown complex coefficients. With this series substituted for $I_a(z')$ and $I_f(y)$ in (19) and (25), we calculate the unknown coefficients by prescribing the integral equations to be satisfied at a sufficient number of points along the dipole and the part of the line $0 \leqslant y \leqslant L$.

The limited simple power series in z and y, respectively, appear to be very convenient and most flexible trial functions for the current $I_a(z)$ and the part $I_p(y)$ of the current $I_f(y)$. Let us, therefore, approximate the currents along the dipole and the line by the following expressions, respectively,

$$I_a(z) = \sum_{m=0}^{M} A_m z^m, \quad d \leqslant z \leqslant h, \quad (30)$$

$$I_f(y) = I_i\,\mathrm{e}^{\mathrm{j}ky} - I_r\,\mathrm{e}^{-\mathrm{j}ky} + \begin{cases} 0, & y \geqslant L, \\ \sum_{n=0}^{N} B_n y^n, & 0 \leqslant y \leqslant L. \end{cases} \quad (31)$$

The current distributions (30) and (31) comprise $M+N+4$ unknown coefficients. Together with four constants, C_1, C_2, C_3 and C_4, the total number of unknowns amounts to $M+N+8$.

On the other side, there are four conditions for currents—(2), (5a, b) and (6)—and the relation (22) expressing the scalar potential condition, so that the total number of unknowns is reduced by five. Besides, since only the ratio I_r/I_i is needed for determination of Y, I_i can be chosen to be equal to unity, and, hence, it remains to evaluate $M+N+2$ unknowns.

According to the point-matching method these unknowns can be determined by satisfying the integral equations (19) and (25) (with the approximations (30) and (31) included) at $M+N+2$ points along the dipole and the part of the line $0 \leqslant y \leqslant L$. In principle these points can be selected arbitrarily, but it seems to be quite natural to select $M+1$ points along the dipole arm ($d \leqslant z \leqslant h$) and $N+1$ points along the part of the line $0 \leqslant y \leqslant L$. In addition, the selected points on the dipole, as well as those on the part of the line, are taken to be equidistant:

$$z_k = d+k(h-d)/M, \quad k = 0, 1, \ldots, M, \quad (32)$$

$$y_p = pL/N, \quad p = 0, 1, \ldots, N. \quad (33)$$

Substituting successively different values z_k for z in (19), and y_p for y in (25), and evaluating the corresponding integrals, we obtain a system of $M+N+2$ linear equations containing the unknown complex coefficients and constants which determine the current distribution functions. By solving the system these unknowns can be evaluated.

4 Numerical Results and Comparison with the Experiment

The method described has been used to calculate the admittance for a number of representative examples. However, before proceeding further with these examples, some preliminary checks of the validity of the theory and estimates of the influence of some parameters on the convergence of the results seem to be appropriate.

First, as in the other applications of the point-matching method and the polynomial approximation for the antenna current distribution, it has been found that low-order approximating polynomials (M, N = 2, 3, 4, depending on the ratio h/λ) yield fairly fast convergence of the results.

The influence of the length, L, of the 'perturbed' part of the line has also been examined and estimated. Contrary to the expectation, its influence has been found to be fairly slight. As an illustration, the theoretical admittances of two particular dipoles, calculated with different ratios L/λ are presented in Tables 1 and 2. In Table 1 the admittance of a dipole of the half-length $h = 0{\cdot}25\ \lambda$ is shown as a function of L/λ. Other geometrical dimensions of the dipole and of the line are defined as follows:

$$\Omega = 2\ln(2h/a) = 10, \qquad d/h = 0{\cdot}01,$$

$$Z_c = 120\ln\left[d/b+\sqrt{(d/b)^2-1}\right] = 300 \text{ ohms}.$$

Table 1

Admittance of a half-wave dipole fed by a two-wire line as a function of L/λ

$h = 0{\cdot}25\ \lambda$, $\Omega = 10$, $d/h = 0{\cdot}01$, $Z_c = 300$ ohms, $M = N = 3$

L/λ	0·001	0·05	0·10
Y mS	9·551 − j4·375	9·365 − j4·255	9·123 − j4·245
L/λ	0·15	0·20	0·25
Y mS	9·146 − j4·176	9·268 − j4·153	9·076 − j4·185

It can be seen from this Table that the length L is not critical at all and that, starting from approximately $L = 0{\cdot}1\ \lambda$, the results converge in a satisfactory manner. In order to verify this conclusion another example has been elaborated and the results are presented in Table 2.

By inspection we conclude that the lengths $L = 0{\cdot}1\ \lambda$ and $L = 0{\cdot}25\ \lambda$ provide practically the same results.

After the influence of the parameter L has been estimated, a theoretical but important check of the theory was performed: the admittances corresponding

Table 2

Admittances of full-wave dipoles with different spacings between the line conductors

$h = 0{\cdot}5\,\lambda$, $\Omega = 10$, $Z_0 = 300$ ohms, $M = N = 3$

L/λ	Y mS $d/h = 0{\cdot}001$	$d/h = 0{\cdot}01$	$d/h = 0{\cdot}05$	$d/h = 0{\cdot}10$
0·10	0·951 + j1·505	0·962 + j1·214	0·996 + j0·448	1·039 − j0·137
0·25	0·952 + j1·505	0·935 + j1·172	0·965 + j0·413	1·069 − j0·119

to a very small spacing between the line conductors ($d = 0{\cdot}001h$ and $Z_c = 300$ ohms), as obtained by the present theory, were compared with those resulting from the theory which is based on the idealized delta-function generator.[11] In both cases the same order of the polynomials, approximating the current distribution along the dipole, was used (for $kh \leqslant 2$, $M = 2$ and for $kh > 2$, $M = 3$). The numerical results are presented in Table 3, where Y_δ denotes admittances corresponding to the delta-function generator and Y those obtained by the present theory.

Table 3

Comparison of admittances of the half-wave dipoles fed by a delta-function generator and by a two-wire line with very small conductor spacing

$h = 0{\cdot}25\,\lambda$, $\Omega = 10$, $d/h = 0{\cdot}001$, $Z_0 = 300$ ohms

kh	Y mS	Y_δ mS
1·0	0·400 + j4·063	0·400 + j4·171
1·2	1·991 + j6·948	2·021 + j7·075
1·4	11·879 + j6·748	12·232 + j7·415
1·6	7·893 − j4·399	7·885 − j4·512
1·8	3·422 − j1·943	3·380 − j3·181
2·0	2·104 − j1·943	2·098 − j1·930
2·2	1·573 − j1·024	1·571 − j0·994
2·4	1·296 − j0·376	1·290 − j0·334
2·6	1·130 + j0·165	1·127 + j0·213
2·8	1·031 + j0·655	1·026 + j0·707
3·0	0·971 + j1·126	0·966 + j1·188
3·2	0·946 + j1·620	0·941 + j1·689

Although the general agreement between the above cited theoretical results is very satisfactory, it should be noted that these results refer to a particular case only. As far as the validity of the theory is concerned, the above agreement is a necessary but not a sufficient condition.

Of course, the most competent support to a theory is provided by experiment. Unfortunately the published experimental data concerning the admittances (impedances) of dipoles driven by a two-wire line are rather rare and often refer to somewhat specific feeding conditions (the antenna as end load with high-impedance stub support, or antenna as a centre load with equal and opposite generators at the ends of the line—Ref. 6, p. 208). Probably the most reliable experimental data, which are very suitable for direct comparison with the theoretical results, are those presented by Angelakos (Ref. 6, Figs. 34.7*a* and 34.7*b*).

In order to eliminate the difficulties connected with the conventional open-wire lines, Angelakos used the image-plane line and monopole antenna in the measurements. The end of the line and monopole antenna were supported either by styrofoam supports or by a high-impedance stub. The measured impedances in the two cases differ significantly and both are given in the cited reference. For the purpose of comparison with the theory the experimental impedances obtained in the measurements with the styrofoam supports have been used. These impedances, converted into equivalent admittances, are shown in Fig. 3.

In Angelakos' experiment the frequency was kept constant at 750 MHz ($\lambda = 0{\cdot}4$ m) and the line had the following dimensions: $b = 3{\cdot}17$ mm, $2d = 19{\cdot}626$ mm ($Z_c = 215{\cdot}4$ ohms). The radius of the antenna conductor was the same as that of the line conductor, i.e. $a = b = 3{\cdot}17$ mm, and the length of the dipole was varied within the limits $1{\cdot}4 \leqslant kh \leqslant 4$ ($k = 2\pi/\lambda$). The

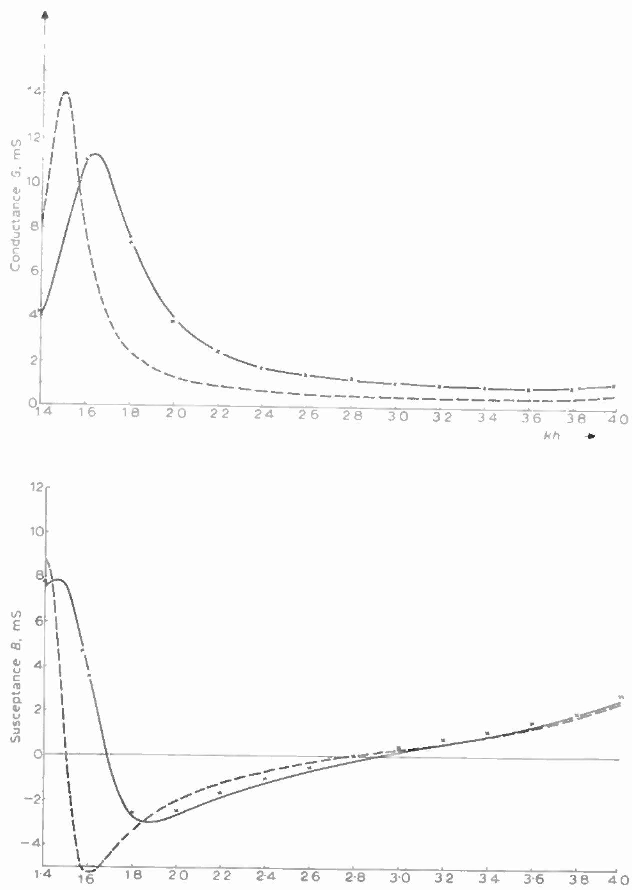


Fig. 3. Theoretical and experimental conductance, G, and susceptance, B, as functions of kh.

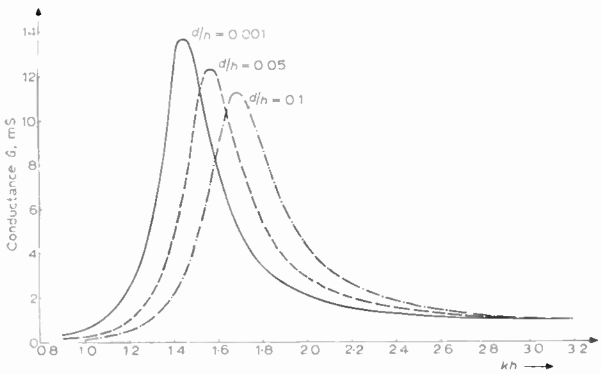
Dipole antenna fed by a two-wire line; $a = b = 3{\cdot}17$ mm, $2d = 19{\cdot}626$ mm, $Z_0 = 215{\cdot}4$ ohms, $\lambda = 40$ cm.

——— present theory

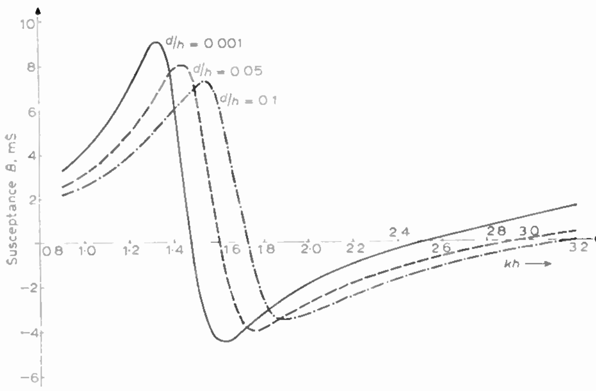
× × × × measured (Angelakos).

Dipole antenna fed by a delta-function generator; $a = 3{\cdot}17$ mm, $\lambda = 40$ cm.

------- theoretical.



(a) Theoretical conductance, G, as a function of kh with d/h as parameter.



(b) Theoretical susceptance, B, as a function of kh with d/h as parameter.

Fig. 4.

same data were used in calculating the theoretical admittances. These were evaluated using polynomials of the order $M = 4$ and $N = 3$, and assuming $L = 0{\cdot}25\,\lambda$. Theoretical admittances are shown on the diagram in Fig. 3 together with the corresponding experimental results. The agreement between the theoretical and experimental data is really excellent and unexpected. A small, constant difference in the susceptance, of about 0·25 mS, can be explained by the shunting effect of the styrofoam support. At the frequency of 750 MHz this difference corresponds to a shunting capacitance of about 0·05 pF only.

For comparison, in Fig. 3 the admittances corresponding to the idealized delta-function and generator and to the same dipole dimensions are also shown. Note the very remarkable discrepancies between the theoretical admittances obtained with idealized and real feeding conditions respectively.

In addition, some other examples of the dipole antenna were analysed and the apparent driving-point admittances calculated. This time all dimensions of the dipole and line were kept constant and the frequency was varied. The geometry of the dipole and line was defined by the following parameters:

$$\Omega = 2 \ln (2h/a) = 10, \qquad Z_c = 300 \text{ ohms},$$
$$d/h = 0{\cdot}001;\ 0{\cdot}05;\ 0{\cdot}1.$$

The ratio b/d is implicitly contained in Z_c. The calculated conductance and susceptance, as functions of kh, are shown in Fig. 4. It is seen that the influence of the line conductors on the admittance is very significant. With the increase of the ratio d/h the maxima of both the conductance and susceptance curves become smaller and both curves are shifted towards larger values of kh.

5 Conclusions

A new theoretical approach to the problem of the symmetric dipole antenna, driven by a two-wire line, has been presented in this paper. The described method treats the dipole and the transmission line as a unique boundary-value problem leading to a system of two simultaneous integral equations. These integral equations have been approximately solved by using the so-called point-matching method and the polynomial approximation for the unknown currents on the dipole and on the line. In order to overcome the difficulties, caused by the mutual coupling between the dipole and transmission line, a new, suitably defined apparent driving-point admittance has been introduced and calculated.

The described method has been used to calculate the apparent admittance for a number of examples. A remarkable dependence of the admittance with respect to distance between the transmission line conductors, as well as the inadequacy of the commonly used, idealized delta-function generator have been found. Excellent agreement of theoretical and experimental results for admittances available in the literature has been established.

6 Acknowledgment

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8 Appendix

Magnetic vector-potential of a semi-infinite two-wire line carrying progressive current waves

Consider a semi-infinite two-wire line, beginning at $y = 0$ (Fig. 5) and ending at infinity ($y \to \infty$). Let the line carry the progressive current wave

$$I(y) = I_i \, e^{jky}, \tag{34}$$

travelling in the negative direction of y-axis.

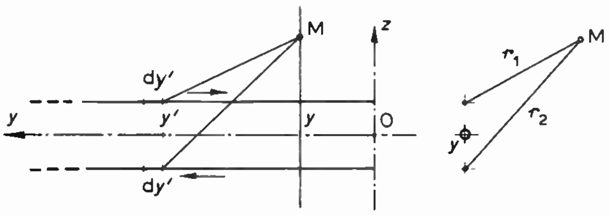

Fig. 5. Notation for semi-infinite two-wire line.

With the proximity effect disregarded, the currents can be located on the axes of the wires and the vector-potential at a point M (r_1, r_2, y), out of the conductors, can be written in the form

$$A_y = -\frac{\mu_0}{4\pi} I_i \int_0^\infty e^{jky'} \times$$
$$\times [\exp(-jkR_1)/R_1 - \exp(-jkR_2)/R_2] \, dy', \tag{35}$$

where

$$R_1 = \sqrt{r_1^2 + (y-y')^2}, \qquad R_2 = \sqrt{r_2^2 + (y-y')^2},$$

r_1 and r_2 being the bipolar coordinates of the point M in the transverse plane y.

Introducing the new variable $u = y' - y$ (y is assumed to be finite) and by denoting the new limits of the integral by $p = -y$ and $q \to \infty$, the integral (35) can be put in the form

$$A_y = -\frac{\mu_0}{4\pi} I_i \, e^{jky} \left[\int_p^q \frac{e^{jk(u - \sqrt{r_1^2+u^2})}}{\sqrt{r_1^2+u^2}} \, du - \right.$$
$$\left. - \int_p^q \frac{e^{jk(u-\sqrt{r_2^2+u^2})}}{\sqrt{r_2^2+u^2}} \, du \right]. \tag{36}$$

By a new change of the variable

$$t = k(u - \sqrt{r^2+u^2}), \qquad dt = \frac{-t}{\sqrt{r^2+u^2}} \, du,$$

(36) reads

$$A_y = \frac{\mu_0}{4\pi} I_i \, e^{jky} \left[\int_{k(p-\sqrt{r_1^2+p^2})}^{k(q-\sqrt{r_1^2+q^2})} \frac{e^{jt}}{t} \, dt - \int_{k(p-\sqrt{r_2^2+p^2})}^{k(q-\sqrt{r_2^2+q^2})} \frac{e^{jt}}{t} \, dt \right]. \tag{37}$$

Since

$$\int_a^b - \int_c^d = \int_a^b + \int_b^c - \int_b^c - \int_c^d = \int_a^c - \int_b^d$$

the integral in the square bracket in (37) can be written as follows:

$$J = J_1 - J_2 = \int_{k(p-\sqrt{r_1^2+p^2})}^{k(p-\sqrt{r_2^2+p^2})} \frac{e^{jt}}{t} \, dt - \int_{k(q-\sqrt{r_1^2+q^2})}^{k(q-\sqrt{r_2^2+q^2})} \frac{e^{jt}}{t} \, dt. \tag{38}$$

Taking into account that $p = -y$, and putting, for abbreviation,

$$Y_{01} = k(-y - \sqrt{r_1^2+y^2}), \qquad Y_{02} = k(-y - \sqrt{r_2^2+y^2}),$$

the integral J_1 can be expressed in terms of sine- and cosine-integral functions:

$$J_1 = \int_{Y_{01}}^{Y_{02}} \frac{e^{jt}}{t} \, dt$$
$$= \mathrm{Ci}(Y_{02}) - \mathrm{Ci}(Y_{01}) + j[\mathrm{Si}(Y_{02}) - \mathrm{Si}(Y_{01})], \tag{39}$$

where

$$\mathrm{Ci}(x) = \int_\infty^x \frac{\cos t}{t} \, dt, \qquad \mathrm{Si}(x) = \int_\infty^x \frac{\sin t}{t} \, dt.$$

The second part of the integral (38) can be evaluated by the help of the mean value theorem:

$$J_2 = e^{jP} \ln \frac{q - \sqrt{r_2^2+q^2}}{q - \sqrt{r_1^2+q^2}}, \tag{40}$$

where

$$k(q - \sqrt{r_1^2+q^2}) \leqslant P \leqslant k(q - \sqrt{r_2^2+q^2}).$$

When $q \to \infty$, $P \to 0$ and

$$J_2 = \lim_{q\to\infty} \ln \frac{q - \sqrt{r_2^2+q^2}}{q - \sqrt{r_1^2+q^2}} = 2 \ln \frac{r_2}{r_1}. \tag{41}$$

Therefore,

$$A_y = \frac{\mu_0}{4\pi} I_i \, e^{jky} \left\{ -2 \ln \frac{r_2}{r_1} + \mathrm{Ci}(Y_{02}) - \mathrm{Ci}(Y_{01}) + \right.$$
$$\left. + j[\mathrm{Si}(Y_{02}) - \mathrm{Si}(Y_{01})] \right\}. \tag{42}$$

If on the same semi-infinite two-wire line ($0 \leqslant y < \infty$), a progressive current wave of the form

$$I(y) = -I_r \, e^{-jky} \tag{43}$$

is present (the wave propagates in the positive direction of the y-axis), the vector-potential can be derived in a similar way and it has the following form

$$A_y = \frac{\mu_0}{4\pi} I_r \, e^{-jky} \times$$
$$\times \{\mathrm{Ci}(Y_{04}) - \mathrm{Ci}(Y_{03}) - j[\mathrm{Si}(Y_{04}) - \mathrm{Si}(Y_{03})]\}, \tag{44}$$

where

$$Y_{03} = k(-y + \sqrt{r_1^2+y^2}), \quad Y_{04} = k(-y + \sqrt{r_2^2+y^2}). \tag{44}$$

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Design of a flexible phase reversal modulation correlator

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SUMMARY

A digital correlator is described for use with intra-pulse phase reversal modulated signals. The length and content of the binary sequence used can be altered at will from the front panel which adds considerably to its flexibility. It is particularly suited to ionospheric sounding but can find application in many areas of finite sequence correlation, e.g. radar.

The paper discusses the reasons leading to the choice of technology, particularly where cost and difficulty of meeting the specification with readily procured components was in question.

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1 Introduction

The requirement for the correlator to be described in this paper originated in the field of h.f. communications where the technique of intra-pulse reversal modulation is practiced in ionospheric sounding. Use of intra-pulse modulation enables the sounding pulses to be transmitted as long (e.g. several milliseconds), relatively low power pulses which, on reception after reflection from the ionosphere, can be time-compressed to improve the sounding resolution and signal/noise ratio. Basic descriptions of the intra-pulse modulation techniques and systems are well documented;[1] this paper is concerned with the design and development of a pulse compression network for use with a system using phase reversal modulation but flexible enough to cover a wide range of the possible signal variables.

Because this work was carried out by a University Department on behalf of a client with a real operational requirement, the application of 'state of the art' technology was limited to those devices available as production items. However, the design is such that it is possible to test the practical limits of some accepted theoretical criteria, e.g. Nyquist sampling criteria, etc.

2 General Considerations

The outline requirement as presented was for a basic p.r.m. correlator for use with a 28-bit phase-reversal modulated signal on a band limited carrier frequency of 100 ± 8 kHz. The total length of the modulated composite pulse was to be 2·8 ms and the duration of the correlated peak signal about 100 μs. In addition, the system was to be made flexible enough to cope with changes of code (within the limit of 28 bits maximum) and possibly changes of the overall time scale and carrier frequency. A further requirement was the ability to function satisfactorily in the presence of multiple signals, some of which may be Doppler shifted in frequency by up to 20 Hz.

The long intra-pulse modulated signal being considered may be treated as a number, say n, of contiguous shorter pulses of equal duration all derived from a single coherent signal source with which they are either in phase or in anti-phase. A practical constraint is that the frequency of the coherent signal source f_c must be related to the duration of the elemental short pulses t_p such that $t_p = l/f_c$ where l is an integer.

The basic action of the pulse compression network is illustrated by Fig. 1. The ratio of maximum to second highest output while there is some signal in the line is referred to as the peak/sidelobe ratio. This ratio is determined by the number of elements in the signal (n) and the chosen code sequence of phase reversals.

On inspection, it can be seen that the action of the pulse compression network is to generate the cross-correlation function between the input signal to the delay line and the signal coding set up by the amplifier states. If the input signal is $V_1(t)$ and the signal with which it is to be correlated as determined by the delay line and amplifier selection characteristics is $V_2(t)$, then the output of the summer represents their cross-correlation

$R_{1,2}(\tau)$ where:[2]

$$R_{1,2}(\tau) = \frac{1}{T_0}\int_{-\frac{1}{2}T_0}^{\frac{1}{2}T_0} V_2(t).V_1(t+\tau)\,dt.$$

The system of Fig. 1 is not a general-purpose correlator since one of the functions being correlated is determined by a hardware configuration, hence it is described as a 'p.r.m. correlator'. The particular range of the 28-unit p.r.m. codes used operationally is found by computer search of available codes, the criterion for selection being optimum peak/sidelobe ratio of the auto-correlation function (which in this case is about 14).

Two basic techniques were studied as means for implementing the requirement—these being analogue and digital techniques. After consideration of the current state of technology, tolerance margins, cost, and degrees of technical risk, it was decided to proceed with the development of a digital system. The major points of the study are set out in the following Sections.

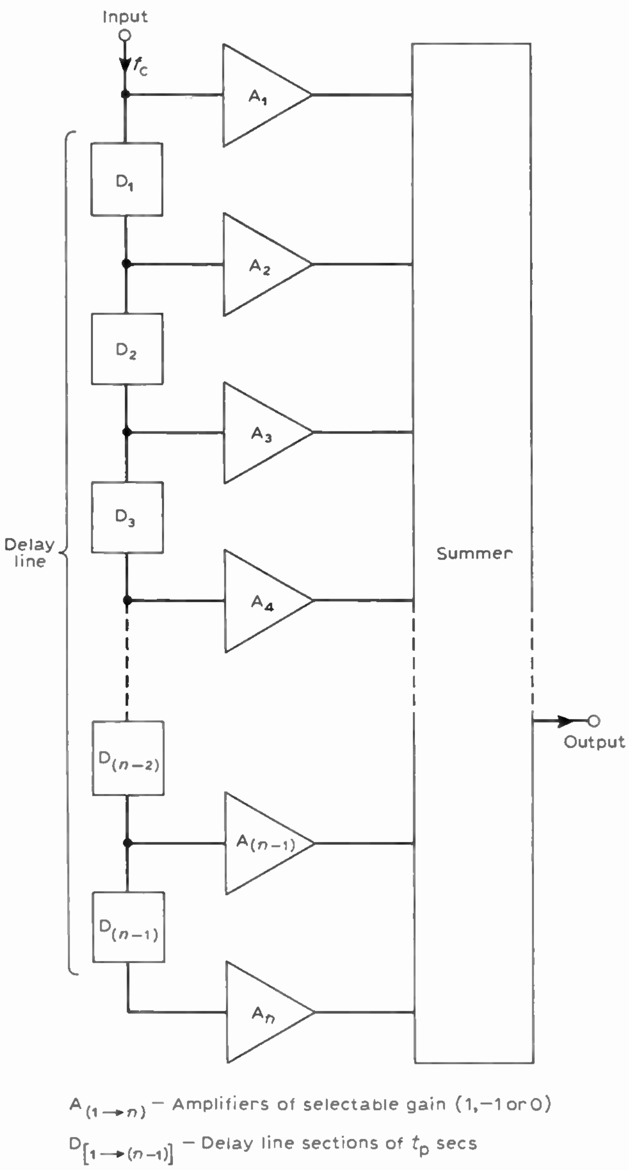


Fig. 1. Basic phase-reversal-modulation correlator.

3 Basic Techniques

3.1 Analogue

The use of analogue techniques to implement the basic system of Fig. 1 is an application of well-known devices. The delay line with its 27 sections and 28 taps ($n = 28$ (maximum) being the requirement) can be obtained in several forms, e.g. lumped constant, distributed line, ultrasonic glass rod, surface acoustic waves, etc. Since the amplifiers are required to multiply their input signals by $+1$, -1 or 0, they can take the form of any of a wide range of integrated circuit operational amplifiers which satisfy the 100 kHz parameter.

Clearly the critical item is the delay line and this may be looked at from two aspects:

(a) the technical specification and cost, and

(b) its effect on the basic flexibility requirement.

The technical specification calls for a 27-section delay line of 100 µs per section. Since the carrier frequency is 100 kHz and the predetection information bandwidth is about 16 kHz, then we must also specify 100 kHz working with a 16 kHz non-dispersive bandwidth. Required tolerances can be derived from the fact that 28 contiguous signals are summed after their phase coherence has been restored by the delay line. If a maximum phase error of $\pm\pi/10$ is postulated as acceptable then the total permissible delay deviation from nominal is $\pm 0{\cdot}5$ µs in 2·7 ms, i.e. about ± 2 parts in 10^4. However, since it is also required to work with Doppler-shifted signals of up to 20 Hz, even with a nominal delay line this maximum Doppler shift would itself introduce a cumulative phase error of about $\pm\pi/10$. This means that, taken at face value, the requirements to specify a realistic delay line tolerance and work over a range of Doppler shifts are incompatible. From the point of view of flexibility, the use of analogue delay lines leaves unrestricted the selection of codes up to 28 bits, but it does not meet the requirement for change of time scale and carrier frequency.

3.2 Digital

Since the delay line is the critical system component, the feasibility of implementing it in digital form was first investigated. Input signals are in analogue form so that it is necessary to convert these into digital form by a sampling analogue to digital converter (a.d.c.). Because of the 100 kHz signal carrier, a necessary feature of the a.d.c. is that its sampling aperture should be less than 0·5 µs, and its conversion time less than the sampling period.[3]

At this early study stage it was not possible closely to define the limits of the possible sampling period, but the gross boundaries were reasonably clear. These derived from the fact that if the sampling rate was determined by the 100 kHz carrier frequency, then the maximum sampling rate was likely to be about 600 kHz. Alternatively, if the sampling rate was determined by the information bandwidth, the minimum sampling rate was likely to be about 30 kHz. Available a.d.c.s in the 600 kHz to 30 kHz range are generally of the 8-bit resolution type which should be more than adequate to cope with signals of, say, 40 dB dynamic range.

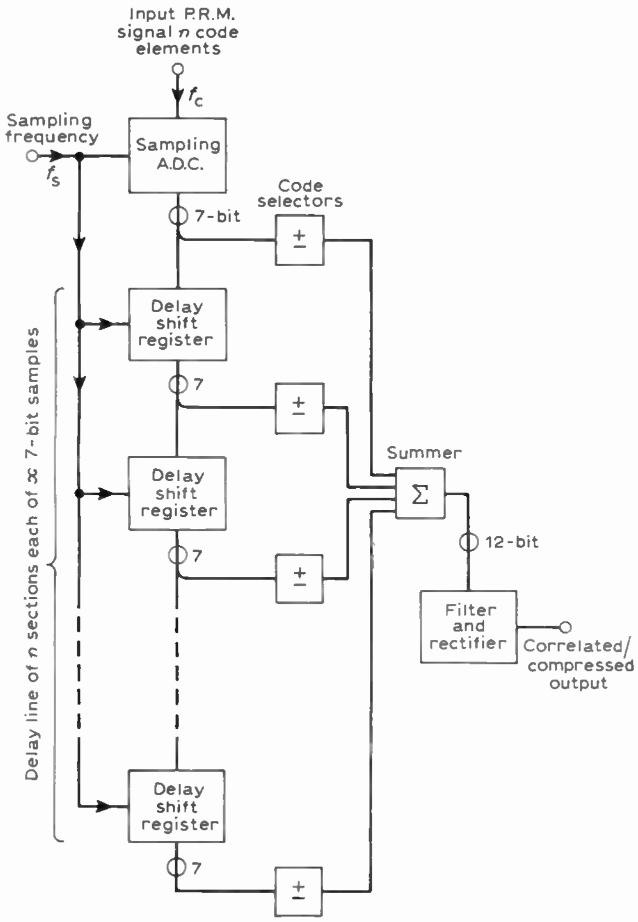


Fig. 2. Block diagram of digital correlator.

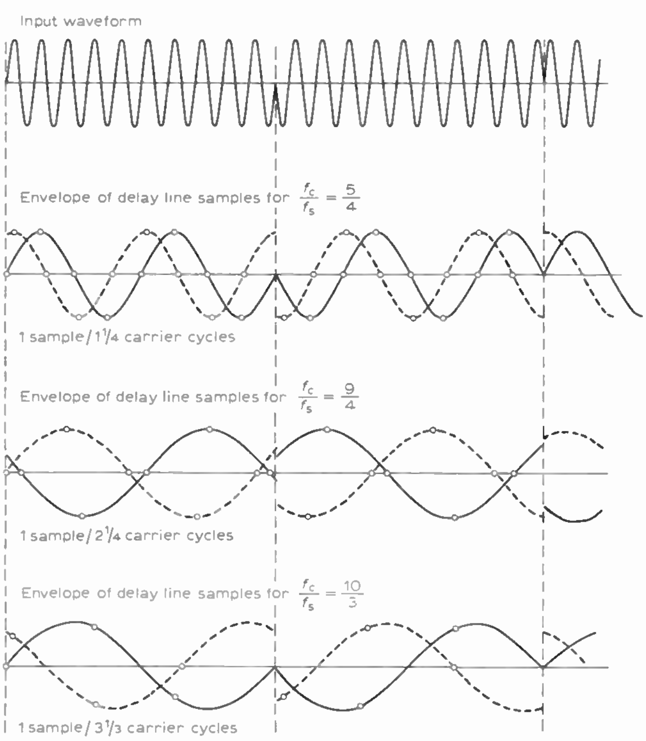


Fig. 3. Input and sampled envelope waveform relationships.

The most obvious implementation of a digital delay line was the construction of a parallel-bit shift register of the appropriate number of bits and tapping points. In this case (as in all digital systems) the precision of the delay line constants is determined entirely by the applied clocking frequency and this point allows an effective tolerance to Doppler shifts in signal frequency and changes of time scale. Summing could either be carried out by converting the delay line outputs back into analogue form and summing operationally after insertion of the appropriate phase reversal code, or the delay line outputs could be summed digitally in an arithmetic operation.

4 System Considerations

The fundamental system choice made initially was that between the analogue and digital approaches. For reasons of flexibility and precision, the advantage appeared to be with the digital technique. When cost considerations were applied, the issue resolved itself quite clearly. Accepting the fact that the delay line is the paramount cost bearing item in either system and comparing basic costs in this area, the comparative figures were as follows:

Analogue: Cost of 100 μs delay lines was quoted at about £100 each. Total delay line cost—not less than £2,700.

Digital: The cost of a digital delay line depended on the number of bits of total storage required and assuming 7-bit sample quantization, the total number of bits would lie between 11,340 and 567. Cost of storing this range of bits in a tapped shift register was about £250, or using available random access memories with sequence controlled addressing (to create effectively a tapped line) about £75 (including addressing logic). To these digital delay line costs it was necessary to add the cost of the input sampling a.d.c. (about a further £70). Total digital costs were between, say, £145 and £320 in the worst case.

Clearly the case for the digital system was made on both technical and cost grounds; further systems considerations are restricted to the digital system.

Figure 2 shows the basic block diagram of the digital correlator; for convenience, the delay line is shown as a series of 7-bit shift registers each of x samples. The relationships between the input signal waveform and the envelope of the digital sample signals in the delay line are shown in Fig. 3 for three different ratios of $f_c : f_s$, where f_c is the signal carrier frequency (100 kHz) and f_s is the sampling frequency. These signal waveforms demonstrate that the period of the envelope of the digital samples must be a sub-multiple of a signal code element.

The delay per section of the delay line is x/f_s seconds and the total delay is nx/f_s seconds (which in this case is 2·7 ms).

The number of carrier frequency cycles per delay section is xf_c/f_s.

Let m be the ratio of f_c to the envelope frequency of the digital signals in the delay line and y the number of digital samples per envelope cycle. Then because the delay per section must be an integer multiple of the envelope period, the number of carrier cycles per sample is m/y where m and y are both integers and m/y is irreducible.

y is limited by the Nyquist criterion, i.e. $\geqslant 2 \times$ envelope frequency.

The resulting sample frequency

$$f_s = f_c \bigg/ \frac{m}{y} = y\,\frac{f_c}{m}.$$

A limit is imposed on the value of m because the envelope period must be a multiple of $1/f_c$ and less than the duration of a code element (for this case less than $10/f_c$.)

In order to avoid difficulties due to operating close to the theoretical limits, it was decided to proceed with the design based on a sampling frequency of $100 \times 10^3 \times \frac{4}{5} =$ 80 kHz. This provided a compressed output signal of two whole cycles duration which, after full-wave rectification, provided two octaves of frequency separation between information and signal ripple, so that filtering could be effectively applied.

A final system parameter remained to be examined—the requirement for the system to work in the presence of selective Doppler shifts. This was assessed on the grounds that an acceptable phase error introduced by the shift over the whole signal length of 2·7 ms was about $\pm 18°$ on the 100 kHz carrier ($\pm \frac{1}{2}$ μs in 2·7 ms) representing a worsening in peak to sidelobe compression ratios of about 10%. Converting this back to a ratio gives a tolerable frequency shift of $\pm$ 1 part in 5400. Therefore the permissible Doppler frequency shift for the assumed criteria equals

$$\frac{10^5}{5{\cdot}4 \times 10^3} = \pm 18{\cdot}5 \text{ Hz};$$

this compares well with the target figure of ± 20 Hz.

It was intended that the system should be so designed as to permit variations in sampling rate and storage capacity so that the theoretical limiting condition could be approached. By so doing, an evaluation program into possible modes of error generation and/or failure could be carried out at a later date.

5 System Design

5.1 Delay Line

From the above, the delay line can now be defined as being $2700 \times 80/1000 = 216$ bits long by 7 bits wide and tapped every 8 bits. Two basic methods exist for implementing such a line:

(i) shift registers, or

(ii) random access memories (r.a.m.).

For (i), the data move physically along a line of storage cells whereas for (ii), the information, once stored, stays in fixed locations and the point of reference moves. In both cases, provision of storage for multiple bit samples is achieved by replication of the basic storage elements. The cells are then operated in parallel.

Although the shift register method is the obvious one to choose because of its conceptual simplicity, it was rejected on the following grounds:

(a) Metal-oxide-semiconductor shift registers are generally not available in tapped form.

(b) Bipolar ones are generally very short (8 bits) because of dissipation and lead-out limitations. They would therefore be required in large numbers for the required storage capacity.

On the other hand, r.a.m.s provide, not only access to any cell position within a single cycle of operation, but also the ability to change the effective tap positions by suitable addressing sequences. Furthermore, considerable storage can be provided on a single chip and at a modest dissipation per bit (< 2 mW).

5.2 Processing

Apart from the arithmetic section, the remaining major point of consideration is the provision of programmable multiplication coefficients for the signals emerging from the 'delay-line' taps. These are limited to ± 1 and 0, and can therefore be programmed using tri-stable panel switches. However, the use of a r.a.m. storage system implies access to each tap output in some sequence according to the addressing method. The switch positions can be multiplexed into the arithmetic unit in the required sequence giving a spin-off of harness and arithmetic logic simplification.†

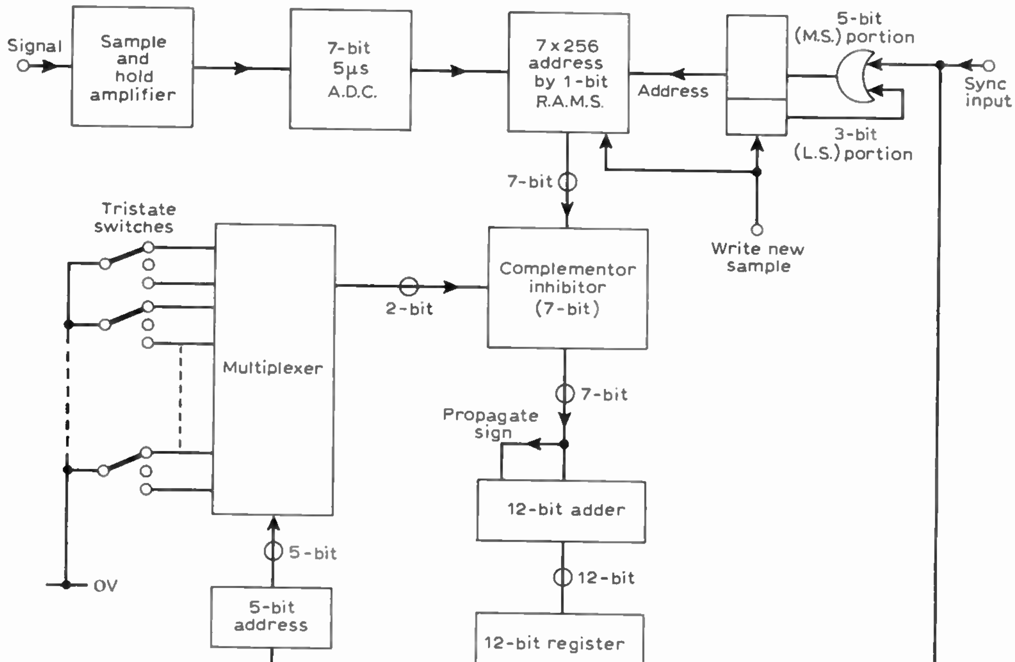
Use of a moving point of reference for the delay line makes necessary a non-uniform switch sampling sequence. (The sampling order stays constant but the sequence starts with a different switch every 8th sample.) This results in the requirement for a separate address generator for the multiplexor and r.a.m. address mechanism.

The arithmetic unit is required to sum 28 weighted 7-bit samples between each signal sampling operation (12·5 μs ≡ 80 kHz). The simple program weights $k(k\varepsilon \pm 1, 0)$ allow the use of a simple programmable inhibitor/ones complementer to be interposed between the 'delay-line' and digital accumulator. The former is controlled by the currently sampled panel switch. In order to accommodate the greatest possible magnitude of grand total at any time, the accumulator needs to be $\geqslant 7 + \log_2 28$ bits wide, i.e. 12 bits.

6 Engineering Constraints

The p.r.m. correlator described in this paper was designed and constructed with a particular specification in mind. However, it is of interest to investigate the nature of the limitations of this basic method of correlation, particularly as it refers to operation speed. Assuming the use of bipolar technology throughout the digital section, the first speed limitation occurs in the

† Such sequenced access to the 'delay-line' and coefficient multiplexor is only possible if sufficient time is available between each signal sample.



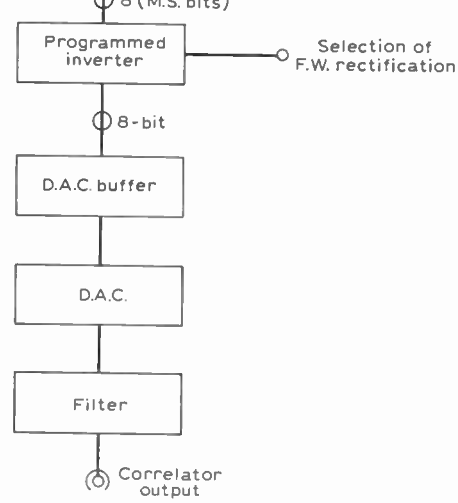


Fig. 4. Digital correlator data-flow.

analogue sampling/a.d.c. area. The sample/hold amplifier is the only part of the machine that has to be specified with the carrier, rather than the information frequency, in mind. It must have an aperture time much less than the period of the carrier (10 μs in this case). However, in generally considering this correlation method, it is not necessary to be tied to a carrier/information frequency ratio of 20 : 1—that is merely a characteristic of the system described in this paper. Clearly a reduction of this ratio would relieve the sample/hold amplifier specification.

After the sample/hold amplifier, the a.d.c. is probably the device working most closely to its 'state-of-the-art' limits. Where the specification demands very high information rates, several a.d.c.s could be used in an interleaved operation. However, such a system could be very costly.

The arithmetic section and r.a.m. has some performance in hand for the system specified. The r.a.m.s have a typical access time of less than 150 ns (including address generation) and the arithmetic unit needs less than 200 ns to process each tap output. The 200 ns can be considerably reduced by the use of 'carry-save techniques' or the use of Wallace adder trees, etc. The r.a.m. access time can be merged to some extent with the arithmetic to effect a further saving of time.

7 Hardware

This is shown in outline in Fig. 4. Sampling of the 100 kHz input signal is triggered by a synchronization pulse derived from the coherent receiver. The 7-bit digitized signal is then stored in seven similarly addressed 256-bit bipolar r.a.m.s. The address for the r.a.m.s is segmented into two portions—3 least significant bits representing the 8 cell positions between 'taps' and 5 most significant bits representing the 28 sections (4 spares). Therefore, by incrementing the 5-bit portion, the point of reference can be moved through 32 positions representing the (32) storage cells. This gives access to data mapped to the r.a.m. at intervals of 100 μs, equivalent to 8 sample intervals (assuming the mapping was originally to contiguous locations).

The data from each 'tap' are passed through a 7-bit inhibitor/complementer to one side of a full adder array. The most significant (sign bit) of the sample is propagated left to stretch the 7-bit sample to 12 bits. This makes it ready for insertion into the 12-bit accumulator.

The panel switches are sampled with a pair of 16 input multiplexors giving a total fan-in of 32. The address for the multiplexors is provided by a 5-bit counter which is incremented in unison with the r.a.m. address count (most-significant portion).

To effect a moving point of reference, the r.a.m. address constantly 'wraps-around' from (2^8-1), i.e.

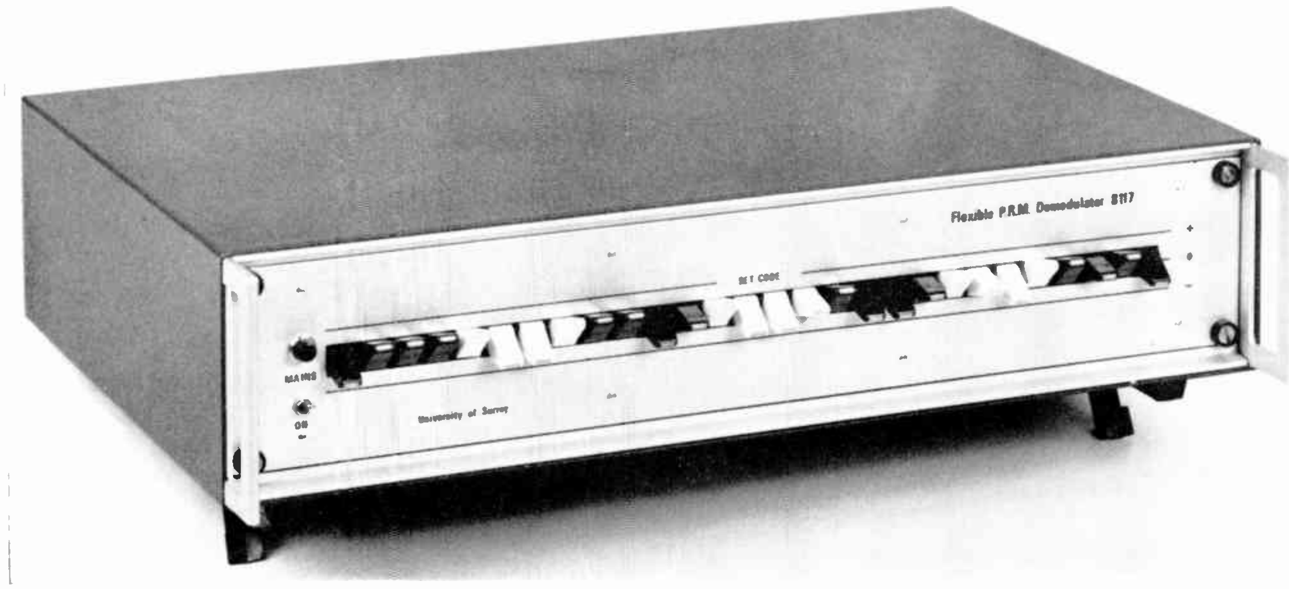


Fig. 5. Prototype flexible p.r.m. correlator.

255 to 0. In order that address generation stays correct, in particular the sampling of the panel switches with respect to the currently sampled tapping point, the 5-bit portion of the r.a.m. address generator must be incremented, not only by the mechanism which inspects each tap (as described), but also by the carries emerging from the least significant portion of the r.a.m. address generator (3 bits). However, the switch multiplexor address is only incremented by the 'tap' inspection mechanism. The addressing sequence is shown in Table 1. At the end of each machine cycle (acquisition of one data sample plus weighted summation of the tap outputs), the accumulator is sampled and cleared for the next cycle. Sampling here consists of strobing the accumulator's contents into a digital-to-analogue converter (d.a.c.) buffer register, ready for display.

A programmable inverter (array of exclusive-OR gates) is interposed between the accumulator and the d.a.c. buffer so that, if desired, the modulus of the correlator output can be obtained. This represents a reasonable approximation to full-wave rectification, being only one quantum in error in the least significant bit of the output when a negative quantity emerges from the accumulator.

Digital-to-analogue conversion is restricted to 8 (the most significant) bits of the accumulator output so as to be about compatible with the precision of the original digitized data.

8 Construction and Performance

A prototype machine has been constructed in the Department of Electronic and Electrical Engineering at the University of Surrey (Fig. 5). Trials have been carried out using a synthesized test sequence of up to 28 bits. The peak-to-sidelobe ratios that have been achieved are as expected from theory. A typical prefilter 28-bit correlated output is shown in Fig. 6.

Correlators of this type can find obvious application in intrapulse modulated radar systems as well as in

Table 1. Random access memory addressing sequence

Sample number	R.a.m. address for reading	Switch multiplexor address	R.a.m. address for writing
0	8, 16, ..., 224	0, ..., 27	0
1	9, ..., 225	0, ..., 27	1

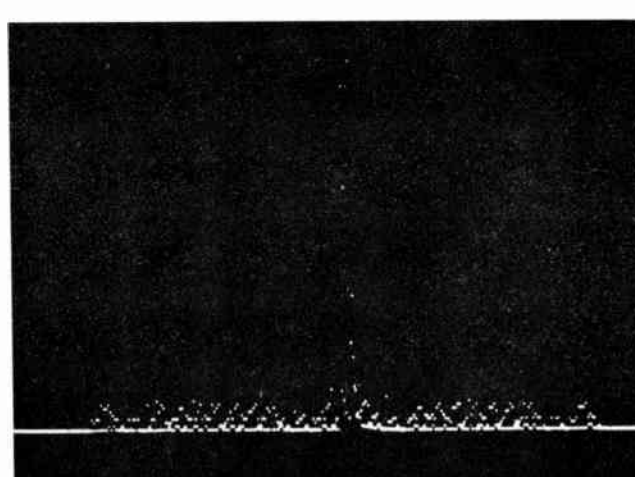


Fig. 6. Typical prefilter 28-bit correlated output.

ionospheric sounding for h.f. communications (for which it was primarily designed).

9 Acknowledgments

We wish to acknowledge the invaluable support of the Electronics Workshop of the Department of Electronic and Electrical Engineering, University of Surrey, without which the programme could not have been completed in the time scale.

The work has been carried out with the support of the Procurement Executive, Ministry of Defence.

10 References

1. Benjamin, R., 'Modulation, Resolution and Signal Processing in Radar, Sonar and Related Systems,' Chaps. 1, 3, 5. (Pergamon, Oxford, 1966.)
2. Taub, H. and Schilling, D. L., 'Principles of Communication Systems' pp. 31–6. (McGraw-Hill, New York, 1971.)
3. Gordon, B. M. and Seaver, W. H., 'Insight into sampled data theories', *Systems (G.B.)*, 2, No. 1, pp. 30–4, January 1974.

Manuscript first received by the Institution on 20th February 1975 and in revised form on 23rd June 1975. (Paper No. 1709/CC 244).

IMEKO Congress comes to London

The 7th Congress of the International Measurement Confederation—IMEKO—opens at the Royal Lancaster Hotel, London, on 10th May. To be officially opened by the Parliamentary Under Secretary at the Department of Industry, Mr. N. G. Carmichael, M.P., the Congress has the theme of 'Practical Measurement for Improving Efficiency', and lasts until 14th May. It is being organized by the Institute of Measurement and Control, and co-sponsors in the UK are the Chemical Society, the Institute of Petroleum, the Institute of Physics, the Institution of Chemical Engineers, the Institution of Electrical Engineers, the Institution of Electronic and Radio Engineers, the Institution of Mechanical Engineers, Measurement Control and Automation Conference, the Metals Society, the Society of Chemical Industry and the United Kingdom Automatic Control Council.

Recognizing that measurement is fundamental to all scientific and technological endeavour, the organizing committee hopes that the Congress will emphasize its value, increase effectiveness and promote the efficient use of resources in industry. The general economic significance of measurement in both the developed and developing world is an additional aspect which, it is hoped, will run through discussions at the Congress.

In addition to Mr. Carmichael, other speakers at the opening session will include Sir Ieuan Maddock, Chief Scientist at the Department of Industry, and the President of the International Committee of Weights and Measures, Dr. J. V. Dunworth.

Some 180 papers from over 20 countries have been selected from over 300 offered—about 20% have come from the UK. Papers will be presented in parallel technical sessions, falling into three general categories; Industrial Application of Measurement (Food manufacture and processing (including agriculture), Chemical/petroleum, Quality control, Transport and traffic, Environmental monitoring and protection, Machine tools, Metals, Energy, papers, Plastics, Textiles); Measured Variables (Electrical quantities, Properties of materials and components, Length, displacement and velocity, Composition control, Optical methods of dimension measurement, New techniques of flow measurement, Correlation methods of flow measurement, Force and mass, Pressure, Temperature, Transducers and sensors); Measurement Theory and Fundamentals (Fundamental mathematics—modelling of measuring systems, Errors and accuracies, Signals and identification, Signal handling equipment and application). These sessions will run from mid-morning until mid-afternoon. From mid to late afternoon there will also be round-table sessions for more informal discussion.

The first part of each morning will be devoted to Plenary Sessions at which five specially invited Survey Lectures will be presented on subjects of general interest by the following internationally recognized experts:

Measurement in Quality Control, by Prof. H. Trumpold (GDR)

Visual Displays and Man/Machine Interfaces, by Prof. F. P. Lees (UK)

New Problems of Instrumention Set by the Control of Large Systems, by Z. Binder (France)

Statistical Methods in Measurement, by Prof. G. I. Kavalerov, Prof. E. I. Tsvetkov and Prof. S. M. Mandel'shtam (USSR)

The Contribution of Systems Theory and Control Engineering to Instrument Science, by Prof. F. Mesch (FRG)

There are over 20 countries in IMEKO and the triennial Congress rotates around the membership. Thus it is unlikely that UK instrumentation specialists will have another opportunity of attending an IMEKO Congress in this country for some considerable time. The Registration Fee at £75.00 includes copies of all papers from Technical sessions. Full details of the Congress together with Registration Form, are available from The Institute of Measurement and Control, 20 Peel Street, London. W8. (Tel. 01-727 3755).

IERE News and Commentary

The Golden Jubilee Convention

The theme of the Golden Jubilee Convention, to be held at the University of Cambridge from 28th June to 1st July 1976, will be 'Electronics in Society'.

Addresses by distinguished speakers and papers from senior electronic engineers will be given in the course of the programme, and technical sessions include: Communications; Cognitive Science and Signal Processing; Microelectronics; Medical Electronics; Automation and Control; Advances in Data Processing.

The Ninth Clerk Maxwell Memorial Lecture will be given in the Cavendish Laboratory on Wednesday, 30th June, and the Convention Banquet will be held on the following evening in King's College.

Further information will be published in the Journal shortly and requests for registration forms should be sent to the Conference Department, IERE, 8–9 Bedford Square, London WC1B 3RG (telephone 01-637 2771).

New Charter for Professional Engineers

A Supplemental Charter drafted by the Council of Engineering Institutions to give the engineering profession a more effective voice is being circulated for discussion to the United Kingdom's 175,000 chartered engineers. Copies are enclosed in all copies of this issue of the Journal sent to IERE members' addresses in the British Isles.

Members are asked to let the CEI have their views by April 30th so that the final draft Supplemental Charter and Bye-Laws when finally approved can be submitted for Privy Council approval by about the middle of this year.

In an accompanying letter, Professor John Coales, immediate past chairman of the CEI, says, '. . . it has become clear from the lively discussion meetings that have been held in many local centres that you as individual engineers want to feel more closely associated with CEI . . . and better represented on the Council. This is therefore the goal which the Board of CEI is seeking to achieve by a quite fundamental change in the structure of CEI.'

Previously, the engineer could express his views in CEI only through his institution representative. The most significant proposed change is that all chartered engineers should be individual members of CEI. Their representatives on the Council would be elected directly by individual chartered engineers through their institutions. It has yet to be finally decided whether the larger institutions should have more than one elected member, although this is included in the draft.

To improve communications between the industrial engineers in the regions and CEI, a Regional Affairs Committee has already been set up. A further change is that non-chartered engineering institutions will be able to become affiliates of CEI and their members, if suitably qualified, will be able to be registered as chartered engineers.

It is possible that a new name may be adopted for the Council of Engineering Institutions. Whilst a decision has yet to be made, the draft Supplemental Charter and By-Laws have used the name 'The Chartered Engineering Institute'.

Submissions Invited for 1976 MacRobert Award

Submissions for the 1976 MacRobert Award—the major engineering award in Great Britain—are being invited and entries should reach the MacRobert Award Office at the Council of Engineering Institutions by 30th April 1976.

Consisting of a gold medal and a cash prize of £25,000, the Award is presented annually by CEI on behalf of the MacRobert Trusts. It recognizes an outstanding innovation in the fields of engineering or the other physical technologies or in the application sciences which has enhanced or will enhance the national prestige and prosperity of the United Kingdom.

The rules allow the prize to be made to an individual, to an independent team, or to a team working for a firm, organization or laboratory—but no team must exceed five persons. The prize may be divided but between no more than two innovations. Submissions for the Award should be sent to: The MacRobert Award Office, CEI, 2 Little Smith Street, Westminster, London SW1P 3DL (Tel.: 01-799 3912). Copies of the rules and conditions can be obtained from the same address.

New Academic Qualifications in Australia

The Institution of Engineers, Australia, has adopted a new policy which will up-grade the academic qualifications required to become a member of the Association of Professional Engineers of Australia (APEA) and practise engineering. It is the Institution which sets the basis for employment in the profession of engineering in Australia, and APEA uses this same basis for acceptance. Those who do not meet the minimum standards of the Institution are not accepted by APEA and are not considered as professional engineers.

The new rule of the Institution, effective from 1980, will require that people who qualify after that date complete a four-year full-time course (or equivalent) to be eligible for admission to Corporate membership and thus be recognized in Australia as professional engineers. The IE Aust. has judged that by that time the interest of the community will require newly qualified engineers to have completed four years of study to be capable of performing professional engineering studies.

There is a transition period up to 1985 allowing admission to Graduate Membership for those passing three-year full-time (or equivalent) courses, but such people must also complete a bridging qualification for admission as Corporate Members. If they have not done so by 1985, they will cease to be Graduate Members.

IERE Member on CEI Board

Following the reorganization of the CEI Board whereby each Institution has only one member, the representative of the IERE is Professor W. A. Gambling (Vice-President); the alternate is Professor W. Gosling (formerly a Vice-President).

Nearly 1000 Companies for IEA/Electrex

With participating companies now approaching the 1000 mark and nearly 21,000 square metres of stand space booked, the International Electrical, Electronic and Instrument Exhibition at the National Exhibition Centre, Birmingham, from 3rd to 7th May 1976, looks like amply justifying the faith of its sponsors and organizers.

A natural combination of two well established events—the Instruments, Electronics and Automation (IEA) Exhibition and the International Electrical Exhibition (Electrex)—this will be one of the first events to be held in the new Exhibition complex. Adjoining Elmdon Airport and on the main railway line a few miles south east of Birmingham, this was officially opened by HM the Queen on 2nd February.

Particularly for the convenience of overseas visitors to IEA/Electrex, the 7th IMEKO Congress has been scheduled to take place in London during the following week and linking travel arrangements have been organized. Details of IMEKO are given on page 135.

Conference on Video and Data Recording

As one of the major events of its Golden Jubilee Year, the Institution is holding its second Conference on Video and Data Recording (the first was in 1973). This will take place at the University of Birmingham from 20th to 22nd July 1976, and associated in the planning have been the Institution of Electrical Engineers, the Institute of Physics, the Institute of Electrical and Electronics Engineers, the Royal Television Society and the Society of Motion Picture and Television Engineers.

Papers will be presented in the following sessions:

- Theory of Recording Processes
- New Recording Techniques and Media
- Coding and Modulation
- Video and Computers
- Instrumentation
- Recording and Replay Heads
- Transcription and Testing

The programme will be supported by an associated exhibition of equipment.

Accommodation will be available in the University's Halls of Residence and further information and registration forms may be obtained from the Conference Department, IERE, 8–9 Bedford Square, London WC1B 3RG (Telephone 01-637 2771).

IEE Conference Publications

Members of the IERE may purchase single copies at reduced rates of the following volumes of papers read at conferences for which the Institution was a co-sponsor:

'Antennas for aircraft and spacecraft', London, June 1975 (IEE Conference Publication 128), £6.00 (normal price £9.00).

'Dielectric materials measurements and applications', Cambridge, July 1975 (IEE Conference Publication 129), £7.80 (normal price £11.90).

'Telecommunication transmission', London, September 1975 (IEE Conference Publication 131), £5.20 (normal price £8.00).

'Optical fibre communication', London, September 1975 (IEE Conference Publication 132), £5.60 (normal price £8.50).

Orders from members wishing to take advantage of these special rates should be placed through the IERE Publications Sales Department, 8–9 Bedford Square, London WC1B 3RG.

Index to Volume 45

The Index for the 1975 volume of *The Radio and Electronic Engineer* (comprising title page and principal contents list, subject index and index of persons) has been published and copies are being sent automatically to all subscribers to the Journal with this issue. Members wishing to obtain the Index for binding with their Journals or for reference will be sent a copy free of charge on application to the Publications Sales Department, IERE, 9 Bedford Square, London WC1B 3RG, by letter, or telephone 01-637 2771, extension 19.

Indexes are included in all bound volumes of the Journal and members who propose to send their 1975 issues for binding need not apply for a copy of the Index beforehand.

CEI and Personal Services

The CEI has set up a Working Party representing all the Chartered Engineering Institutions to discuss the personal services which each Institution might give to its members. Mr. W. B. Hopkins (Fellow) has been appointed to represent the Institution on this working Party.

SRC Regional Meetings on Postgraduate Training and Research in Universities

The Science Research Council in conjunction with the Confederation of British Industry, Council of Engineering Institutions, Council of Scientific and Technical Institutions and the Committee of Vice-Chancellors and Principals (CVCP) is planning to hold a series of five regional meetings in the period April/May 1976; the venues are to be Cardiff, Birmingham, Glasgow, Leeds and London. The future SRC policy for support of postgraduate training will be discussed in the light of the two recent SRC reports* and the interim report of the CVCP working group on postgraduate training. Also for discussion will be the new SRC methods for support of research in universities and polytechnics as described in the SRC Annual Report for 1974/5.

Universities and polytechnics are being invited to nominate representatives to attend the meetings but any non-academic members interested in attending may obtain details and tickets from the Secretary, CEI, 2 Little Smith Street, London SW1P 3DL, or from the appropriate CEI Regional Committee Secretary whose address may be obtained from the Secretary of your local IERE Section. Final arrangements for these meetings will be announced as soon as possible.

The British Institute of Non-Destructive Testing

On 1st January 1976, the Society of Non-Destructive Examination (SONDE) and the Non-Destructive Testing Society of Great Britain (NDTS) amalgamated to form The British Institute of Non-Destructive Testing. This new Institute will represent both NDT engineers and the NDT industry throughout the United Kingdom, as one body. Existing members of both Societies are being transferred to the appropriate grades of membership of the new Institute.

The first President of the Institute is Mr. K. F. Williams. Mr. I. M. Barnes is Secretary of the Institute and its offices are at 53–55 London Road, Southend-on-Sea, Essex, SS1 1PF (telephone: (0702) 354252/3).

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* The report of the Working Party on Postgraduate Training (Chairman, Professor Sir Sam Edwards, F.R.S.) and the Second Report of the Joint SRC/SSRC Committee on Interdisciplinary Postgraduate Education (Chairman, Lord Ashby, F.R.S.).

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

BENEVOLENT FUND

INCOME AND EXPENDITURE ACCOUNT FOR THE YEAR ENDED 31st MARCH, 1975

1974 £	£		£	£
297		Grants and Donations		543
100		Purchase of Bursaries		100
74		Postage, Telephone and Stationery		88
3,029		Surplus for the year carried to Reserve Account		3,221
£3,500				£3,952

1974 £	£		£	£
791		Subscriptions and Donations		777
2,709		Dividends and Interest on Investments		3,175
£3,500				£3,952

BALANCE SHEET AS AT 31st MARCH, 1975

		Reserve Account		
	45,886	Balance as at 1st April 1974	48,960	
		Add: Surplus on Income and Expenditure Account		
	3,029	for the year	3,220	
		Surplus on Redemption of Investments		
	45	(Net)	4	
48,960				52,184
		Sundry Creditors		
		Amount due to the Institution of Electronic and Radio Engineers		
—				83
£48,960				£52,267

46,038		**Investments at Cost**		48,933
		(Market value as at 31st March 1975 £39,700 (1974 £33,869))		
		Current Assets		
	2,550	Bank Balances on Current and Deposit Accounts		2,945
	282	Income Tax repayment claim	310	
	79	Sundry Debtors	79	
		Amount due from the Institution of Electronic		
	11	and Radio Engineers	—	
2,922				3,334
£48,960				£52,267

For Trustees
Signed
IEUAN MADDOCK (*Chairman*)
G. D. CLIFFORD (*Honorary Secretary*)
S. R. WILKINS (*Honorary Treasurer*)

AUDITOR'S REPORT TO THE TRUSTEES OF THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS BENEVOLENT FUND

In my opinion the above accounts show a true and fair view of the Benevolent Fund's affairs at 31st March 1975 and of the surplus for the year ended on that date, and comply with the Rules thereof.

50 Bloomsbury Street,
London WC1B 3QT.
5th August, 1975

Signed: R. O. O. FREEMAN
Chartered Accountant
Honorary Auditor

Report of the IERE Benevolent Fund Trustees for 1974-75

It is 25 years since the Benevolent Fund of the Institution was started. Due to the generosity of members, the Trustees have never had to deny help to a deserving case; moreover, as shown in the Balance Sheet, the support that has been given in the past has enabled the Fund to have a capital sum of over £52,000, the revenue from which now considerably helps offset the reduction in the number of subscribers during the past two years. Current lack of support is no doubt attributable to inflation, and it is as well that the Fund is in possession of assets.

Disbursements during the year were, for the most part, given to help in the support of the widows and families of deceased members, and particularly in the education of their children. The education bursaries which have been purchased in the past have been fully used and the Trustees have also collaborated with other funds in giving assistance to deserving cases.

Members are earnestly requested to consider whether they can make a contribution, however modest, to this their own Benevolent Fund, and the most beneficial way they can do this is by executing a simple Deed of Covenant undertaking to contribute a fixed sum each year for at least seven years. This enables the Fund to recover income tax at the standard rate (at present 35%) from Inland Revenue on the sum contributed. A form of Deed for this purpose is available from the Honorary Secretary to the Trustees at 9 Bedford Square, London WC1B 3RG.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

BENEVOLENT FUND

SCHEDULE OF INVESTMENTS AS AT 31ST MARCH, 1975

Nominal		*Cost* £
£100	Associated Electrical Industries Ltd. 6% Debenture Stock 1978/83	99
£1,000	Barnet Corporation 7¾% 1982/84	982
112	Bowater Paper Corporation Ltd. Ordinary Stock	322
£2,000	British Transport Government Stock 1972/77	1,864
£170	Burmah Oil Co. Ltd. Ordinary Stock	702
£800	Commonwealth of Australia 5½% Loan Stock 1977/80	788
1,200	Henderson Industrial Trust	150
£4,000	Consolidated 4% Stock	3,527
340	Courtaulds Ltd. 25p Ordinary Shares	500
£500	Courtaulds Ltd. 7% Debenture Stock 1982/87	467
750	Currys Ltd. 25p Ordinary Shares	1,580
400	De La Rue Co. Ltd. Ordinary Shares of 50p	656
1,600	E.M.I. Ltd. 50p Ordinary Stock Units	2,003
£210	E.M.I. Ltd. 8½% Convertible Unsecured Loan Stock 1981	210
320	English China Clays Ltd. 25p Ordinary Shares	501
£2,375·46	Funding Stock 5¼% 1978/80	2,000
£2,225	Funding Stock 5½% 1982/84	1,994
£2,250	Funding Stock 6½% 1985/87	2,007
£1,830	Funding Loan 6% 1993	1,001
2,249	G.E.C. Ltd. 25p Ordinary Shares	885
332	Grattan Warehouses Ltd. 25p Ordinary Stock Units	622
£200	Great Universal Stores Ltd. 5⅞% Redeemable Unsecured Loan Stock	106
£300	Great Universal Stores Ltd. 8¼% Unsecured Loan Stock 1993/95	248
	Carried forward	£23,214

Nominal		*Cost* £
	Brought forward	23,214
£150	I.C.I. Ltd. 7¼% Unsecured Loan Stock 1986/91	148
1,000	I.C.I. Ltd. Ordinary Stock Units of £1	2,925
£1,000	Islington Corporation 10% Redeemable Stock 1982/3	995
£100	London County 6% Loan Stock 1975/78	99
531	Lonrho Ltd. 25p Ordinary Shares	690
1,000	Marks & Spencer Ltd. 25p Ordinary Shares	1,772
£4,000	New Zealand 7¼% Stock 1977	3,972
1,000	Plessey Co. Ltd. 50p Ordinary Shares	1,329
156	Reed Group Ltd. £1 Ordinary Shares	392
£35	Reed Group Ltd. 10% Unsecured Loan Stock 2004/09	35
1,000	Shell Transport and Trading Co. Ltd. 25p Ordinary Shares	3,140
£1,000	Slough Corporation 8¾% Redeemable Stock 1979/80	990
£200	Tanganyika Concession Ltd. 50p Ordinary Stock Units	703
1,210	Transport Developments Group Ltd. 25p Ordinary Shares	743
£1,225	Treasury Stock 7¾% 1985/88	992
£1,320	Treasury Stock 8¼% 1987/90	1,005
£1,990·69	Treasury Stock 8½% 1980/82	2,000
£1,125	Treasury Stock 9% 1994	999
£1,500	Treasury Stock 12% 1983	1,433
796	Unicorn General Trust	125
£1,200	3½% War Loan	498
400	Watts, Blake, Bearne & Co. Ltd. 25p Ordinary Shares	437
500	Whitbread & Co. Ltd. 25p 'A' Ordinary Shares	297
		£48,933

Letters

From: **I. M. Joyce,** C.Eng., M.I.E.R.E., **G. A. Cooper,** C.Eng., M.I.E.R.E., **R. Parsons,** C.Eng., M.I.E.R.E., **G. Ryder,** C.Eng., M.I.E.R.E. and **R. L. Thompson,** C.Eng., M.I.E.R.E.

Some Views of the Institution

Those members who pay subscriptions by the direct debit system and missed the announcement of increased subscriptions on p.692 of the November Journal will by now have received the Deputy Finance Secretary's letter requesting authorization for a 25% increase.

This call for increased subscriptions, when read in conjunction with the gloomy editorial of the December Journal, surely made some members wonder whether their continued membership was achieving anything. There is disenchantment within the profession, much of it directed against the Engineering Institutions whom, by their rigid adherence to learned society activities, have failed to achieve the status and power in the community for their members that the older professions, such as Law and Medicine, have achieved. It is surprising that a suspicion of this disenchantment has reached the Editor of the Journal, since the affairs of the Institution do not appear to be structured with feedback in mind.

It is absurd that a compulsory subject entitled 'The Engineer in Society' should be added to the 'would-be' engineer's examination load with no provision for discussion of this subject after graduation. Surely to this end local sections should have regular meetings, closed to the public, to discuss matters concerning the Professional Engineer and the services the Institute should provide.

For example, many members may have sympathy with the sentiments expressed by P. J. Hewitt and A. F. Willis in their letters, in the April Journal. Regular meetings with a free exchange of views on these and other subjects, far too numerous to mention here, would at least give members a feeling of belonging to an organization which is concerned about the profession (i.e. the members) rather than a publishing organization designed to impress the estimated 500 genuine subscribers to the Journal with the intellectual abilities of the authors.

With regard to the increased subscriptions, approved without question by the 74 corporate members present at the A.G.M., i.e. less than 1% of the corporate membership, most private individuals and organizations consider ways of cutting costs in these inflationary times. One obvious target is the cost of publishing and posting the Journal, this forming 25% of the Institution's budget. Is it not time that the practice of other institutions is adopted whereby members receive a news sheet automatically but the Journal is only sent on receipt of an additional subscription?

Another target is postal charges. Mr. Tyler's letter was sent almost simultaneously under separate cover to the local section meeting list. If this happened for all 15,000 members the cost of postage, not counting the envelopes, amounts to a staggering £975. It is interesting to note that the Executive Committee has had its proposal that freehold premises outside London be acquired, approved. Let us hope that this takes place soon in view of the rent and rates increases forecast of around £34,000 a year.

The actual costs of the library service are not itemized in the budget, but surely a service which only 400 readers used, i.e. 2·6% of the membership, and which is duplicated by the Public Library Service, Universities, Colleges and many Companies, must be examined in the light of increasing costs.

In conclusion, will members who have sympathy for some or all of the views expressed in this letter please stand up and be counted so that the Council knows what the members it represents really think, rather than assuming, as they probably do now, that all is well.

79 Fir Copse Road,
Purbrook, Hants.
PO7 5HY
27th January 1976

I. Joyce
G. A. Cooper
R. Parsons
G. Ryder
R. L. Thompson

In reply:

If engineers are to achieve professional status which has parity in the public mind with that of lawyers and doctors, they must concern themselves with the standards of competence and the means of maintaining professional ability. Should Mr. Joyce and his colleagues wish to discuss the place of the engineer in society they have a very active Local Section whose Committee would undoubtedly welcome help in the arrangement of meetings on this and other relevant subjects. The columns of the Journal, too, contrary to what the writers appear to believe, are not closed to the expression of constructive comment on the Institution's work. Thus one may reasonably paraphrase the late President Kennedy: Ask not what your Institution can do for you, but rather what you can do for your Institution. Already the Institution has provided the five Members, and thousands of others, with a recognized professional qualification and a continuing means of contributing to the standing of their profession.

It is a well known and regrettable fact of life that the attendance at the annual general meeting of almost any organization is but a small percentage of the eligible membership. A journey to London would of course have been necessary but, having made that, as Corporate Members the writers could have asked their questions and received direct answers from the officers of the Institution.

The publishing practice referred to has certainly been considered in the past and, with its variants, is constantly kept in mind. So far it has been concluded that the Institution's role of advancing radio and electronic engineering, and thereby the professional interests of its members, is fulfilled most effectively by the present style of the Journal, which incidentally also demonstrates to industry and others the potential intellectual ability of members—not only of authors. It should be remembered that many members, and not only those overseas, are not in the fortunate position of the five Members from Hampshire and do not have ready access to technical libraries and information services to keep them in touch with the developments in their branch of learning.

Turning to the sums which have been done in the letter, the Deputy Secretary's communication in fact went only to the relatively small number of members in the United Kingdom who at present pay their subscriptions by the Direct Debit system; the enclosure of such a restricted communication with Local Section meeting programmes or even with the Journal posting would have clearly presented a data handling problem of some complexity—and expense.

As to the cost of the Institution's Library, this is indeed difficult to quantify but the existence of a professional, learned institution without its own library would be unthinkable to the majority of members whether or not they need frequently to call upon its specialized resources. On a statistical point, 400 borrowers represents more than 2·6% of the potential users since books are not on loan to members abroad; furthermore it takes no account of those using the library by personal call, by telephone or through the post.

(Continued at foot of facing page)

Colloquium Report

'Wedding Calculators to Instruments'

Organized by the Automation and Control System Group and held in London on 12th November 1975

The Chairman, Mr. M. S. Birkin, opened the colloquium, which was attended by over 100 delegates, by stressing how developments in electronics had given us desk-top programmable calculator systems of quite phenomenal power. He went on to say that it was now possible to have systems which could be readily programmed by scientist and engineers without the need for them to have to resort to the employment of specialist computer programmers. He added that the level to which programmable calculators had now been developed was such that their power was well in excess of many of the earlier conventional computer systems.

The programme started with a paper by Mr. R. Tinson (Wang Electronics) on the 'Development Philosophy of the Programmable Calculator'. Mr. Tinson traced the development of programmable calculators and made particular emphasis of the input and output facilities of these machines. In particular he described how the introduction of the use of BASIC language in calculators gives a versatility in operation not possible on purely algebraic language machines.

The next speaker, Mr. S. M. Jaeckel (HATRA) described in his paper 'The HATRA Instrumental Colour Pass/Fail System using the Harrison/Shirley Digital Colormeter'. Mr. Jaeckel gave an extremely interesting description of how a Wang calculator had been interacted with a digital colorimeter and an output writer to give an automatic means of colour matching with accurately quantifiable and repeatable results.

This was followed by a paper on the 'P652 Microcomputer System' by Mr. D. Aggleton (British Olivetti) who described the functional working of the system. This was followed by Mr. D. Moon (Metal Trim) who described in a paper entitled 'Applications of a Microcomputer to Control Production Equipment' a practical application of a programmable calculator to control directly the position of metal purlins and a punch press by data output from the calculator to a pneumatic control system. A random access extended memory had been added to the system to enable every required combination of bolt hole pattern to be stored. The system was operated by a press operator, the only input being the customer's dimensions through the numeric entry key board. The system had been extremely satisfactory in operation and had saved its initial cost many times over.

During the lunch break those attending could take the opportunity to examine working calculator-controlled systems in an exhibition arranged in the Institution Library. The firms exhibiting calculators were Hewlett Packard, Wang, Tektronix, British Olivetti and Monroe International.

The first contribution after the lunch interval was by Mr. W. Davies (Monroe International) and Mr. N. Challacombe (Keithley Instruments) who described how a Monroe Calculator could be interfaced to the range of instruments produced by Keithley.

The presentation following was by Mr. H. Rippener (Tektronix U.K.) who spoke on 'The Use of Calculators in the Automation of Data Acquisition'. He described how rapid strides had been made in using programmable calculators for data acquisition and said that in many application areas where data logging of a small number of variables was required, the calculator-based data acquisition system was more attractive in terms of ease of use and cost than its counterpart the mini-computer. Mr. R. Brown (Smiths Industries) then described in his paper 'Calculator-controlled Testing of Rate Gyroscopes' an application of a Tektronix 31 calculator and a Solartron 1177 frequency response analyser to make automatic measurements of rate gyroscope parameters.

The final paper was by Mr. D. Peacock (Hewlett Packard) on the application of the 'Hewlett Packard Interface Bus—A New Method of Interfacing Instrumentation' in which he described how there is a need for a standard interface to interconnect instruments, peripherals and computers or calculators, in order to simplify the building of systems. The development of the Hewlett Packard Interface Bus (HP-IB) has resulted from this requirement, and he said that the International Electrotechnical Commission (IEC) has adopted the HP-IB as a draft recommendation for a new international standard. He then described in detail how the HP-IB worked and the versatility of the system for implementing a wide variety of test equipment.

Letters (*continued from facing page*)

In spite of what has been suggested, the Council, its Committees and the Institution's staff do between them meet and talk with a large cross-section of the membership and from all these meetings policy is evolved to serve the vast majority of members.

I repeat that Council welcomes comments which help to advance the objects of the Institution and therefore the interests of its members. *Editor*

From: **T. M. Ball,** M.Inst.P.S. C.Eng., M.I.E.R.E.

The Institutions and the Government

It was interesting to read in the January issue of the Journal that the CEI's Chairman had written to the Prime Minister offering the help of professional engineers in reviving British industry. One can but support this offer—however the Engineering Institutions have other colleagues who would support this action if so requested, since many Chartered Engineers are, like myself, members of other professional bodies and as such hold influential positions in industry which can be used to support this revival. Perhaps Professor Coales should approach such Institutes as Purchasing and Marketing for their support.

Unfortunately his call to the Government to protect the independence of the professional engineer may well fall on deaf ears, since many of us who are, for instance, in the aircraft industry, are having to form ourselves into a union to protect both ourselves and our industry's future which is under the threat of nationalization.

17 Rivercourt Road,
Hammersmith,
London W6 9LD
12th February 1976

T. Michael Ball

The Electronics Industry of South Korea

One of the most remarkable success stories of South Korea's industrialization has been the electronics industry. Beginning in 1958 with the assembly of radios from imported components, it now employs more than 100,000 men and women in about 400 factories.

It was not until 1966, however, that world manufacturers began to appreciate the advantages of using Korea's abundance of cheap, diligent labour to solve the growing cost difficulties of such a labour-intensive industry.

First into the field were three American companies—Motorola, Signetics and Fairchild. That small beginning soon became a flood, with the Japanese leading the way. Today there are 168 Japanese electronic companies operating in Korea through 139 joint ventures and 29 wholly Japanese. The United States comes a poor second with 27, of which 14 are wholly owned. However, the American operations tend to be larger. In addition, there is a Dutch, a West German and three Hong Kong companies.

As an indication of the extent to which foreign capital has poured into Korean electronics the Ministry of Commerce and industry reported earlier this year that of the $724M invested by foreign enterprises in Korean domestic industry up to December 1974 no less than 17% went into the electronics sector.

With a growth rate of nearly 60% annually for the past six years—and the bulk of production going direct to hard currency overseas markets—it is hardly surprising that the industry accounts for nearly 6% of the nation's total G.N.P.

Development Plan for the Industry

Such rapid development owes much to the aggressive support provided by the South Korean Government. In 1966 electronics was designated as a strategic export industry and provided with various financial aids and preferential tax treatment. Early production had been confined to radio sets, telephones and similar low technology products. Because of a shortage of technical knowledge the work was mainly achieved by domestic companies entering into licensing agreements with foreign manufacturers.

During the first five-year development plan, however, electronics began to take on a new importance and in 1965 domestic production of record players began. That was followed a year later by the assembly of television sets.

The industry really began to develop in the late 1960s with the manufacture of semiconductors such as transistors, integrated circuits and components for the booming computer industries of the world.

Since then tape recorders, desk and hand-held calculators and colour television sets have extended the wide range of Korean-made electronic products finding their way to world markets, often carrying the brand name of their Japanese and United States partners. Today about 80% of output is sold in more than 50 countries.

From a total production of $22M worth in 1966 the industry reached $734M in 1974. In the same period exports grew from $4M, representing 18% of production, to $553M, or 68% of the total.

The production target for 1975 was $780M, of which about $620M worth was earmarked for export. But with demand in Japan and the United States down by about 28% and 14% respectively, that will take some achieving. To stimulate orders from depressed markets export prices have been lowered by amounts varying from 2% on radios to nearly 30% on calculators.

In spite of slackening demand, the Government is pressing ahead with expansion plans aimed at a six-fold increase in the value of production by 1981. However, there will be a small drop in the proportion exported—from 68% in 1974 to 52% in 1981. That is planned to take into account the growth of domestic demand as Korean wages increase and local materials are substituted for imports.

The expansion plan calls for a further 560 factories to be built at a cost of $1,081M of which it is hoped $530M will be foreign investment.

Incentives for Foreign Companies

In October 1973 the Government established an industrial estate at Gumi in Gyeongbug province, near the Seoul–Pusan motorway, exclusively for electronic and allied industries. In addition to the provision of normal services for industrial estate users, a wide range of incentives have been offered to foreign investors.

They have been exempted from a range of income, corporate and property taxes for the first five years and receive a 50% reduction for the next three years. The remittance of profits and dividends is guaranteed from the first year and the remittance of principal foreign investment is guaranteed after two years of business operation.

It is not surprising that within a very short time 60 companies were in occupation with room for a further 250 on the 2,600-acre site. By 1977–78 exports from the estate could be earning $400M a year.

Low Labour Costs

The key attraction from the beginning has been the low wages paid to everyone from a skilled engineer to a labourer. A new study produced by the Korean Ministry of Commerce and Industry, 'Investment Opportunities for the Electronic Industry', contains this significant statement:

'At the present time Korea has no legal minimum wage system. Basic wages comprise only a portion of workers' income and vary with the size of companies' or personal ability. The monthly starting wage is now $130 for engineers, $80 for technicians, $30 for workers and $90 for clerical or administrative workers.

'Employers pay one to four months bonus annually to employees and serve usually one hot meal a shift, but this is not legally required. Transportation compensation is optional, depending on factory location'.

The minimum age for starting work is 13, although employers are required to keep a register of workers under 18 and must obtain parental approval in writing.

The standard working day in the electronics industry is eight hours a day for six days a week. But that can, and often is, extended to 60 hours a week or more.

Good attendance, frequently interpreted as 100% attendance, must be rewarded by an extra eight days leave with pay a year. Workers who achieve 90% lose five of the eight days. Legal holiday entitlements are a further 16 days a year.

Clifford Webb

Reprinted from The Times *of 26th September 1975 by permission of the Editor.*

For how much longer can the higher paid take the brunt of wage restraint?

Everyone in industry is closely following reports in the press of the discussions between the Government and the trade unions about pay policy after the ending of the £6 limit. A limit based on a cash figure, as recommended by Mr. Jack Jones, seems still to be very much on the cards.

From the point of view of managers and other more highly paid workers, the implications of another round of incomes policy on this basis would be very serious. The table below shows how rapid would be the further decline in differentials, after tax, between different income bands—indicating a further large fall in real standards of living for managers and highly skilled men. The assumption made is of a 4% increase with a maximum £10 limit and a minimum 'safety net' of £2.40.

Obviously, we do not know what the tax rates will be for 1976–77, and the table is based on current rates. Between August 1975 and August 1977, a labourer will receive a 16·1% increase in his pay packet, while managers (£15,000–£30,000) will gain by only 0·8 to 1·8%. Whatever the rate of inflation in this period, these managers will have virtually no means of protecting themselves against the continuing fall in the real value of their take-home pay.

Almost everyone would agree that for a brief period of crisis it was fair and sensible to allow the better-off to carry an even heavier share of tax burden than would be the case in normal times. But discrimination against the higher paid has been a strong feature of incomes and taxation policy for so long that it constitutes an attack.

The effects are alarming, and it will require positive action to contain, let alone reverse them. Good people are being lost through emigration. For those who stay, extra work for the good of the business becomes hardly worth the effort, and some are already starting to opt out by paying increasing attention to their leisure activities.

Others, more concerned to maintain the family living standard, are seeking different ways to supplement their incomes. We shall find many more wives going to work, and this will have an undoubted effect on unemployment rates among the women and young people whose jobs they take. It will also undermine voluntary work.

It cannot make economic or social sense—in any form of economy known to the world—to put loyal and dedicated managers and higher paid workers, whose contribution is crucial to our international industrial competitiveness, under such relentless pressure.

Having the best possible managers and technicians in industry is of vital importance to the trade union movement, as it is to the whole country. Therefore incentives are necessary to get such people into industry in Britain, and then to encourage them to undertake and discharge the ever growing responsibilities which fall on them. Yet differentials are narrowing to a point that undermines commitment and damages the whole community.

The projected figures for August 1977 show that a managing director at present on £10,000 a year has, in terms of take-home pay, a worth only 2·97 times that of a person doing the least demanding work such as fetching, carrying, sweeping and cleaning. The managing director, on the other hand, has all the pressure on him of carrying responsibility for running a £1M business with 150 employees, being responsible for the development, manufacture, and sale of its products. Yet, if we compare this man with another managing director who has the responsibility of a very much larger business (with say £130M turnover and some 9,000 employees) we find that the latter's £30,000 per year salary is reduced by taxation to only 1·7 times more than that of his much less burdened colleague.

The only relevant measure of reward for those making their relative contributions towards the success and viability of their businesses is post-tax earnings. Yet most press or television reports of the relative pay of a labourer or a managing director dwell on the gross pay. This does not give any indication of the real differences in the standards of living, and permits political egalitarians to exploit differences which, in fact, do not exist.

KENNETH BOND
Deputy Managing Director, GEC

Reprinted from The Times of 2nd March 1976 by permission of the Editor.

Comparisons of Net Earnings at August 1975, August 1976 and August 1977

Married man with two children under 11

		Gross Pay Aug. '75	Gross Pay Aug. '76	Gross Pay Aug. '77	Net Pay Aug. '75	Net Pay Aug. '76	Net Pay Aug. '77	Net Pay Increase Aug. '75–'77	Differential of Net Pay to (1) Aug. '75	Aug. '76	Aug. '77
		(including Family Allowances)									
		£	£	£	£	£	£	%			
Labourer	1	2,100	2,400	2.520	1,899	2,108	2,205	16·1	1·00	1·00	1·00
Superintendent	2	4,000	4,300	4,472	3,134	3,343	3,474	10·8	1·65	1·59	1·58
Senior Manager	3	8,000	8,300	8,632	5,506	5,658	5,812	5·6	2·90	2·68	2·64
Managing Director	4	10,000	10,000	10,400	6,386	6,386	6,546	2·5	3·36	3·03	2·97
	5	15,000	15,000	15,500	8,130	8,130	8,280	1·8	4·28	3·86	3·76
	6	20,000	20,000	20,500	9,451	9,451	9,567	1·3	4·98	4·48	4·34
	7	30,000	30,000	30,500	11,266	11,266	11,351	0·8	5·93	5·34	5·15

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meeting on 17th February 1976 recommended to the Council the election and transfer of the following candidates. 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 communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

Meeting: 17th February 1976 (Membership Approval List No. 218)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

COLLINS, Bernard Patrick. *Henley-on-Thames, Oxfordshire.*
KING, Reginald Alfred. *Catterick, North Yorkshire.*
WILLIAMS, Clifford. *Cramlington, Northumberland.*

Transfer from Graduate to Member

BRIERLEY, Norman Leslie. *London.*
EVASON, John Richard. *Hazlemere, Buckinghamshire.*
GRAHAM, Wade Stuart. *Southampton.*
WILLIAMS, Kieran Francis. *Malahide, County Dublin.*

Direct Election to Member

COOKE-WELLING, David John. *Tunbridge Wells, Kent.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

HARRIS, Richard William. *Gamlingay, Cambridgeshire.*
JONES, James Roderic. *Portsmouth.*

Direct Election to Associate Member

HUGHES, Harrison Paul B. *Uxbridge, Middlesex.*
MORRIS, Geoffrey Thomas. *Darlington, North Yorkshire.*
OMENESA, Julius Onimisi. *Camberwell, London.*
SEDEGAH, Englebert M. Z. *Leeds.*
VLCEK, Pavel. *Gloucestershire.*

STUDENTS REGISTERED

ALEXANDER, John Reginald. *Slough, Buckinghamshire.*
BURKE, Martin James. *Dublin.*
CHAN, Chue. *London.*
CHAN, Ming Wah. *London.*
DUTHIE, Peter Joseph. *Kirby Muxloe, Leicestershire.*
FROST, Garry David. *Stowmarket, Suffolk.*
FLAVIN, Philip Graeme. *Felixstowe, Suffolk.*
GEAL, Christopher. *Bracknell, Berkshire.*
METCALFE, Colin Stuart. *Birmingham.*
POTTER, Howard Rhydian. *Neath, West Glamorgan.*
TEONG, Mee See. *London.*
XIROTAGAROS, Arhyrios. *Leicester.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

NAIR, K. Venugopalan. *Bombay.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

YU, Kai Wing. *Hong Kong.*
LAU, Kim-Wa Henry. *Hong Kong.*

Direct Election to Graduate

LAI, Huen Yu. *Hong Kong.*

Transfer from Associate to Associate Member

IRIMISOSE, Peter Oberuomo. *Kaduna, Nigeria.*

Transfer from Student to Associate Member

LOH, Wing Hoe. *Singapore.*
ROSE, Arnold Howard. *Geneva, Switzerland.*

Direct Election to Associate Member

OSEI-ANSAH, Samuel. *Takoradi, Ghana.*
TAN, Peng Yam. *Singapore.*
YOUNG, Victor S.A. *Freetown, Sierra Leone.*

Direct Election to Associate

AKINKUNMI, Akintunde. *Lagos, Nigeria.*

STUDENTS REGISTERED

AL-AHMAD, Hussain A.M. *Basrah, Iraq.*
CHAN, Sai Shing. *Hong Kong.*
CHUA, Colin. *Kuala Lumpur, Malaysia.*
CHEUNG, Chi Wah Daniel. *Hong Kong.*
CHEUNG, Fu Wah. *Hong Kong.*
FOK, Man Moon. *Hong Kong.*
PANG, Chi Shing. *Hong Kong.*
PUN, Ping Lan. *Hong Kong.*
TOH, Swee Hoe. *Kuala Lumpur, Malaysia.*
YAU, Huatt Chai. *Kuala Lumpur, Malaysia.*
YEUNG, Wing Sun. *Hong Kong.*

P.C.M. Equipment for Railway Communications

The first commercial 30-channel pulse-code modulation system to go into service in the UK has been handed over to British Rail by GEC Telecommunications Limited. The system connects New Street station in Birmingham with Birmingham International, the new main-line station built at the National Exhibition Centre. As the system is operating within the 25 kV overhead electrified network several problems created by this environment had to be overcome.

GEC 30-channel p.c.m. equipment enables thirty high-quality speech-bandwidth circuits carrying speech or data signal to be transmitted over two de-loaded pairs of a paper-covered quad-twin (p.c.q.t.) cable. Route length and noise and crosstalk on the line have no cumulative effect on signal quality: the performance depends on the ability of the equipment to detect the presence or absence of a pulse.

At the terminal 30 voice signals—300 to 3400 Hz—are multiplexed together into a single 2048 kbit/s digital output in the form of a high-density bipolar line code type HDB3. The multiplex equipment generates 32 time slots, 30 of which are used for voice channels, one for frame alignment, and one for signalling information. The 30 analogue channels are sampled successively at 8 kHz and a discrete eight-bit digital output code is produced for each sample. The signalling conditions associated with each of the 30 analogue channels are separately coded and time shared into slot 16.

The 2048 kbit/s digital output can be passed either to a p.c.q.t. balanced-pair cable, or to additional multiplex equipment for transmission over higher speed transmission media. Alarm indications provide for a number of fault conditions to be displayed at either terminal.

For transmission over audio-cable links, regenerative repeaters are spaced at 1·83 km (2000 yd) intervals (to coincide with the usual loading-coil separation). Each repeater employs automatic equalization and adjusts for cable losses over the range 4 to 37 dB at 1 MHz (equivalent to 200 m to 2 km of 0·63 mm p.c.q.t. cable). Power is fed to regenerative repeaters over the line phantom from a terminal.

Other 30-channel p.c.m. contracts for British Rail include: a system between Didcot and Reading in the Western Region which is part of a re-signalling scheme associated with the new 125 mile/hour high-speed trains; and a system for the East Coast re-signalling scheme in the Scottish Region. This latter system (to connect Edinburgh with Drem, Dunbar, Grantshouse, and Tweedsmouth) will include a standby line system which will be switched in automatically if the working line system fails.

The Duke of Edinburgh Announces Fellowship of Engineering at CEI Tenth Anniversary Conference

The Duke of Edinburgh, Founder President of the Council of Engineering Institutions, announced the establishment of a Fellowship of Engineering at a CEI banquet in Guildhall on 10th February to mark the tenth anniversary of the award of a Roy Charter to the Council.

Prince Philip, who was retiring from the presidency of CEI—now assumed by Lord Hinton of Bankside—said: 'The founder members will be those engineers who are presently Fellows of the Royal Society, together with a number of other distinguished engineers selected by the Chairman of CEI together with the Presidents of the Institutions. Further Fellows will be elected by the Fellowship up to a grand total of 1000 but not more than 60 in any one year.

'The Statutes say that the Senior Past President of the CEI shall be the Senior Fellow if he wishes to accept. I can tell you that I very much wish to accept and I shall do all I can to get this body of Fellows off to the best possible start.' Prince Philip said that he was sure that the Fellowship, whose purpose was to advise the President and anyone else including the Government, would do much to raise the standing and prestige of engineering in our modern society and to represent the interests of engineering in the corporate councils of the nation.

Earlier Prince Philip had said that the power for our economy came from the engines of industry and commerce, and just like mechanical engines they would not produce power unless all their components were properly designed and fitted together. 'Engineers', he said, 'are probably the most vital component of all. So it stands to reason that the organization of the engineering profession is critical to our whole national performance. It also stands to reason that the control of the profession must lie in the hands of engineers themselves. It is impossible for any highly complicated organization to be managed by people without experience or comprehension of its problems. . . .'

'I believe that the formation of the CEI ten years ago', continued Prince Philip, 'marked a turning point in the evolution of the engineering profession. It was a tacit recognition that the old carefree days of the fragmentation of the profession had come to an end. Quite a lot of engineers will have, I suspect, been a bit disappointed about progress since then but I think there is a natural reluctance even among the most imaginative and progressive engineers to accept without question sweeping changes in the future of their institutions. In this way they are merely reflecting what appears to be a national characteristic.'

Prince Philip said that fragmentation of the profession was both an expensive luxury and a fruitful field for demarcation disputes. 'I may not be an engineer myself', he went on, 'and I may not be able to speak for engineers but I can speak about engineers and it is my complete conviction that these latest developments in the CEI are in the best interests of the whole engineering profession from Chartered Engineers to Technicians. There are bound to be different levels of professional and technical competence but all levels are dependent on each other and there must always be freedom to climb the ladder of qualification from the bottom to the very top . . .

'Fragmentation has many attractions but this is an industrial and democratic country and we have got to live by industrial and democratic rules. Engineers simply cannot indulge in the luxury of cosy groups. Democracy means counting heads and votes, and engineers will only begin to exert their rightful influence if they have one instrument through which to voice their views.'

In handing over the Presidency to Lord Hinton, Prince Philip said that the time had come when it should be taken over by a distinguished and capable engineer. A new purpose was opening for CEI and he could think of no-one better to lead it through the challenging times ahead.

In reply Lord Hinton said that Prince Philip had been far more than a titular head of the organization always available and always anxious to give advice, and added, 'If in your ten years as President, you found that engineers rarely respond willingly to advice, you have only discovered what many less distinguished men have learned before you'. Lord Hinton said that inevitably today there were separate ladders for craftsmen, for technicians and for chartered engineers but hoped that everybody would remember that all were working on the same job, that different skills were complementary to one another and 'that our common interests can be achieved only by regular, sympathetic and understanding communications between the men on all the ladders. I hope that with a changed organization CEI will play an active part in achieving this unity between all the men and the women who have those different skills which go to make up engineering'.

Lord Hinton said that it was not easy to achieve the sort of constitution for CEI which many people would like it to have but believed that with patience, tolerance and wisdom on the part of all the bodies concerned the changes now planned could give to engineering the unity that it so badly needed. 'But change is needed not merely at the centre', he continued, 'but also at the grass roots, and the grass roots of the Institutions are in their branches! Many of the branches of the large Institutions are, today, bigger than the whole Institution of Mechanical Engineers when I joined as a Graduate in 1921 and yet the branches are run largely by part-time and voluntary workers.

'Surely it would be possible to make the Branch boundaries of all the member Institutions coincident; surely it would be possible to provide a full-time staff at common branch headquarters which would service all the CEI member institutions at that branch. Surely it would be possible to welcome members of all CEI institutions to all Branch Meetings irrespective of which Institution had sponsored the lecture and surely it would be possible to welcome to those meetings the members of technician institutes and societies.

'I believe that co-ordination and co-operation would be welcomed at Branch level and that it would lead to the disappearance of many of the present difficulties within CEI.'

In proposing the health of the guests at the banquet, Mr. Tony Dummett, who succeeded Professor John Coales as CEI Chairman earlier in the day, said that CEI was easily thought of as an organization remote from the realities of everyday life, but in fact it had no point or real existence without the working engineers whose interests ultimately it sought to represent. In a tribute to his predecessor, Professor Coales, Mr. Dummett said that it was a compound of strength and diplomacy that had made him a remarkable chairman. 'We started his year in some uncertainty; we finish it with our course firmly set, with several important reforms already under way and, above all, with CEI making a greater impact on national affairs than ever before.'

In spite of his essential contribution to modern society, continued Mr. Dummett, the engineers' public image is poor. 'Indeed for some time past, "engineer bashing" has become almost a popular sport. In the last year to two, a variant—"CEI bashing"—has become even more popular. We in CEI are determined to put this to rights.'

The need to get closer to individual engineers, said Mr. Dummett, was what had prompted the writing of a new constitution for CEI to include them as members. 'There is no point in pretending', he said, 'that there have not been discussions, even differences as to how best this could be done. It has, indeed, led to what I can describe as a new parlour game—CEI Constitution writing, a variation of scrabble . . . most constitutions can work if there is the will to make them do so. Let us remember Benjamin Franklin's words of two hundred years ago, "We must indeed hang together, or most assuredly we shall all hang separately".'

Members' Appointments

CORPORATE MEMBERS

H.R.H. The Duke of Kent, G.C.M.G., G.C.V.O. (President of the IERE) has been elected to Honorary Fellowship of the Institution of Electrical Engineers.

Mr. A. J. Davidson (Fellow 1964) who has been a Project Manager with the Overseas Development Administration, working on contracts in Zambia and Jamaica and latterly in the Yemem Arab Republic, has now joined Engineering and Power Development Consultants Ltd. as a consultant engineer.

Mr. T. B. McCrirrick (Fellow 1962, Member 1954, Graduate 1953) who has been Assistant Director of Engineering at the British Broadcasting Corporation since 1971, will take up the duties of Deputy Director on May 20th. Mr. McCrirrick

has been with the BBC since 1943 and has held a variety of posts on the operations side, and he will continue to be responsible to the Director of Engineering for professional engineering standards of operations and maintenance throughout the Corporation. He has served on the Institution's Papers Committee since 1973.

Col. I. W. Peck (Ret.) (Fellow 1966, Member 1959) who joined EASAMS Ltd., Camberley, in 1974, as Project Co-ordinator, Military Systems, is now a Project Co-ordinator with the Defence Systems Unit of Plessey Radar at Stoke Park, Slough.

Mr. David Simpson (Fellow 1962, Member 1958) has been appointed Managing Director of Gould Advance Ltd. From

1969 to 1974 he was with George Kent Ltd., initially as Group Director, and for the past year as Managing Director of Scientific and Medical Instruments Ltd. From 1962 to 1969 he was Managing Director of Hewlett-Packard Ltd. in Bedford and South Queensferry. Mr. Simpson was a member of the Institution's Council from 1971 to 1973.

Mr. R. F. Burns (Member 1973, Graduate 1964) has moved from the position of Group Leader with ITT Marine to a similar position in Rank Pullin Controls, where he is concerned with civil telecommunications development.

Sqn. Ldr. A. Dagger, Dip. Ed., RAF (Ret.) (Member 1972, Graduate 1964) is now a Senior Lecturer in Electrical Crafts at Wigan College of Technology. His last service appointment was that of Head of Training Resources, No. 1 Radio School, RAF, Locking.

Mr. F. A. Davenport (Member 1972, Graduate 1966) who was a Supervisory Engineer with Yorkshire Television Ltd., has been appointed Chief Engineer, RCA (Jersey) Ltd.

Sqn. Ldr. J. B. Dawson, RAF (Ret.) (Member 1970, Graduate 1966), formerly a Lecturer at the RAF College, Cranwell, is now Senior Lecturer in Radar Techniques at the REME School of Electronic Engineering, Arborfield.

Mr. G. P. McIntosh (Member 1973, Graduate 1972) has been appointed Chief Engineer and Manager of Lindsay CATV Ltd., Lindsay, Ontario. He was previously with International Aeradio Ltd., working at installations in the United Kingdom and abroad, latterly as Technical Supervisor at the Company's Toronto location.

Mr. M. Mulcay (Member 1967, Graduate 1963) has joined the Vidar Division of TRW as Manager, Radio Products, at Mountain View, California. His previous appointment was Technical Marketing Manager with Raytheon Data Systems in Norwood, Massachusetts.

Brigadier A. Needham, M.A. (Member 1968) has taken up the post of Director of Electrical and Mechanical Engineering at H.Q. BAOR. He was previously at the Headquarters of REME Support Group, Woolwich.

Mr. J. N. Price (Member 1973) has gone out to Ibadan, Nigeria, as Area Field Contract Controller for GEC Telecommunications Ltd. in connection with the expansion of the microwave radio network for the Federal Ministry of Communications. His previous overseas appointment was on secondment from GEC to the Malawi Post Office from 1971 to 1975.

Mr. D. A. Radage (Member 1972, Graduate 1967) is currently Senior Instrumentation and Control Engineer in the Central Engineering Department of Domtar Ltd., Montreal. He has been in Canada since 1968, working for various instrument companies.

Lt.-Col. G. D. Rush, REME (Member 1963) has completed three years with the Electrical Quality Assurance Directorate in Bromley, Kent, where he has been Principal Quality Officer for Army Radar, and is now at the Headquarters of the Directorate of Electrical and Mechanical Engineering (Army) in London as SO I (Communications).

Mr. J. R. Seaburne-May (Member 1973, Graduate 1968) has been appointed a Project Co-ordinator for naval gunnery contracts with EASAMS Ltd. He retired from the Royal Navy with the rank of Lieutenant last September.

Mr. M. E. Tickner (Member 1973, Graduate 1969) has been appointed Group Marketing Manager of Jiskoot Autocontrol Ltd., Tunbridge Wells. He has been with the Company since 1970, working on the production of process automation equipment.

Mr. A. C. Wadsworth (Member 1970, Graduate 1965) is now a Senior Development Engineer in the Avionics Development Laboratory of the MEL Equipment Company Ltd., Crawley, whom he joined in 1971.

Mr. J. A. Wyatt (Member 1965, Graduate 1963) who was formerly with the Data Systems Division of ITT, has been appointed Quality Assurance Manager with Chubb Alarms Manufacturing Ltd., Enfield, Middlesex.

NON-CORPORATE MEMBERS

Mr. John Savage (Companion 1972) who joined Plessey Avionics and Communications as Commercial Director in 1974, has been appointed to the Company's newly created post of Director, Strategic Marketing.

Mr. Michael Crawley (Graduate 1963) who joined Signetics in 1966 and has been Distribution Manager at the Linlithgow plant for the past two years, has been confirmed in this appointment by Mullard Ltd., who took over Signetics International commercial operations in the UK last November.

Mr. O. O. Gbeminiyi, M.Sc. (Graduate 1969) has been appointed Acting Head of the School of Engineering at Auchi Polytechnic, Nigeria. He joined the Polytechnic in 1974 as Acting Head of the Department of Electrical and Electronic Engineering, after three years as an Experimental Officer at the University of Ife. Mr. Gbeminiyi worked for various UK electronics organizations between 1962 and 1970, and then took the M.Sc. course in Systems Engineering at the University of Surrey.

Mr. W. W. Herbert (Graduate 1963) who has been with Home Office Communications since 1956, latterly as Chief Wireless Technician at the Central Communications Establishment, Harrow, has moved to the Wireless Depot at Marley Hill, Newcastle-upon-Tyne, where he is concerned with quality assurance.

Mr. Ooi Swee Tong, B.A.I., B.A. (Graduate 1975) has completed a degree course in engineering at Trinity College, Dublin, and, following technical training in television engineering in London, is now a trainee engineer with Robert Bosch (Malaysia) in Bayan Lepas, Penang, Malaysia.

Mrs. S. R. Roome, B.Sc. (Graduate 1975) who worked as an advisory computer programmer at the University of Sussex for a year after obtaining her degree, has taken up an appointment as Assistant Avionics Engineer with Hawker Siddeley Ltd. at Dunsfold Aerodrome.

Mr. J. N. Slater (Associate Member 1973, Associate 1969), formerly with the BBC Engineering Information Department, has recently joined the Independent Broadcasting Authority as an Information Officer in the Engineering Information Service.

Obituary

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

George William Bagshaw (Fellow 1930) died on 1st October 1975 at the age of 78 years.

Mr Bagshaw was active in the radio industry in the Sheffield area for nearly the whole of his professional life. Before and during the war he was Chief Engineer and manager of J. G. Graves Ltd and continued with this company when it was merged with the Plessey Group to become the Wireless Telephone Company Ltd. He remained on the board of the company and became Managing Director when it was transferred to Amphenol Borg Ltd., becoming Electronic Insulators Ltd.; he retired in 1970. Mr Bagshaw was instrumental in promoting occasional Institution meetings in the South Yorkshire area immediately after the war, and he subsequently took part in the formation of the present Yorkshire Section.

Francis William Charmer (Member 1966) died on 22nd August 1975, aged 41 years. He leaves a widow and a son and daughter.

Born and educated in Manchester, Frank Charmer entered the Post Office Engineering Department in Manchester as a youth in training in 1952, and during the next ten years obtained technical qualifications which eventually gained him membership of the Institution. For the first part of his career he was concerned with work on telephone exchange installation and maintenance. In 1963 he was appointed Executive Engineer, and on his transfer to Senior Executive Engineer in 1968 he moved over to postal service automation.

For many years Mr. Charmer had lectured in telecommunications engineering at technical colleges in Salford and Openshaw, and in 1970 he left the Post Office and was appointed a Lecturer II at Openshaw Technical College, where he taught ONC engineering, City and Guilds telecommunications to full technical certificate, and, more recently, advanced electronics. In October 1974 Mr. Charmer was seconded by the Manchester Education Authority to read for a M.Sc. degree in data communication at the University of Manchester Institute of Science and Technology. Unfortunately he was not able to complete his thesis but on the strength of his examination results he was awarded a first-class Honours Diploma in Technical Science.

John Michael Colbert (Member 1968, Graduate 1966) died on 24th November at the age of 49 years. He leaves a widow and four children.

Michael Colbert worked at Vickers Armstrong Ltd., Weybridge, as a maintenance engineer during the war and subsequently went into the RAF as a mechanic. On demobilization he went to All-Power Transformers Ltd., Byfleet, as an electronic technician, and then from 1951 to 1956 was with the War Department as a Technical Assistant. He returned to industry in 1956 at Cottage Laboratories Ltd., working on circuit development, and in 1961 he joined Peto Scott Ltd., Weybridge (later Pye TVT Ltd.), eventually becoming Group Leader of the Monitor Section.

In May 1972 Mr. Colbert went to Edwards Clifford Ltd., Brighton, to do development work concerned with X-rays, and in January 1973 he was appointed Senior Engineer with Redifon Flight Simulators Ltd., Crawley.

Reginald Edmund Dean (Associate 1947) died on 9th November 1975 at the age of 54. He leaves a widow.

After completing his secondary education at Fleetwood, Mr. Dean studied for the PMG's certificate at the Northern Counties Wireless School, Preston, and from 1939 to 1943 was a Radio Officer with Marconi International Communication Company. He was then commissioned in the RNVR and served as Radar Officer on North Atlantic convoys, and later was in charge of the radar school at Liverpool. On demobilization he went to the Ministry of Civil Aviation and was for a time at Prestwick Airport. He then went into industry and was for a number of years with Pye TMC at Airdrie. Mr. Dean later held the posts of Senior Test Engineer and Calibration Engineer with Pye Scottish Telecommunications Ltd., for whom he worked up to the time of his death.

Captain Nicholas Samuel John Hayward, REME (Member 1975, Graduate 1962) died on 14th June, aged 38. He leaves a widow and three children.

After leaving Welbeck College, Sam Hayward went to Sandhurst and was commissioned into the REME in 1948. He subsequently served at Field Workshops in the Far East and in Europe, returning to the UK in 1968–69 to take the Officers' Long Electronic Course at the School of Electronic Engineering at Arborfield. In March 1973 Capt. Hayward was posted to the Radar Branch of the Technical Group, REME, at the Royal Radar Establishment, Malvern, as second in command, and two years later joined the Guided Weapons course at the Royal Military College of Science, Shrivenham. It was during a visit with that course to the Glasgow area that he collapsed and died.

William Swan, B.Sc., M.Sc. (Member 1948) died in December 1975 at the age of 57. He leaves a widow and married daughter.

Born and educated near Edinburgh, William Swan entered the RAF in 1934 as an Aircraft Apprentice, studying at the Electrical and Wireless School and served as a Wireless Operator Mechanic on ground stations and aircraft equipment until 1943, when he reached the rank of Warrant Officer. He was then commissioned and continued to serve in the RAF as a Technical Officer in the Signals Branch until he retired in 1961.

Mr. Swan then went to Canada, where he studied for his B.Sc. degree in mathematics and physics at the University of Manitoba. In 1965 he was appointed an instructor in the Department of Electronics at Southern Alberta Institute of Technology, where he remained until 1971, when he joined the Department of Computing Science at the University of Calgary, shortly after obtaining an M.Sc. degree of the University.

Ronald Frederick Worrad, M.Sc. (Member 1964, Graduate 1963) died on 9th July 1975, at the age of 50 years. He leaves a widow and two sons.

After war service in the Royal Navy as a Radio Mechanic, Mr. Worrad worked in the radio industry for some three years and then, in 1949, entered technical education at North Staffordshire Technical College. In 1957 he moved to Cheshire and for some five years was at South Cheshire College of Further Education. In 1965 he obtained the M.Sc. degree of the University of Aston in Birmingham and was appointed as Lecturer at Rotherham College of Technology. Two years later he moved to Bristol and at the time of his death he was a Lecturer in the Engineering Department of Bristol Polytechnic.

Forthcoming Institution Meetings

London Meetings

Wednesday, 21st April

ELECTRONICS PRODUCTION TECHNOLOGY GROUP

Colloquium on AUTOMATIC PRODUCTION

The Royal Institution, 21 Albemarle Street, London, W1, 2 p.m.

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

Wednesday, 28th April

AUTOMATION AND CONTROL SYSTEMS GROUP AND COMPONENTS AND CIRCUITS GROUP

Colloquium on STEPPING MOTOR DRIVES

IERE Lecture Room, 10 a.m.

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

Thursday, 29th April

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

A Novel Approach to Marine Surveying

By J. M. Thompson (*Decca Survey*)
IERE Lecture Room, 6 p.m.

Wednesday, 5th May

COMPONENTS AND CIRCUITS GROUP

Colloquium on FUSES AND CIRCUIT PROTECTION

IERE Lecture Room, 2 p.m.

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

Wednesday, 12th May

EDUCATION AND TRAINING GROUP

Colloquium on TRAINING OF ELECTRONICS ENGINEERS AND TECHNICIANS IN THE ARMED SERVICES

School of Electronic Engineering, Arborfield, Reading, 10.25 a.m.
Advance registration necessary. For further details and registration forms, apply to Meetings Secretary, IERE.

Thursday, 13th May

AUTOMATION AND CONTROL SYSTEMS GROUP

The Automated Warehouse
Speaker to be confirmed.

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

Tuesday, 18th May

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Advances in Earth Station Technology and Satellite Communications

By R. I. Harris and D. Longhurst (*SRDE*)
IERE Lecture Room, 6. p.m.
(Tea 5.30 p.m.).

Wednesday, 19th May

MEASUREMENTS AND INSTRUMENTS GROUP

Colloquium on DYNAMIC ANALYSIS INSTRUMENTATION

IERE Lecture Room, 2. p.m.

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

Wednesday, 26th May

COMPONENTS AND CIRCUITS GROUP

Colloquium on INDUSTRIAL CATHODE RAY TUBES

IERE Lecture Room, 2 p.m.

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

Thames Valley Section

Thursday, 29th April

ANNUAL GENERAL MEETING

at 7 p.m. followed by a lecture on

Mobile Radio in the Era of Spectrum Congestion

By Professor W. Gosling (*University of Bath*)

Caversham Bridge Hotel, Caversham Road, Reading, 7.30 p.m.

South Western Section

Wednesday, 28th April

JOINT MEETING WITH IEE

Developments in Marine Navigation Aids

By S. F. Appleyard (*Marconi International Marine*)

Lecture Room, School of Chemistry, University of Bristol, 6 p.m. (Tea 5.30 p.m.)

Monday, 3rd May

ANNUAL GENERAL MEETING

Royal Hotel, College Green, Bristol, 7 p.m.
(Coffee available afterwards)

Beds & Herts Section

Tuesday, 27th April

ANNUAL GENERAL MEETING

followed by a lecture on
Electronics in the Tuning of Musical Instruments

By L. H. Bedford, C.B.E., Past President.
Hatfield Polytechnic, College Lane, Hatfield, 7.45 p.m.

East Midlands Section

Tuesday, 11th May

JOINT MEETING WITH IEE

Special Effects in Colour Television

By A. B. Palmer (*BBC TV Studios*)

EMEB Offices Sports and Social Club, Wigman Road, Bilborough, Nottingham, 7.30 p.m. (Refreshments available after the lecture.)

South Midlands Section

Tuesday, 27th April

Guided Weapons

By Captain R. G. W. Maltby (*REME*)

To be followed by the

ANNUAL GENERAL MEETING

Golden Valley Hotel, Cheltenham, 7 p.m.

Friday, 7th May

Golden Jubilee Barn Dance
Bredon Village Hall,

8 p.m. to 12 a.m.

Tickets £1.50, including buffet, from any Committee member or from the Honorary Secretary, D. W. Iliffe, C.Eng., MIERE. 'Portland', 16 Stantons Drive, Swindon Village, Cheltenham, Glos.
Telephone: Cheltenham 26323.

Yorkshire Section

Friday, 9th April

ANNUAL GENERAL MEETING

Leeds Polytechnic, 8 p.m.

North Western Section

Thursday, 22nd April

ANNUAL GENERAL MEETING

followed by a lecture on

The Philips Video Disc

By K. J. Bogel (*Philips Electrical Ltd.*)

University of Manchester Institute of Science and Technology, 6.15 p.m. Light refreshments will be available beforehand.