The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

The European Prospect for Electronics

SOME of the realizations of the past and hopes for the future of the British electronics industry are set out in the annual reports issued in the past month by three of the leading trade associations of the industry, namely, the Electronic Engineering Association (EEA), the British Radio Equipment Manufacturers' Association (BREMA) and the Radio and Electronic Component Manufacturers' Federation (RECMF).

As might be expected, all three Associations refer to the entry of the United Kingdom into the European Economic Community, welcoming this important event with guarded optimism. The EEA is perhaps the least optimistic, pointing out that the purchasing policies for capital equipment in European markets have been characterized by strongly nationalistic preferences: for communications equipment the UK has supplied only between 1 and 2% of the total EEC market and for computing systems approximately 5%. The prospects 'for expansion of the latter are however considered good since Europe is regarded as the world's fastest growing market for computers and peripherals. On the radio and television receiver and audio equipment front, BREMA comments that the existing overseas marketing networks built up by British manufacturers for other industrial products will provide a basis for increasing sales. On the 1972 figures, exports have been only about £13·5 million of which £4 million was to EEC countries and about 75% in value was audio equipment, so that there would seem to be ample scope here.

On a related problem, the reports discuss the threat to all European electronics industry of Japanese competition—principally as yet in consumer electronics but seen by the EEA as spreading into the capital equipment market. Reference is made by BREMA and RECMF to the industry delegation which visited Tokyo to discuss some form of voluntary control on radio and television exports to the UK: both were non-committal about the likely outcome. RECMF sees some compensation for its members in selling British-made components to Japanese factories set up in Britain.

Implicit in greater international co-operation is the acceptance of agreements on an increasing number of technical matters. BREMA and EEA refer to EEC Directives on safety of low voltage equipment and on interference requirements, and clearly the existing IEC standards will form the basis of much work in these fields. The importance of harmonization of standards is recognized in the developments in the BS 9000 system for electronic components of assessed quality and in the complementary activity with other European countries in relating the British standard to CENEL 1. The setting-up of the European Electronic Component Manufacturers' Association, under which the relevant technical bodies CEMAC and CEPAC, previously concerned respectively with active and passive components and now merged as CEMEC, is reported by RECMF as a welcome move.

The effectiveness with which British industry as a whole is able to make the most of opportunities offered by our membership of EEC will depend on guidance it receives on Government policy. This point was made by the retiring President of EEA, Dr. B. J. O'Kane, when speaking at his Association's annual luncheon and he urged the Secretary of State for Trade and Industry to meet the industry from time to time to discuss such policy matters on a formal or semi-formal basis. The response by the Minister of Aerospace (Mr. M. Heseltine) referred to the DTI, the banking community and manufacturing industry as operating a three-part team for overseas trade, and such encouragement must surely brighten the prospect for the electronics industry.

Contributors to this issue

Dr. C. J. Paull graduated with a B.Sc.(Eng.) degree from the Department of Electrical Engineering, University College London, in 1966. He then served on the staff of the London Electricity Board for a short time before embarking on research in signal synthesis and processing techniques at the University College of Wales, Swansea, where he subsequently obtained both M.Sc., and Ph.D. degrees. Dr. Paull held the Wayne Kerr Industrial Fellowship at Swansea prior to taking up his present appointment as a Lecturer in the Department of Electrical and Electronic Engineering at Nottingham University.

Mr. W. A. Evans graduated with a B.Sc. in pure science with honours in Physics at the University College of Swansea, in 1957 and later obtained an M.Sc. degree in information engineering at Birmingham University in 1958. His industrial experience has included computer design with Ferranti Packard Electric Ltd., Toronto, and participating in the Canadian space program whilst a member of the Scientific Staff of RCA Victor Company Montreal. In 1966 he re-entered academic life as a Lecturer at the University College of North Wales, returning to Swansea as a Lecturer in the Department of Electronic and Electrical Engineering in 1968. His interests lie in electronic instrumentation and at Swansea he directs a research group particularly concerned with system identification and waveform synthesis.

Mr. R. J. Clarke trained as a student apprentice at the English Electric Valve Co., Chelmsford, with academic periods at the then Loughborough College of Advanced Technology between the years 1957 and 1961. After spending a further year working on television camera tubes he moved to Standard Telephones and Cables, North Woolwich, joining the submerged repeater test equipment division. In 1964 Mr. Clarke returned to Loughborough to take an M.Sc. degree by research into wideband field-effect transistor amplifiers and in 1966 he was appointed to his present position as a Lecturer in the Department of Electronic and Electrical Engineering.

Mr. A. M. Hayes joined the University of Bradford in June 1970 as a research student to investigate the design of correlation equipment for flow measurement. He graduated at Cambridge in 1967 in a combination of Mathematics and Electrical Sciences. Prior to his present appointment he was with Rotax, now Lucas Aerospace, of Bradford, where he was concerned with the design of digital position measurement equipment for machine tools.

Mr. G. Musgrave started his engineering career with A. Reyrolle & Co. as a student apprentice and gained his first degree from Bradford University. Later he was awarded the M.Sc. degree of the University of Manchester Institute of Science and Technology as a result of research into aspects of pulse width modulation. He then returned to Bradford University as a Lecturer and he took up his present appointment as Reader in Electrical and Electronic Engineering at Brunel University in January 1973.

Dr. D. R. Wilson (Fellow 1972) is a graduate of the Faculty of Technology, University of Manchester, now the Institute of Science and Technology. After a period on the staff of the Electrical Engineering Department of the Faculty, he entered industry holding appointments with GEC–AEI (Electronics) and with Plessey Radar, where he worked mainly on the design of servo controls for aerials. In 1972 Dr. Wilson was appointed Head of the Department of Engineering and Systems Technology at the Polytechnic of Central London. A fuller note on his career was published in the Journal in September 1969.

Mr. M. Burl went to Centrum, a Swedish communications company, as an electronics technician on leaving school in 1963. He subsequently joined the British Post Office at Dollis Hill in 1967 where he worked on digital signal processing techniques. In 1968 Mr. Burl completed his Higher National Certificate and started a sandwich degree course in electronic engineering at the Polytechnic of Central London; all his training in the sandwich course was obtained at the Post Office. Since graduating in June 1972, Mr. Burl has been pursuing a research programme into servo filter methods at the PCL.

A method of generating a high-quality circle test pattern for television

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SUMMARY

In this paper a method of generating a circle test pattern for television is described. An important feature of the method is the use of switching matrices as the source of line information. Corresponding generator organization is given and the generated test pattern is shown.

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1 Introduction

For the study of the geometrical quality and linearity of the television picture, a circle pattern is one of the most important test patterns for a television test generator. The imperfections can be recognized as deformations of the test circle.

High accuracy and stability are required of the test circle generation method to achieve reliable results. These requirements cannot be easily satisfied by analogue generation methods. It is then rather difficult to determine which deformations arise from deflexion non-linearities and which from the distortion of the initial pattern.

The required accuracy and stability of the operation can be achieved if the pattern is generated by numerical means. A previous paper[1] described a computational method of circle generation based on recursive incremental algorithms which were used for line information generation.

The present paper discusses another digital approach to the circle generation problem. The line information required is generated by switching matrices, thus reducing the computation time and device complexity. The generator is realized with medium-scale integrated t.t.l.-circuits which make the construction also economically feasible.

2 General Considerations on Circle Generation

The method of image scanning of a circle on a television screen is presented in Fig. 1.

All the principal television systems utilize 2:1 interlaced scanning. Solid lines in Fig. 1 belong to the odd-field and dotted lines to the even-field of the image. Circle points B_v and D_v on line v ($v = 1, 2, \ldots, R+1$) and $B_{\bar{v}}$ and $D_{\bar{v}}$ on line $\bar{v}$ ($ = 1, 2, \ldots, R$) are the in and out points of the white inserts respectively. The locations of the in and out points are measured from the beginning of each line as multiples of a unit length. The unit is chosen to be the distance between adjacent lines, resulting in equal resolution in horizontal and vertical
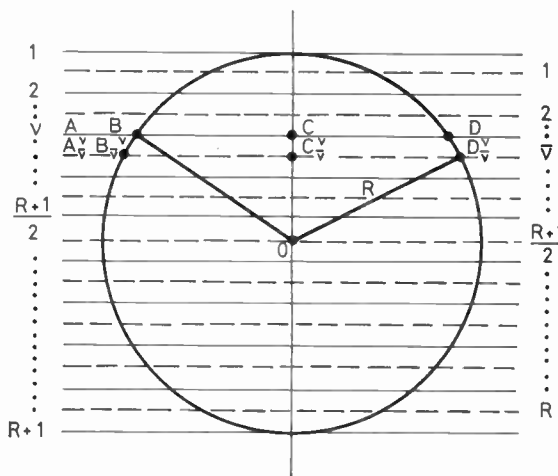

Fig. 1. Image scanning on television screen.

directions in circle writing. The radius of the circle is then R units.

The positions of B_v and D_v measured from the beginning of the line v, can be determined according to eqns. (1) and (2):

$$A_v B_v = A_v C_v - B_v C_v \qquad (1)$$

$$A_v D_v = A_v C_v + B_v C_v \qquad (2)$$

These equations can also be applied in the even-field changing v into $\bar{v}$.

The lengths of line segments $B_v C_v$ and $B_{\bar{v}} C_{\bar{v}}$ can be generated numerically. The structure of the circle is analysed by a general purpose computer in terms of scanned image parameters. Using this information one switching matrix for each field can be designed to generate the length of $B_v C_v$ or $B_{\bar{v}} C_{\bar{v}}$ from the ordinal number of corresponding line. By this means rounded integer-valued approximations of $B_v C_v$ or $B_{\bar{v}} C_{\bar{v}}$ are obtained.

3 Function Generation by a Switching Matrix [2]

In Fig. 2 an integer-valued approximation of function $f(x)$ is presented. The domain of the function is $x_0 \leq x < x_N$. The approximation function $F(x)$ is defined by eqn. (3).

$$F(x) = \text{int} \{f(x_i) + 0 \cdot 5\}, \qquad (3)$$

for $x_i \leq x < x_{i+1}$ and $i = 0, 1, \ldots, N$.

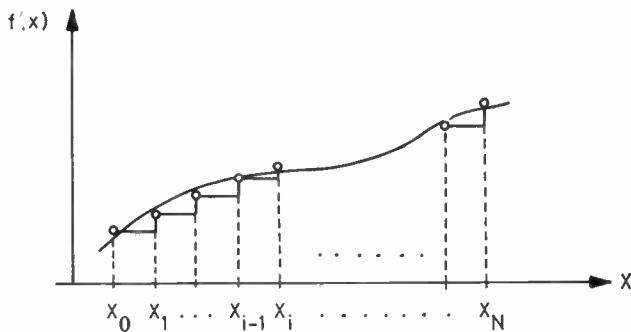The operator int $\{.\}$ takes the integer part of the value of the function in parenthesis.
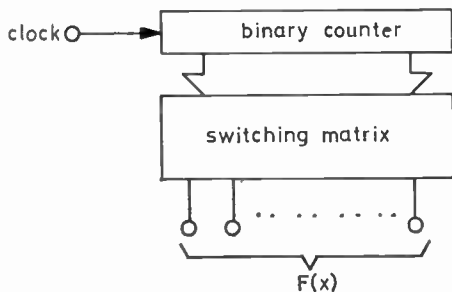

Fig. 2. Approximation of function $f(x)$.

Fig. 3. Function generation by a switching matrix.

The discrete values of the function $F(x)$ can be generated by a switching matrix of Fig. 3. Clock pulses to the binary counter denote the discrete steps of the argument. The input to the switching matrix is the state of the binary counter and the output is the rounded integer-valued approximation of $f(x)$ in binary form.

When the number of discrete values of the function increases the use of a switching matrix for function generation becomes impractical because of its size. The practical range of use can, however, be considerably expanded if the increment $\delta(x_i)$ of the function is generated rather than the value of the function itself. The definition of $\delta(x_i)$ is

$$\delta(x_i) = F(x_i) - F(x_{i-1}). \qquad (4)$$

Then the value of the function $F(x_i)$ can be calculated as the sum of the increments.

$$F(x_i) = \delta(x_0) + \sum_{j=1}^{i} \delta(x_j), \qquad (5)$$

where

$$\delta(x_0) = F(x_0).$$

4 Switching Matrix for the Generation of Line Information in Odd-fields [3]

Using the techniques described in the previous section the circle function can also be generated. The number of lines in the odd field belonging to the circle pattern in Fig. 1 is $R+1$. Then the number of increments of different absolute values needed for the generation is $(R-1)/2$. R is chosen to be 215 units. Then the number of increments is 107 and the capacity of the binary counter in Fig. 3 must be seven bits.

The numerical values of the increments can be calculated from eqn. (6):

$$\delta(v) = \rho(v) - \rho(v-1), \qquad (6)$$

where

$$v = 0, 1, 2, \ldots, R,$$
$$\delta(0) = 0$$

and

$$\rho(v) = \text{int} \{\sqrt{R^2 - (R - 2v)^2} + 0 \cdot 5\}.$$

The maximum absolute value of the increments is 29, which can be represented with five bits In Table 1 the values of the increments $\delta(v)$ are tabulated with corresponding ordinal number of the line v in decimal form. Letter 'd' in the columns $\delta(v)$ and $\delta(\bar{v})$ denotes 'don't care' terms, which can be used to simplify the logic expressions. Converting the values in Table 1 into binary form and notating the variables coding v and $\delta(v)$ by A_7 through A_1 and B_{5t} through B_{1t} respectively in the odd-field Table 2 is obtained, where the beginning of the converted Table 1 is given to show various notations.

Simplification of logic functions $B_{it}(i = 1, 2, 3, 4, 5)$ of seven variables is manually very cumbersome. Therefore the functions have been simplified with a digital computer using Quine-McCluskey method. The computer program performs also the selection of prime implicants to

Table 1

Decimal values of the increments for odd- and even-fields ($\delta(\bar{v})$ odd, $\delta(v)$ even)

$v, \bar{v}$	$\delta(v)$	$\delta(\bar{v})$	$v, \bar{v}$	$\delta(v)$	$\delta(\bar{v})$	$v, \bar{v}$	$\delta(v)$	$\delta(\bar{v})$	$v, \bar{v}$	$\delta(v)$	$\delta(\bar{v})$
0	29	21	33	2	2	66	1	1	99	0	0
1	12	15	34	2	2	67	1	1	100	1	1
2	9	10	35	2	2	68	1	0	101	0	0
3	8	8	36	1	1	69	0	1	102	0	0
4	7	8	37	2	2	70	1	1	103	0	0
5	6	6	38	2	2	71	1	1	104	0	0
6	5	6	39	1	2	72	1	0	105	0	0
7	5	5	40	2	1	73	0	1	106	0	d
8	5	5	41	1	2	74	1	1	107	d	d
9	5	4	42	2	1	75	1	0	108	d	d
10	4	5	43	1	2	76	0	1	109	d	d
11	4	4	44	2	1	77	1	0	110	d	d
12	3	4	45	1	2	78	0	1	111	d	d
13	4	3	46	2	1	79	1	1	112	d	d
14	4	4	47	1	1	80	0	0	113	d	d
15	3	3	48	1	2	81	1	1	114	d	d
16	3	3	49	2	1	82	0	0	115	d	d
17	3	4	50	1	1	83	1	1	116	d	d
18	3	3	51	1	2	84	0	0	117	d	d
19	3	2	52	1	1	85	1	0	118	d	d
20	3	3	53	1	1	86	0	1	119	d	d
21	2	3	54	2	1	87	0	0	120	d	d
22	3	3	55	1	1	88	1	1	121	d	d
23	2	2	56	1	1	89	0	0	122	d	d
24	3	3	57	1	1	90	0	0	123	d	d
25	2	2	58	1	1	91	1	1	124	d	d
26	2	2	59	1	1	92	0	0	125	d	d
27	3	3	60	1	1	93	0	0	126	d	d
28	2	2	61	1	1	94	1	0	127	d	d
29	2	2	62	1	1	95	0	1			
30	2	2	63	1	1	96	0	0			
31	2	2	64	0	1	97	0	0			
32	2	2	65	1	1	98	0	0			

Table 2

Beginning of converted Table 1

Ordinary number of a line v									Increment					
Decimal	Binary								Decimal	Binary				
	A_7	A_6	A_5	A_4	A_3	A_2	A_1			B_{5t}	B_{4t}	B_{3t}	B_{2t}	B_{1t}
0	0	0	0	0	0	0	0		29	1	1	1	0	1
1	0	0	0	0	0	0	1		12	0	1	1	0	0
2	0	0	0	0	0	1	0		9	0	1	0	0	1
3	0	0	0	0	0	1	1		8	0	1	0	0	0
4	0	0	0	0	1	0	0		7	0	0	1	1	1

cover the values of the function.[4-6] The result is obtained in sum-of-products form. To reduce the number of gate inputs the functions are presented using only NAND logic operations. The results are given in eqns. (7) through (11). The realization of these functions is straight-forward.

$$B_{5t} = \overline{\overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_4 \bar{A}_3 \bar{A}_2 \bar{A}_1}} \tag{7}$$

$$L_{4t} = \overline{\overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_4 \bar{A}_3}} \tag{8}$$

$$B_{3t} = \overline{\overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_4 \bar{A}_3 A_2 A_4 \bar{A}_3 \bar{A}_2 A_1 A_3 A_2 \bar{A}_1}} \tag{9}$$

$$B_{2t} = \overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_6 \bar{A}_3 \bar{A}_4 A_1 A_2 A_1 \cdot A_3 \bar{A}_2 A_4 \bar{A}_4 A_4 A_1 \cdot \bar{A}_6 A_3 \bar{A}_4 \bar{A}_2 A_1 \bar{A}_4 \bar{A}_2 A_5} \cdot$$
$$A_6 \bar{A}_5 \bar{A}_4 \bar{A}_3 \bar{A}_2 A_1 A_3 \bar{A}_2 \bar{A}_1 \tag{10}$$

$$B_{1t} = \overline{A_7 \bar{A}_6 A_1 \bar{A}_5 \bar{A}_2 A_4 A_6 A_1 \bar{A}_3 \bar{A}_2 A_4 A_6 \bar{A}_5 \bar{A}_2 A_1 A_4 \bar{A}_3 \bar{A}_2 A_3 \bar{A}_2 A_5 \bar{A}_4 A_3 \bar{A}_2 A_1 A_7 \bar{A}_6 \bar{A}_3 A_2}$$
$$A_4 \bar{A}_3 \bar{A}_2 \bar{A}_6 A_1 \bar{A}_7 \bar{A}_5 A_1 \bar{A}_6 A_1 \bar{A}_5 \bar{A}_4 A_3 \bar{A}_7 \bar{A}_2 A_2 \bar{A}_5 A_3 \bar{A}_5 \bar{A}_3 A_3 \bar{A}_7 \bar{A}_4 \bar{A}_3 A_1 \bar{A}_5 A_2 A_4 \bar{A}_3 \bar{A}_5 A_1 \bar{A}_5 A_2 \bar{A}_1 \tag{11}$$

(The arrow head in these formulae denotes that the negation operation bar continues on the next line.)

5 Switching Matrix for the Generation of Line Information in Even-fields

The number of lines in the even field belonging to the circle pattern in Fig. 1 is R. The numerical values of the increments in the even field can be calculated from eqn. (12):

$$\delta(\bar{v}) = \rho(\bar{v}) - \rho(\bar{v}-1), \tag{12}$$

where

$$\bar{v} = 0, 1, 2, \ldots, \quad R-1,$$
$$\delta(v) = 0$$

and

$$\rho(\bar{v}) = \text{int} \left\{ \sqrt{R^2 - (R - 2\bar{v} - 1)^2} + 0.5 \right\}.$$

The maximum absolute value of the increments is 21, which can be represented with five bits. In Table 1 the values of the increments $\delta(\bar{v})$ with corresponding ordinal number $\bar{v}$ of the line are tabulated in decimal form.
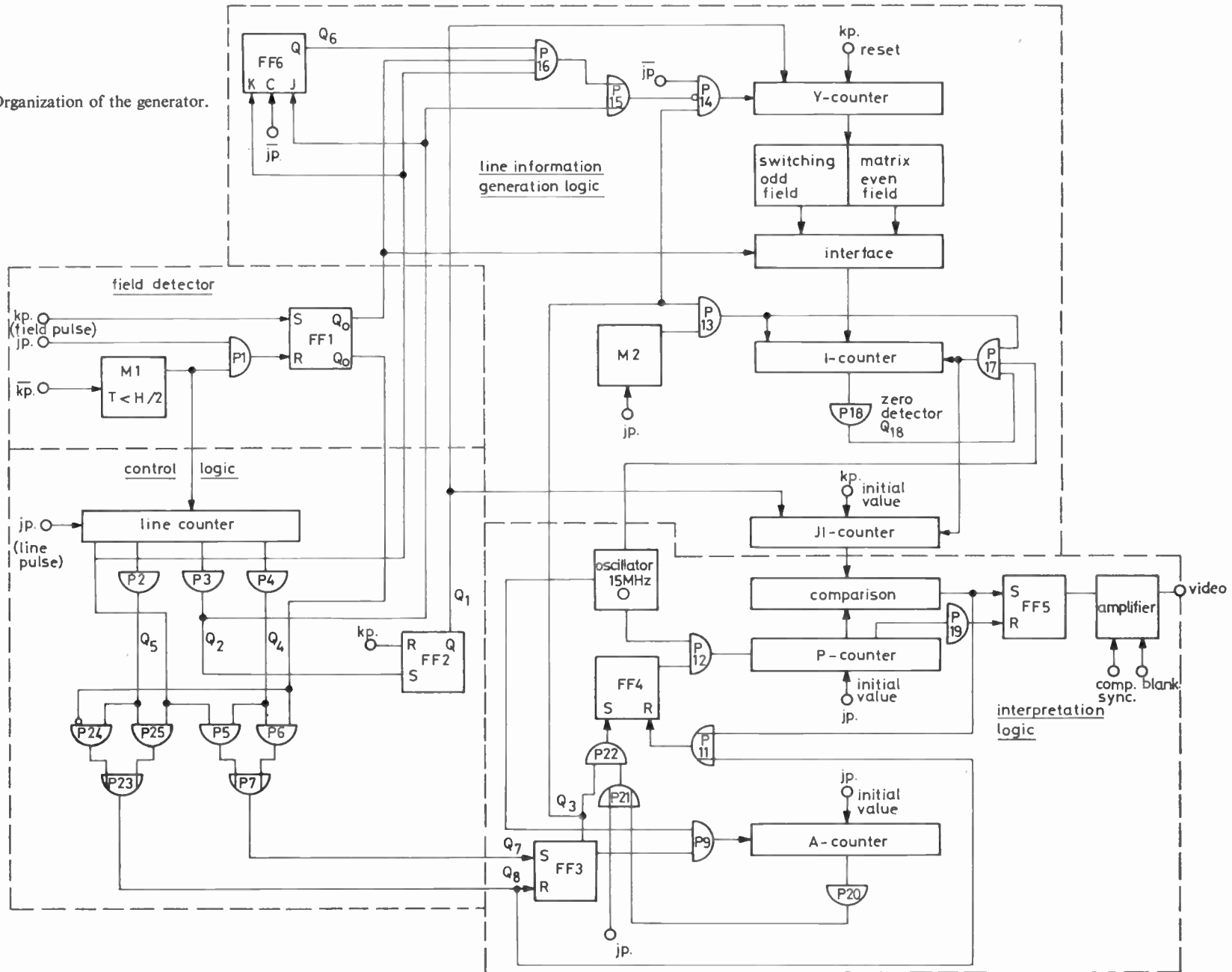
As in the previous Section the conversion into binary form is performed with the notations A_7 through A_1 and B_{51} through B_{11} for the variables encoding $\bar{v}$ and $\delta(\bar{v})$ respectively. After the simplification with a digital computer the final forms of the logic functions B_{51} through B_{11} eqns. (13) through (17) are obtained.

$$B_{51} = \overline{\overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_4 \bar{A}_3 \bar{A}_2 A_1}} \tag{13}$$

$$B_{41} = \overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_4 A_3 A_1 A_2 A_3 \bar{A}_2 A_1} \tag{14}$$

$$B_{31} = \overline{A_7 \bar{A}_6 \bar{A}_5 \bar{A}_3 A_2 A_4 \cdot A_4 A_1 \cdot A_4 A_3 A_1 A_2 A_4 A_3 A_2 A_1} \tag{15}$$
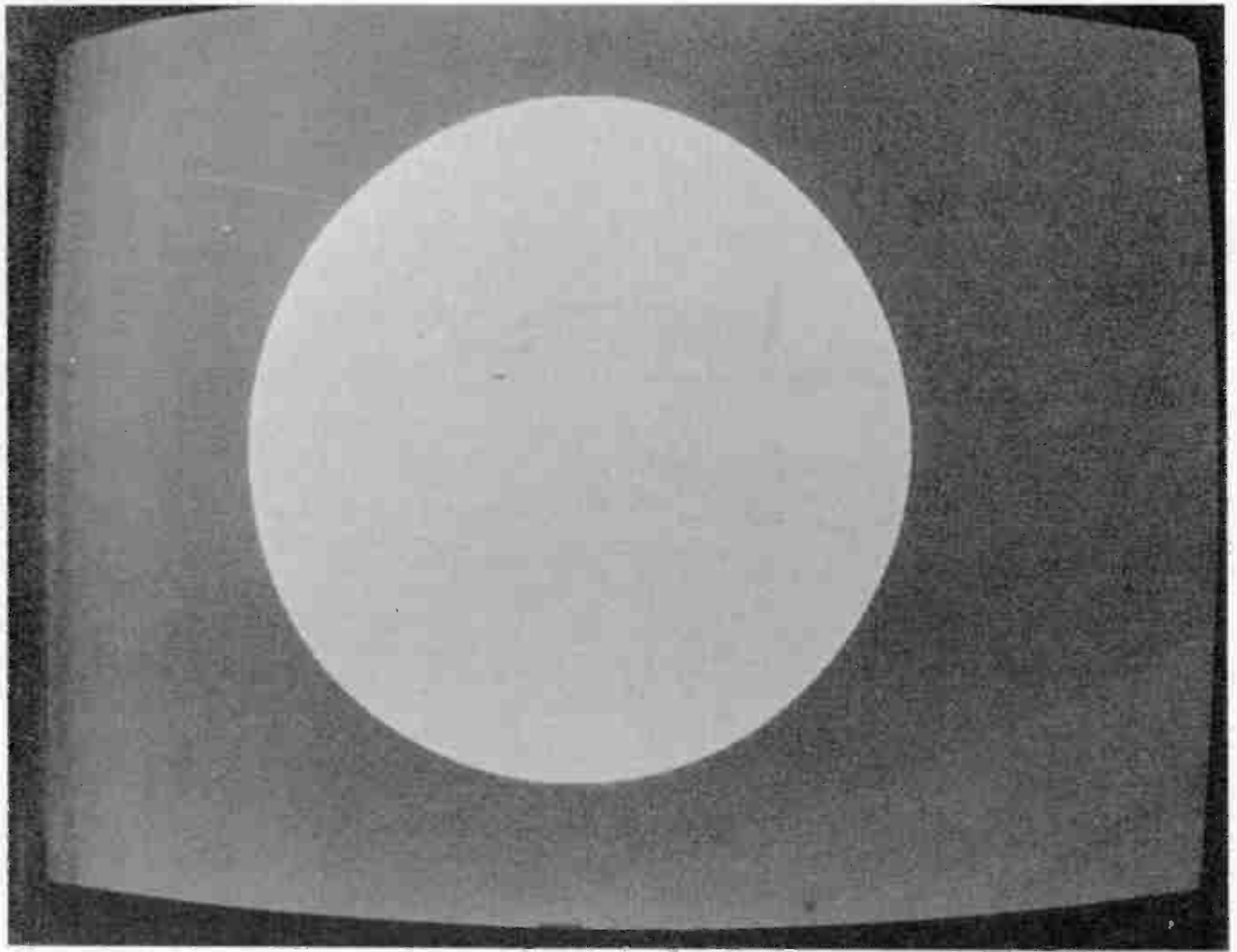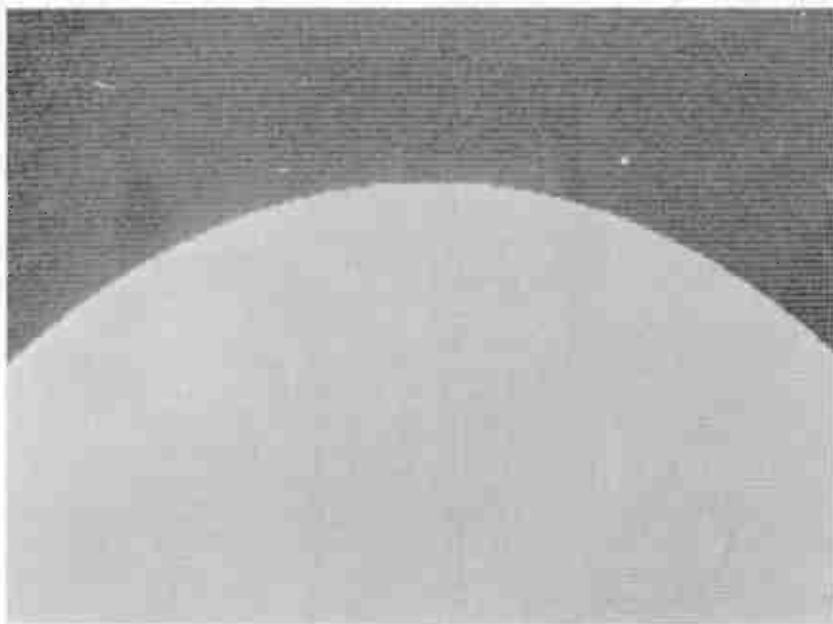
Fig. 4. Organization of the generator.

$$B_{21} = \bar{A}_7\bar{A}_6\bar{A}_5\bar{A}_4\bar{A}_1\bar{A}_2\bar{A}_4\bar{A}_2\bar{A}_1\bar{A}_3\bar{A}_1\bar{A}_4\bar{A}_2 \cdot \bar{A}_5\bar{A}_2\bar{A}_1\bar{A}_4\bar{A}_3 \cdot \bar{A}_6\bar{A}_3\bar{A}_1\bar{A}_4\bar{A}_4\bar{A}_2\bar{A}_5\bar{A}_4\bar{A}_3\bar{A}_2\bar{A}_1\bar{A}_2\bar{A}_1 \quad (16)$$

$$B_{11} = \bar{A}_7\bar{A}_5\bar{A}_4\bar{A}_3\bar{A}_2\bar{A}_6\bar{A}_1\bar{A}_2\bar{A}_1 \cdot \bar{A}_6\bar{A}_2\bar{A}_1\bar{A}_3\bar{A}_4\bar{A}_3\bar{A}_2\bar{A}_1\bar{A}_5\bar{A}_4 \cdot \bar{A}_3\bar{A}_1\bar{A}_6\bar{A}_2\bar{A}_3\bar{A}_1\bar{A}_6\bar{A}_4\bar{A}_1$$

$$\bar{A}_6\bar{A}_5\bar{A}_4\bar{A}_3\bar{A}_2\bar{A}_3\bar{A}_2\bar{A}_1\bar{A}_7\bar{A}_1\bar{A}_4\bar{A}_4\bar{A}_3\bar{A}_2\bar{A}_4\bar{A}_3\bar{A}_1\bar{A}_2\bar{A}_1 \cdot \bar{A}_3\bar{A}_7\bar{A}_4\bar{A}_4\bar{A}_1\bar{A}_5\bar{A}_4\bar{A}_2\bar{A}_1 \cdot \bar{A}_5\bar{A}_2\bar{A}_4\bar{A}_3\bar{A}_1\bar{A}_7\bar{A}_4\bar{A}_1$$

$$\bar{A}_6\bar{A}_5\bar{A}_4\bar{A}_3\bar{A}_2\bar{A}_1 \quad (17)$$

6 Organization and Operation of the Generator

6.1 Operation of the Generator

The organization of the test generator based on switching matrices is presented in Fig. 4. It has been designed to meet the requirements of the system CCIR-B.[7] In this system there are 625 lines, 40 of which belong to the field blanking period. Thus the number of lines which can be seen is 585. The diameter of the circle has been chosen to be 75% of the height of the picture. Then the radius of the circle R is 215 units. For the proper operation of the generator it is necessary to detect which field will be written next on the screen. The detection is performed with the monostable multivibrator M1, gate P1 and RS-flip-flop FF1. The variable Q_0 is true in the odd field and false in the even field.

With the decoding gate P4 and gates P5, P6 and P7, FF3 is set depending on the field in the coincidence of the front edge of 88th or 89th line pulse. The output pulse of M2 which is triggered in the coincidence of the front edge of line pulse loads the increment corresponding the line into the I-counter. The variable Q_{18} is true if the contents of the I-counter is not equal to zero. Then the pulses of the clock oscillator are applied to the I-counter which counts down until the contents is equal to zero again. Equal number of pulses is applied to the JI-counter which counts down for $v \leq (R+1)/2$ and counts up for $v > (R+1)/2$. Thus the JI-counter stores the length of the line segment $A_v B_v$ for each line. The horizontal coordinate of the middle point of the circle is applied as initial value to the counter.

In the coincidence of the rear edge of the line pulse the Y-counter counts up once. Thereafter the switching matrix generates the increment corresponding to the next line. Thus the value of the increment is ready when the next line pulse is applied.

At the end of the line pulse the interpretation logic of the generator is started. The interpretation logic consists of comparison logic, P- and A-counters, a number of gates, and P22 and flip-flops FF3, FF4 and FF5. P- and A-counters count the clock pulses until the contents of P- and JI-counters are equal. Then the comparison logic sets FF5 and the sweeping electron beam is brightened on the screen. At the same time FF4 is reset, through P11 and the P-counter is stopped. A-counter continues counting up. When the contents of A-counter equal 565 which is equivalent to the horizontal position of the middle point of the circle the decoding gate P20

sets FF4 and the P-counter starts again. When the P-counter has reached the state 565, FF5 is reset through P19 and the sweeping electron beam is switched off.

The next line pulse loads a new increment into the JI-counter and the operation as described is repeated. The whole procedure is valid up to the line 195. Then Q_2 becomes true and FF2 is set; Q_1 is true.

The logic variable Q_1 is false for $v \leq (R+1)/2$ and true for $v > (R+1)/2$. Q_1 determines the counting direction of the Y-and JI-counters. Q_2 prevents with gates P14 and P15 the 195th line pulse to be applied to the Y-counter.

In the odd field also the 196th line pulse is blocked with FF6, gates P16, P15 and P14. With these blockings the absolute values of the increments are made symmetrical with respect to the line $v = (R+1)/2$. Thereafter Y-counter counts down and the JI-counter up. At the line 302 in the odd field and 303 in the even field the whole circle has been written and the action is interrupted for the rest of the field.

6.2 Generated Circle Test Pattern

The generator described above has been constructed with excellent success. In Fig. 5 the generated circle test pattern is presented. It has been photographed on the screen of the television monitor Disa 93A77. In Fig. 6 and Fig. 7 the region of circumference around the points $v = 1$ and $v = (R+1)/2$ are shown magnified. The complete symmetry can be detected.

6.3 Comparisons

Another numerical method for circle generation is based on the basic geometrical properties of a circle function.[2] A set of recursive incremental algorithms are used to compute the squares of line segments $B_v C_v$ (Fig. 1). Another set of algorithms are used for the necessary square root extraction. These algorithms are based on the properties of the squares of successive integers. The quality of the circle test pattern generated by this means is comparable with the one presented in this paper.

The generator based on switching matrices has the advantage of simpler line information generation logic: combinational circuits with minimum of sequential logic vs. relatively complex sequential computing unit. This advantage makes the generator preferable to the computing generator in simplicity of circuitry and to many other methods used in performance.

7 Conclusions

In this paper a new method for the television test circle generation has been presented. The positions of the points on the circumference are determined by numerical means based on the use of switching matrices as the source of line information. The corresponding generator configuration has also been given.

The accuracy and stability of the method is found to be excellent. The simplicity of the method of line information generation over computational methods has made it especially favourable for circle pattern generation.

Fig. 5. Generated circle test pattern.

Fig. 6. Structure of the circle in the region around the point $v = 1$.

Fig. 7. Structure of the circle in the region around the point $v = (R + 1)/2$.

8 Acknowledgments

9 References

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10 Appendix 1 Error Considerations

10.1 *Synchronous Interpretation*

The locations of the circle points B_v and D_v on line v in Fig. 8, where the sweeping electron beam is switched on and off, respectively, is determined by the period of a high-frequency oscillator. The period T is selected in such a way to make the horizontal and vertical resolutions equal. Then the frequency of the oscillator is

$$f_c = 1/T = \frac{\alpha . N}{H - a} = 15 \text{ MHz},$$

where

α is aspect ratio, 4/3,

N is the number of lines belonging to the image, 585,

H is the line period, 64 μs and

a is line blanking period, 12 μs.[7,8]

If the interpreting clock oscillator is synchronized with the line frequency, the relative positions of A_v and the front edge of clock signal are steady at the beginning of every line. Then the error caused by interpretation is zero and the error in the generated circle pattern is completely due to the line information generation method.

10.2 *Asynchronous Interpretation*

If the interpreting clock oscillator is running free, the front edge of the first clock pulse coming after the rear edge of line pulse can be situated anywhere in the time interval 0–T (Fig. 8). Then the maximum error in the
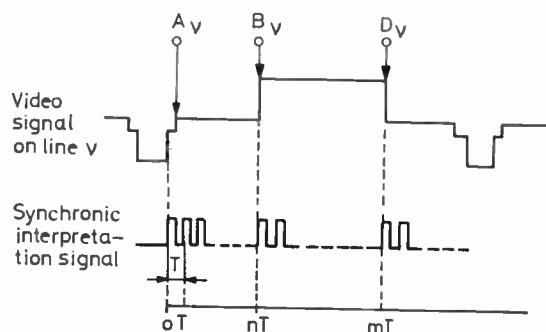
Fig. 8. Synchronous interpretation.
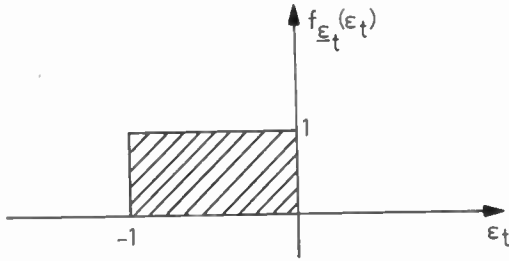
Fig. 9. Density function of ε_t.
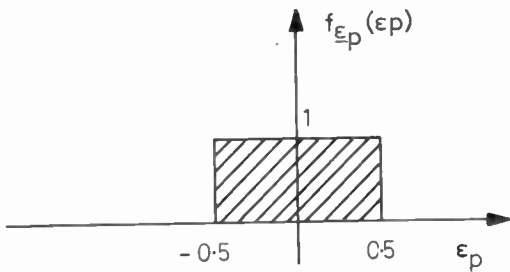
Fig. 10. Density function of ε_p.
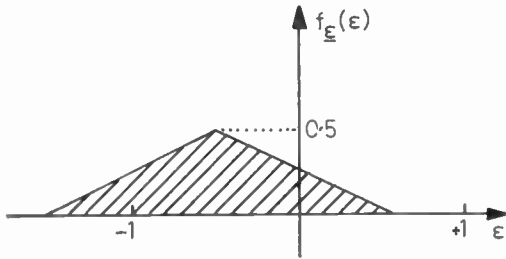
Fig. 11. Density function of ε.

locations of B_v and D_v is one unit length. It can be supposed that the error ε_t caused by the asynchronous interpretation is uniformly distributed in the range $(-1, 0)$. In Fig. 9 the density function of the random variable ε_t is presented.

It is supposed that the error ε_p caused by line information generation method is approximately uniformly distributed in the range $(-0.5, +0.5)$ (Fig. 10).

Then the total error ε in the circle writing is the sum of partial errors:

$$\varepsilon = \varepsilon_t + \varepsilon_p \qquad (18)$$

The density function of ε is according to Papoulis[9]

$$f_\varepsilon(\varepsilon) = \int_{-\infty}^{\infty} f_{\varepsilon_p \varepsilon_t}(\varepsilon - \varepsilon_p, \varepsilon_p) d\varepsilon_p, \qquad (19)$$

where $f_{\varepsilon_p \varepsilon_t}(\varepsilon_p, \varepsilon_t)$ is the joint density function of the variables ε_p and ε_t. Because ε_p and ε_t are independent random variables, $f_{\varepsilon_p \varepsilon_t}(\varepsilon_p, \varepsilon_t)$ equals

$$f_{\varepsilon_p \varepsilon_t}(\varepsilon_p, \varepsilon_t) = f_{\varepsilon_p}(\varepsilon_p) f_{\varepsilon_t}(\varepsilon_t). \qquad (20)$$

Formulae (19) and (20) yield with some manipulation

$$f_\varepsilon(\varepsilon) = \begin{cases} 0.5(\varepsilon + 1.5) & \text{for} \quad -1.5 \leq \varepsilon \leq -0.5 & (21a) \\ 0.5(-\varepsilon + 0.5) & -0.5 \leq \varepsilon \leq 0.5 & (21b) \\ 0 & \text{otherwise.} & (21c) \end{cases}$$

In Fig. 11 the density function of total error ε is presented. Mean and variance of the total error are

$$y = 0.5$$
$$\sigma^2 = 0.17.$$

The Author

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Correlator design for flow measurement

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Based on a paper presented at a Colloquium on Correlation held in London on 25th November 1971

SUMMARY

General purpose correlators are too expensive for flow metering applications and do not give a velocity measurement directly. Two methods of simplification can be used, quantization and what the authors have termed the two-point difference method. Some of the implications and advantages of these techniques are discussed.

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1 Introduction

In 1961 Butterfield, Bryant and Dowsing published a paper[1] describing a new method of measuring the speed of steel strip in a rolling mill. The method had three special features:

(i) The transit time of the strip between two points was measured by cross-correlation of the outputs of sensors at those points.

(ii) Existing surface imperfections on the strip, instead of superimposed marks, were used to identify areas of the strip.

(iii) The sensors of these imperfections had no mechanical contact with the strip so there was no obstruction to the flow.

Appreciating the advantages and the application possibilities of this method, research teams in many parts of the world began their own investigations. Papers describing their results have appeared regularly ever since.[2−7]

A correlation flow meter has two basic components, a transducer (sensor) system and a correlator. It is with the second of these that we are concerned here.

General-purpose correlators are too expensive for flow metering applications and do not give the velocity measurement directly. In the design of a flow (meter) correlator two methods of simplification are used, quantization, and what the authors have termed the two-point difference method. Some of the implications and advantages of these techniques will be discussed.

2 Theory of Flow Measurement by Cross-Correlation

A mathematical model for the simple correlator flow meter has been developed in the Appendix, giving

$$y(t) = \int_0^\infty b(\alpha)\, x(t-\alpha)\, d\alpha + z(t),$$

and

$$R_{xy}(\tau, T) = \int_0^\infty b(\alpha)\, R_{xx}(\tau-\alpha, T)\, d\alpha + R_{xz}(\tau, T). \quad (1)$$

If a single-phase flow such as steel strip[1, 7] with point sensors and identical filters is considered, then the relationship between the correlator inputs $x(t)$ and $y(t)$ is simple time delay, d, with $b(\alpha) = \delta(\alpha-d)$.

With definite sensor sizes and uneven velocity profiles, however, the simple model is unrealistic. $b(t)$ can have finite width and be asymmetric about its maximum. The resulting broad, asymmetric cross-correlation functions have been observed by Fisher and Davies,[8] Veltman et al.[4] and Kashiwagi.[6]

Finding the mean time delay in such a case can be difficult. Fisher and Davies contend that the envelope of the correlation functions at different sensor spacings (and so presumably at different speeds) touches at the point of mean delay. Veltman et al. suggest that the 'centre of gravity' of the correlation function given by

$$\tau' = \frac{\displaystyle\int_{-\infty}^{\infty} \tau R_{xy}(\tau)\, d\tau}{\displaystyle\int_{-\infty}^{\infty} R_{xy}(\tau)\, d\tau}, \quad (2)$$

should be used. Kashiwagi is apparently more optimistic, having obtained excellent results from a two-point correlation technique.

Whatever the system the finite averaging time T causes errors in the estimate $R_{xy}(\tau, T)$, leading to errors in the estimate of time delay. Theory based on Burdic's analysis[9] suggests that the variance of the estimates of time delay d is given by

$$\text{var}(d) = \frac{K}{\left[B^3 T \frac{R_{xx}(0)}{R_{zz}(0)} \right]}, \qquad (3)$$

where B is a bandwidth term; $R_{xx}(0)/R_{zz}(0)$ is the ratio of the autocorrelations of $x(t)$ and $z(t)$ at zero delay and is thus the mean square signal-to-noise ratio of the downstream signal $y(t)$; K is a constant depending on the system and in particular on the definition of B.

Preliminary results from an experimental system have confirmed this form of dependence for varying averaging times and signal/noise ratios.

Mesch *et al.* have investigated the error from mismatched low-pass filters and, for a useful model of a flow

(a) Serial correlator.

(b) Parallel correlator.

system, have been able to quantify the intuitively obvious linear dependence of the delay time error and the difference between the time constants T_1 and T_2, as

$$\text{error in } d = \frac{T_2 - T_1}{4}. \qquad (4)$$

3 Correlator Design: Review

Three common designs of general-purpose correlators are shown in Fig. 1.

The serial method calculates $R_{xy}(\tau, T)$ for a series of values of τ in turn. It is consequently slow and usually unsuitable for on-line flow measurement.

The parallel method gives high speed at a correspondingly high cost; it is useful for experimental work but far too expensive for a flow meter.

The serial/parallel method, which is cheaper than the parallel method but not by enough, gives high speed by updating the correlation functions at a series of delay values between sampling operations. The well-known Hewlett Packard Model 3721A Correlator uses this method.[10]

Input storage, tape or digital, can speed up a serial correlator but tape systems are still slow and a digital store can be used more efficiently in the serial/parallel method. The reader is referred to References 10, 12 and 13 for a more detailed discussion of these designs.

Figure 2(a) shows an obvious method of finding the position of the peak of the correlation function estimate. Each correlation value is compared with the highest already found and the larger of the two is stored with the corresponding delay value. A polar correlator of this design has been built at the University of Bradford and has been used with some success on experimental flow rigs (Correlator Mk. I).

Another method is that used by Jordan[11] where the peak position is detected by overflow in the correlation store as shown in Fig. 2(b).

With these systems time delay is measured as $n\Delta\tau$ where n is an integer and $\Delta\tau$ is the sampling period. For the typical requirements of 1% resolution of time delay over a 10–1 range, 1000 delay increments are needed,[11] with a sampling frequency often well in excess of the ideal, Nyquist frequency.

Clearly some drastic cost reduction is necessary for a flow measurement correlator; the requirement of velocity
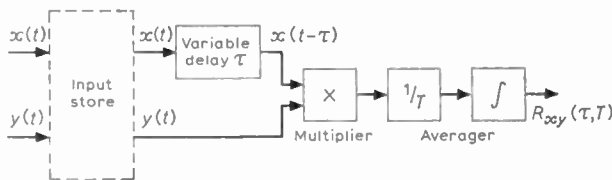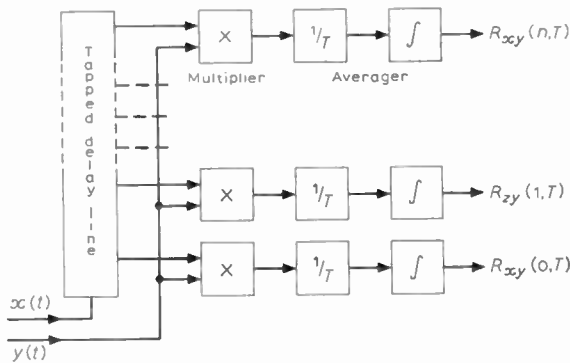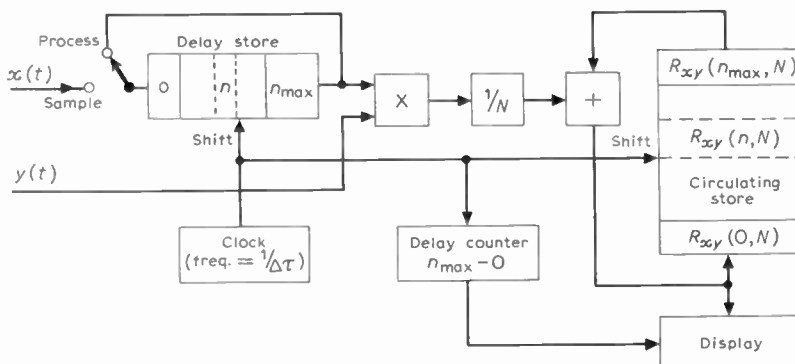
(c) Serial/parallel correlator. $R_{xy}(n, N)$ is estimate of $R_{xy}(\tau)$ from sampled data $\tau = n\Delta\tau$, $T = N\Delta\tau$ where $\Delta\tau$ is the sampling period.

Fig. 1. Correlator designs.

display is an added difficulty. A method of cost reduction which has been very successful in other correlation applications is coarse quantization of one or both inputs.
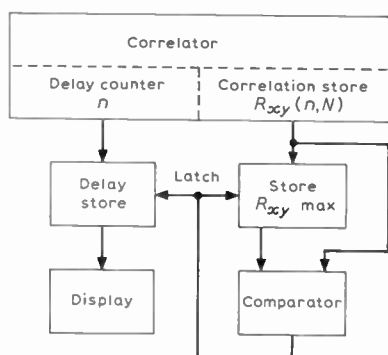
(a) Peak detection by comparison
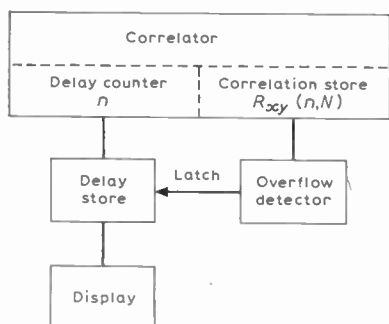
(b) Peak detection by overflow

Fig. 2. Methods of peak position detection.
The correlator is the series/parallel correlator of Fig. l(c).

4 Quantization

Theory first developed by Widrow[14, 15] has shown that fine (for example, 8-bit or 12-bit) quantization (analogue-to-digital conversion) of the inputs to a digital processor is not always necessary for accurate estimation of the statistics of the inputs. Widrow and Watts[16] realized that correlation equipment in particular would be greatly simplified if coarse quantization of one or both inputs were used. Theory and experiment demonstrated that high accuracy of measurement was still possible for many random processes. For a review of quantization theory the reader is referred to Korn[13] and for a classification, to Chang and Moore.[17]

Watts introduced the name Stieltjes correlator for one which quantizes one of the inputs to just a few levels, typically eight or less. Finnie[18] also examined this method with an alternative form of analysis and on his feasibility study was based the Hewlett Packard Model 3721A correlator.[10] This design quantizes one input to seven *levels* and the other to seven *bits*. The simplest form in this classification is known as the relay correlator and has one-bit (two-level) quantization for one channel.

A digital correlator uses coarse quantization of both channels and its simplest form (one bit) is known as a polar correlator. For Gaussian data with zero mean it can be shown[16] that the polar correlation function, ρ_P, is

related to the normalized direct correlation function, ρ_D, by

$$\rho_{Pxy} = \frac{2}{\pi} \text{ arc sin } (\rho_{Dxy}). \qquad (5)$$

A second remarkable result of Widrow's analysis is the use of 'dither'. Not all inputs to a correlator will satisfy Widrow's quantizing theorem.† However the sum of two signals will satisfy the theorem if one of the two does so independently. Carefully chosen dither signals when added to the inputs of a Stieltjes or digital correlator, will ensure that unbiased estimates of the true correlation functions are obtained.

Though equation (5) is non-linear, we observe that the maximum values of ρ_P and ρ_D occur at the same delay. Therefore, if the peak position of ρ_D is the measure of the time delay between the two inputs then so is the peak position of ρ_P. In many flow systems, therefore, a polar correlator can be used without the need for dither, whereas in applications where the shapes of the true correlation functions have significance, then the non-linear relationship necessitates the use of dither or a more elaborate quantization.

Though the estimates of correlation functions from quantized data are usually unbiased the dispersion (variance) of the estimates is increased. To compensate for this the integration (averaging) time T is increased.

It is important to know by what factor T must be increased to give the same confidence limits to an estimate from coarse quantized data as those associated with an estimate from fine quantized or analogue data averaged over time T.

This aspect has been studied by several investigators[19-21] and it appears that there is no simple solution because the relative estimation errors depend on the degree of correlation between the inputs. For low correlation coefficients, such as those which occur in signal detection, factors of $\pi^2/4$ for polar estimates and $\pi/2$ for relay estimates have been derived.

Whether the same holds for the estimates of peak position (time delay) is a problem currently under investigation.

The Stieltjes and digital correlators use simpler delay lines and multipliers and require less storage for the correlation function estimates. In a polar correlator complex analogue-to-digital converters are replaced by Schmitt triggers and multiplication reduces to a comparison of two binary digits. Such a correlator can be built for hundreds instead of thousands of pounds.

5 Difference Methods

Figure 3 shows a correlator design in which correlation functions $R_{xy}(\tau_1, T)$, $R_{xy}(\tau_2, T)$ at two delay values τ_1 and τ_2 are evaluated. From their difference is derived the input to a voltage controlled oscillator (v.c.o.) which

† As Korn observes, 'Actually, real random data cannot possibly satisfy the conditions . . . exactly. . . . Nevertheless, many real signals satisfy the quantizing theorems so nearly that excellent approximations result, and the errors of such approximations can be neatly predicted.'
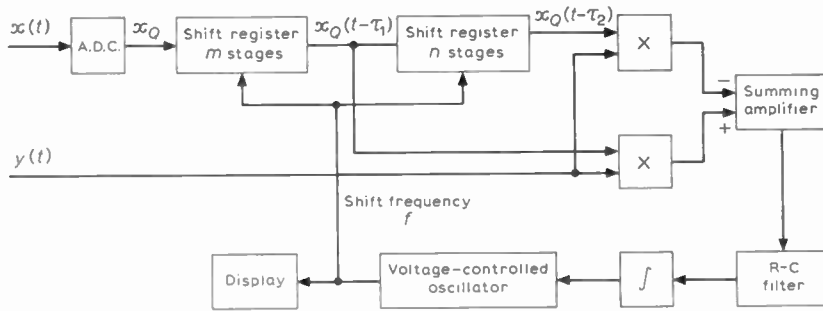
Fig. 3.

Difference method of cross-correlation for flow measurement. Delays τ_1 and τ_2 are m/f and $(m+n)/f$ respectively.)

governs the shift rate of the delay line and hence the delay values τ_1 and τ_2. τ_1 and τ_2 can be moved automatically into the balance position shown in Fig. 4. With a symmetric correlation function as shown the shift frequency is then proportional to the mean flow velocity.

To the authors' knowledge the method was first disclosed by Kashiwagi[5] but similar methods appear to have been discovered independently by Dr. Veltman's team at Delft,[22] Prof. Mesch's team at Karlsruhe,[7] Kent Instruments Ltd.[23] and McGunigle.[25]

If τ_1 and τ_2 are close together then $R_{xy}(\tau_2, T) - R_{xy}(\tau_1, T)$ is proportional (approximately) to the derivative $d/d\tau(R_{xy}(\tau, T))$ of the correlation function estimate at $\tau = (\tau_1 + \tau_2)/2$ so in the absence of a concise alternative we use the term 'two-point difference' to describe these methods.

Difference methods have obvious attractions. The system is less complex and consequently less costly than the 'multi-point' correlators and the shift pulse rate (sampling frequency) is proportional to the flow velocity (in the balance condition) so a velocity display is easily derived.

A third advantage is that the continuously variable sampling period gives fine resolution of the time delay measurement. Further, it is possible to select n_1 and n_2 (where $\tau_1 = n_1 \Delta\tau$, $\tau_2 = n_2 \Delta\tau$) so that the sampling frequency $1/\Delta\tau$ is close to the Nyquist frequency giving maximum signal usage at minimum cost.

The method also has its drawbacks. One of these, the slow response to flow rate changes, has been examined by Mesch et al.[7]

Asymmetric correlation functions may also cause difficulties but Kashiwagi's[6] results suggest that these can be overcome.

Another problem is that it is necessary to have a broad smooth correlation function in order to provide adequate difference signals under changing flow conditions. However, the best accuracy is obtained with wideband noise signals leading to narrow correlation functions. For such systems the correlation functions at more than two points are required particularly if a reasonable flow range is to be covered.

6 A Multi-Point Correlator with Controlled Sampling Frequency

A sampling frequency controller can also be added to a multi-point correlator with a peak finder such as that of Fig. 2(b) to satisfy most of the requirements of a flowmeter.

If the sampling frequency, f, is adjusted so that the measured time delay, d, is always a pre-chosen fixed number M of sampling periods $\Delta\tau$ ($\Delta\tau = 1/f$) then as

$$d = M\Delta\tau = M/f,$$

$$f \propto \frac{1}{d} \tag{6}$$

so f is proportional to the flow velocity.

The University of Bradford Correlator Mk. II, shown in Fig. 5, uses this principle. It has been designed to assist the evaluation of correlation flowmeter systems. The 128-point, polar correlation function can be displayed on an external oscilloscope with a linear (0–$127\Delta\tau$) or compressed (0–$535\Delta\tau$) delay time range. In the prototype M is 60 giving a $1\cdot7\%$ resolution of delay time. The controller can automatically adjust the sampling frequency in any 32:1 range within the limits of 7 kHz and 7 Hz. Frequency is displayed on a panel meter and is also available as an output to a digital frequency meter.
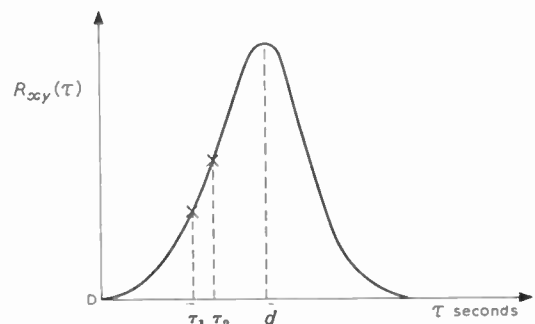
(a) Error condition $R_{xy}(\tau_2) \neq R_{xy}(\tau_1)$
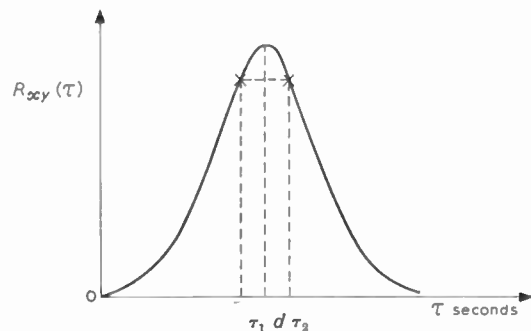
(b) Balance condition $R_{xy}(\tau_1) = R_{xy}(\tau_2)$

Fig. 4. Operation of difference correlator.
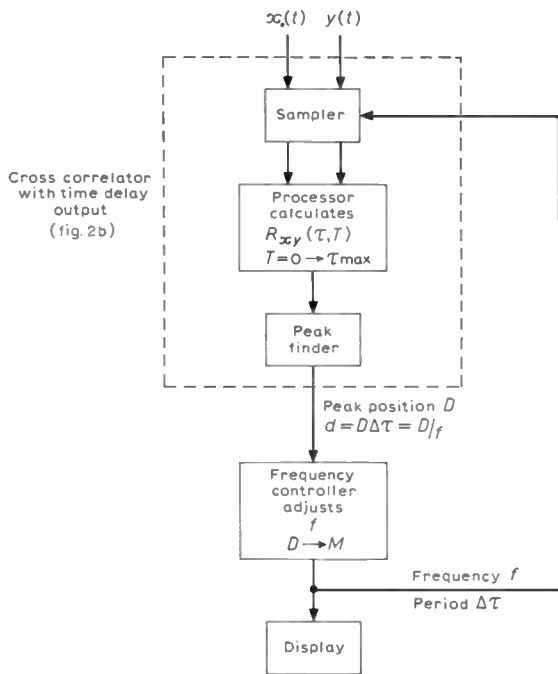
Fig. 5. Multi-point correlator with sampling frequency control (When $D = M$, $f \propto 1/d \propto$ flow velocity.)

A feature of the design is that the correlator and frequency controller are all digital and thus avoid the stability problems of analogue circuits.

The correlator has proved to be a useful alternative to the more expensive, general-purpose instruments for experimental work; but in its present form it is itself rather expensive for an industrial flowmeter. However, the oscilloscope display circuits, wide frequency range and possibly the large store would not be needed once the particular application were known, leading to quite considerable savings.

In an effort to reduce costs still further, the authors are currently investigating a design with coarse frequency control, using the principle described in this section, to give a fast response when the flow rate is changing and fine frequency control using the difference method when the flow is steady.

The experimental correlator described by Bullock[26] has this coarse/fine feature but is rather complex. The authors believe that a simplified form will be more than adequate for most flow systems.

7 Conclusions

One of the major obstacles to the industrial application of correlation flowmeters has been the high cost of the available correlators. Coarse quantization and difference methods (and doubtless modern large scale integration techniques) can overcome this problem.

Much depends on transducer development as the low cost correlators rely on highly correlated, wide bandwidth inputs for their accuracy and response time.

8 Acknowledgments

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Fig. 6. General model of correlation flow meter. $n_0(t)$, $n_1(t)$, $n_2(t)$ are noise.

10 Appendix: Mathematical Model of a Correlation Flowmeter (Fig. 6)

Assume that the sensor outputs $u_1(t)$, $u_2(t)$ are related by

$$u_2(t) = \int_0^\infty a(t, \alpha)\, u_1(t-\alpha)\, d\alpha + n_0(t),$$

where $a(t, \alpha)$ is the generalized impulse response of Bendat[24] and $n_0(t)$ is noise uncorrelated with $u_1(t)$.

Further assume that when the flow is steady $a(t, \alpha)$ is time invariant and can be written $a(\alpha)$.

The correlator inputs $x(t)$, $y(t)$ are given by

$$x(t) = \int_0^\infty h_1(\alpha)\, u_1(t-\alpha)\, d\alpha + n_1(t),$$

$$y(t) = \int_0^\infty h_2(\alpha)\, u_2(t-\alpha)\, d\alpha + n_2(t),$$

with

$$u_2(t) = \int_0^\infty a(\alpha)\, u_1(t-\alpha)\, d\alpha + n_0(t),$$

when the flow is steady.

Let $x(t) \leftrightarrow X(f)$ denote a Fourier transform pair. Then in an obvious notation, making use of the 'convolution theorem'

$$X(f) = H_1(f)U_1(f) + N_1(f),$$
$$Y(f) = H_2(f)U_2(f) + N_2(f),$$
$$U_2(f) = A(f)U_1(f) + N_0(f),$$

so that

$$Y(f) = B(f)X(f) + z(f),$$

where

$$B(f) = H_2(f)A(f)H_1^{-1}(f),$$

and

$$Z(f) = -B(f)N_1(f) + H_2(f)N_0(f) + N_2(f).$$

In the time domain

$$y(t) = \int_0^\infty b(\alpha)\, x(t-\alpha)\, d\alpha + z(t).$$

Further, the estimate of the correlation function $R_{xy}(\tau)$ is

$$\begin{aligned}
R_{xy}(\tau, T) &= \frac{1}{T} \int_0^T x(t-\tau)\, y(t)\, dt \\
&= \frac{1}{T} \int_0^T \int_0^\infty x(t-\tau)\, b(\alpha)\, x(t-\alpha)\, d\alpha\, dt + \\
&\qquad\qquad + \frac{1}{T} \int_0^T x(t-\tau)z(t)\, dt \\
&= \int_0^\infty b(\alpha)\, \frac{1}{T} \int_0^T x(t-\tau)\, x(t-\alpha)\, dt\, d\alpha + R_{xz}(\tau, T) \\
&= \int_0^\infty b(\alpha)\, R_{xx}(\tau-\alpha, T)\, d\alpha + R_{xz}(\tau, T).
\end{aligned}$$

The application of a commutated filter to the design of a frequency response analyser

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Based on a paper presented at a Colloquium on Recent Developments in Systems Performance Measurement held in London on 10th April 1973.

SUMMARY

A new structure is discussed for a frequency response analyser incorporating a four-path filter to execute the correlation process. A detailed analysis of the filter is presented. The instrument covers the frequency range 10 Hz $\leqslant f \leqslant$ 100 kHz and its advantages include higher frequency operation and increased dynamic range compared with previous designs and a voltage programmable sweep capability over a frequency range in excess of 1000 : 1.

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List of principal symbols

ω input signal frequency (rad/s)

ω_r input commutation frequency (rad/s)

ω_0 output signal frequency (rad/s)

ω_{if} output commutation frequency (rad/s)

N number of capacitors or commutator segments

T_i commutation time between segments or the dwell time on any one segment for input commutator

T_{if} as above, but for output commutator

p path index number

k number of completed revolutions of contact arm

$e(t)$ input signal

$e_p(k)$ potential difference on capacitor p after k revolutions of contact arm

α initial phase displacement

1 Introduction

In recent years switched or commutated N-path filters have attracted much attention, particularly since they have a frequency characteristic that may be used to realize a high-Q bandpass filter without inductors. The literature has, in general, concentrated almost exclusively on the theoretical aspects with little consideration given to practical application. The imbalance between theory and practice tends to obscure many of the important issues and can, in fact, prove generally misleading to the investigator who is new to the field. It is therefore the intention of this paper to examine the circumstances of one particular application and, further, to demonstrate how the favourable characteristics of the N-path filter may be exploited to advantage within the design of an instrument dedicated to the characterization of electrical networks and other systems.

2 Frequency Response Analysers

The response of a network or linear system to a sinusoidal input is expressed as an amplitude/phase relationship or, alternatively, in terms of in-phase and quadrature components. For less critical applications, within a relatively noise-free environment, a sinusoidal test oscillator, amplitude ratiometer and zero-crossing phasemeter may provide sufficient accuracy. However, with electro-mechanical systems in particular, static friction, coulomb friction, backlash, non-linear compliances, non-linear motors, interference and non-harmonically related noise sources, are all common causes of signal degradation. When working with these imperfect systems it is necessary to employ a detector which rejects, as far as possible, both harmonics and noise.

The general block diagram of a frequency response or transfer function analyser designed as a dedicated instrument is shown in Fig. 1. The system or network under test is excited by a sinusoidal source and the unknown output is correlated both with the system input and a signal in quadrature with this input. Thus, if the normalized input is sin ωt and the output is $V \sin (\omega t + \phi)$,
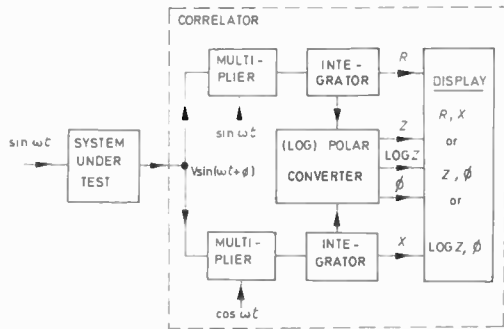
C. J. PAULL and W. A. EVANS

Fig. 1. System transfer function analyser.

the averaged outputs after N cycles are:

$$\frac{1}{NT}\int_0^{NT} V \sin(\omega t + \phi) \sin \omega t \, dt = \tfrac{1}{2}V \cos \phi = \frac{a}{2}$$

and

$$\frac{1}{NT}\int_0^{NT} V \sin(\omega t + \phi) \cos \omega t \, dt = \tfrac{1}{2}V \sin \phi = \frac{b}{2}$$

where

$$V = a^2 + b^2 \quad \text{and} \quad \phi = \tan^{-1}\frac{b}{a}$$

Harmonics due to non-linearities are rejected since:

$$\frac{1}{NT}\int_0^{NT} V \sin n(\omega t + \phi) \frac{\sin}{\cos} \omega t \, dt = 0 \quad n = 2, 3, 4 \ldots$$

The dependence of the noise response upon integration time has been investigated by Elsden and Ley[1] who have also published typical curves for both the sine and cosine channels.

In consequence of the above considerations, the necessary ingredients for such an instrument must include an excitation source (sinusoidal), a correlator and a display unit. For frequencies typically in the range $10^{-4} < f < 10^3$ Hz, the generation of low distortion sinusoids by digital methods is relatively straightforward and for this reason instruments adopting an all-digital approach to the generation and processing of signals have predominated over the range.[1] Above these frequencies the high sampling rates require faster and more expensive analogue-to-digital (a/d) and digital-to-analogue (d/a) converters, as well as one or more high speed multipliers, even to secure a relatively modest increase in operating frequency. The alternative approach calls for an additional phase-locked oscillator and high frequency sampler unit to effectively extend the range of a lower frequency instrument. In either case, the a/d and d/a conversion processes with typically eight-bit word lengths restrict the dynamic range and overall accuracy of the result. The need, therefore, to explore alternative means of achieving higher operating frequencies is a compelling one, particularly for the smaller user, where the initial capital expenditure required for the existing alternatives may be prohibitive.

3 Hybrid Design Principles

The adoption of a substantially different hybrid approach to the design of a frequency response analyser

offers distinct advantages both in terms of performance and cost reduction. The mode of operation of such an instrument is most readily appreciated by following the forward signal path, as indicated in Fig. 2. A function generation unit originates the triangle and two-phase square wave outputs, as shown, the frequency being continuously variable within each switched decade from 10 Hz to 100 kHz. Non-reactive shaping of the triangle is used to synthesize the sinewave which is then buffered out to the external terminals by a power amplifier. When an unknown network or system is connected between this source and the detector terminals, the forward signal path is returned within the instrument via an input buffer amplifier. The low impedance output from this amplifier is suitable for driving a special-purpose N-path filter. This filters the signal as well as reducing it to a constant frequency of 125 Hz. Since the relative amplitude and phase information is retained through the frequency translation process, low-pass filtering yields an undistorted, noise-free, sinusoid on which precise measurements can readily be taken. Amplitude and phase are displayed separately on two suitably calibrated moving-coil meters, using conventional low frequency precision detector circuits.

The principal three sections:

(a) The signal source
(b) Correlator (N-path filter)
(c) Display unit

will now be considered in greater detail and criteria established that will enable the total performance of the complete instrument to be assessed.

4 The Signal Source

If the commutating filter discussed in the next section is to behave in the ideal manner, the generation of the four-phase clock or reference waveforms must be precisely defined in terms of frequency and phase with respect to the primary sinewave source. This source could be a conventional RC or LC oscillator of good harmonic purity provided four equi-spaced phase positions could be accurately located on the sinewave output. The 0° and 180° phase angles might be accurately identified using zero crossing detection. This leaves the quadrature phase positions which could be located only approximately using direct acting peak detection circuitry. Possible alternative solutions could be found with a two-phase oscillator or a frequency doubler. A third, preferred, solution avoids these difficulties by generating the signal with a triangular relaxation oscillator. The triangular waveshape results from alternately charging and discharging a capacitor from a current source and sink of equal magnitude. A hysteresis switch operates directly on the triangle output to define both positive and negative maxima for permitted voltage excursions. A quadrature square wave A is therefore immediately available at the hysteresis switch output without recourse to any additional circuitry of possibly indefinite precision. The in-phase square wave B is generated from the triangle wave using a high speed comparator connected as a zero crossing detector. The required four-phase commutating filter clock waveforms are obtained by suitable gating
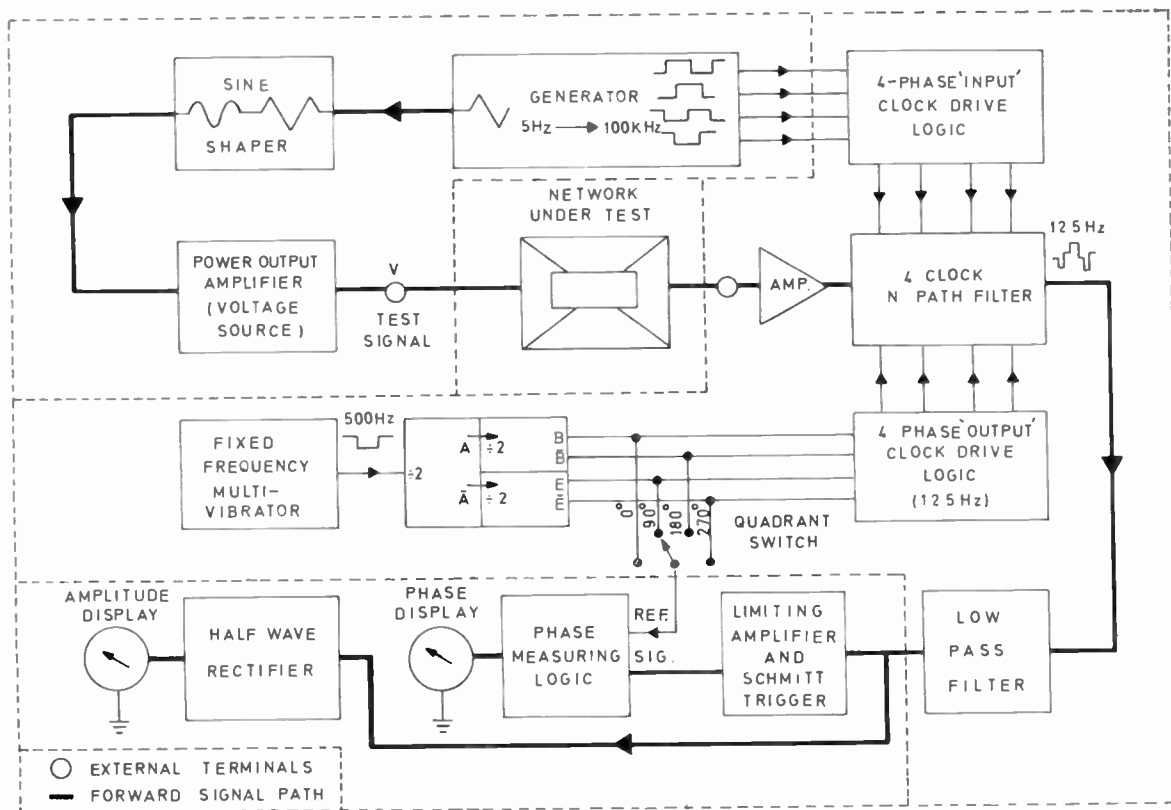
Fig. 2. Frequency response analyser.

to yield the following logical combinations AB, $\overline{A}B$, $\overline{A}\overline{B}$ and A$\overline{B}$ respectively.

The triangle relaxation oscillator has a further virtue that a sweep capability may be conveniently incorporated by voltage control of the current source and sink. Amplitude integrity of the triangle waveform is maintained during sweeping due to the 'anti-bounce' characteristics of the circuit. This is not the case for the LC and RC circuits.

4.1 Sine Wave Synthesis Technique

A variety of non-reactive shaping techniques have been developed for synthesizing a sinusoid voltage or current waveform from a triangular input. The methods may be conveniently subdivided into two categories based either on the use of the non-linear characteristics of an active device[2], or a piecewise linear approximation using diode resistor networks[3]. The former approach is not particularly attractive due to the unavoidable problems associated with adequately defining the operating region, particularly with temperature variation. Piecewise linear approximation methods, though widely used, suffer from certain limitations that include zero crossover and peak distortion as well as a pronounced susceptibility to second harmonic distortion. A careful setting-up procedure is normally obligatory if good spectral purity is to be achieved.

A new system developed by the authors operates by approximating a sinewave to any desired accuracy by summing an appropriate number of suitably proportioned trapezoids[4]. The final circuit was constructed in a fully temperature-compensated form, providing exceptional stability and purity of waveform. For a seven-segment approximation, the calculated resistance values gave a measured total distortion figure below 0.2% without invoking a complex setting-up procedure. Where the distortion at the detector input is also of this order, it is subsequently shown in Table 1 that the phase and amplitude reading errors will be less than 0.1% and 0.1° respectively. In practice, the harmonic content at the detector may exceed that at the source; for example, in the testing of a high-pass filter. Nevertheless, it is most unlikely that the resultant errors would become unacceptable.

4.2 Source Impedance

In the normal operating mode, the output impedance of the sinewave source must be kept as low as possible to approximate a voltage source, since the instrument is primarily designed to measure the complex ratio of voltages at the detector and source, respectively, under conditions of minimal loading at the detector. To plot the h, y, z, or r parameters for any linear network, it is necessary to make available both voltage and current sources to be used in conjunction with either a voltage or current detector input.

In practice, a very low source impedance ($< 1\,\Omega$) can only be realized by using overall feedback around the

final stage of the output amplifier. It then becomes possible that the amplifier will oscillate under certain conditions of capacitive loading due to the introduction of a second pole into the close loop transfer response of the amplifier. It is common practice under these conditions to specify a maximum permissible value for capacitive loading. Similar problems arise when designing a current output. Nevertheless, a reasonable compromise solution resulted in nominal voltage and current outputs of $< 5\,\Omega$ and $> 100\,\text{k}\Omega$ respectively, stable under all conditions of loading. The source voltage is arranged to be continuously variable throughout four switched decade ranges from 3 mV to 3 V r.m.s. max. The inclusion of a variable signal amplitude control allows measurements to be referenced to any arbitrary datum, as may frequently be desirable.

5 Correlator (Four-path Filter)

The interactive relationships between the system blocks shown in Fig. 2 may be considered to be fixed by the character of the switching four-path filter. This is reduced to its essentials in the schematic of Fig. 3(a) from which it can be recognized as a series-switched, four-path commutating filter driven from a voltage source V_{in}. Since the load impedance is specified as near infinite, no charge may be transferred to or from any channel capacitor except via the input commutator.

The commutators, shown in their mechanical equivalent form, incorporate moving contact arms which rotate at ω_0 and ω_{if} radians per second respectively. A break-before-make action is necessary to prevent shorting of adjacent segments. The four channels or paths, being identical, consist of single-pole low-pass filter sections of time-constant RC. In normal operation ω_0 is made equal to ω_{in} of the input signal and therefore each low-pass filter section is exposed to the same quarter sinusoid on successive cycles. For the remaining 75% of each cycle the channel inputs are left floating. Since, ideally, the capacitors can lose or gain charge only when the commutating switch is closed, the voltage that builds

(a) Four path filter with mechanical commutators

(b) Equivalent circuit for single channel.

Fig. 3. The commutated filter

up on each capacitor must be a true average over that quarter cycle for which the section is active. The second commutating switch effectively 'looks at' the quasi-d.c. capacitor voltages in sequence to generate an output waveform. This waveform appears as a crudely quantized approximation to the sinusoidal input signal.

When the input frequency differs from ω_0, the average value of the input voltage will change with each successive time interval that the capacitor is connected. Because of this lack of synchronism between the input and switching frequencies, the voltage across any capacitor varies at a rate equal to the difference frequency. However, since the charging rate is limited by the RC time-constant, this capacitor voltage becomes smaller and smaller, approaching zero as the difference between signal and switching frequencies becomes larger. In effect, the low-pass characteristic is translated into a bandpass characteristic centred about ω_0.

5.1 Input Commutator

An approximate general analysis appropriate to this type of commutated filter has been provided by Le Page et al.[5], where the input signal is assumed to be quantized and to remain constant over that time interval for which the commutator brush makes contact with any one segment. For a sinusoidal input, the quantized signal amplitude is set equal to the true amplitude at that instant in time when the brush is at the centre of the segment. However, for a small number of segments this approximation leads to an appreciable error in the steady state gain of the filter. This condition is therefore better analysed by adopting the average signal amplitude over the relevant commutation interval, since this is the manner in which the filter physically responds. An analysis of the filter is given in Appendix 1 where it is shown that the capacitor potential $e_p(k)$ in the steady state is given by:

$$e_p(k) = E_0 \sin\left[\omega(p+kN)T_i + \alpha + \theta\right]$$

where

$$E_0 = \frac{EN}{\pi} \sin\frac{\omega T_i}{2} \frac{q}{\sqrt{q^2 + 2(1-q)(1-\cos \omega N T_i)}}$$

and

$$\theta = -\tan^{-1}\left[\frac{\sin \omega N T_i}{\frac{1}{(1-q)} - \cos \omega N T_i}\right].$$

5.2 The Output Commutator

The output waveform is regenerated from the capacitor potential $e_p(k)$ through the action of the second commutator rotating at a frequency $\omega_{if} = 2\pi/NT_{if}$. Since the two commutators operate asynchronously, it is not practicable to obtain exact expressions for the output signal $e_0(t)$. However, for input frequencies close to the resonance frequency
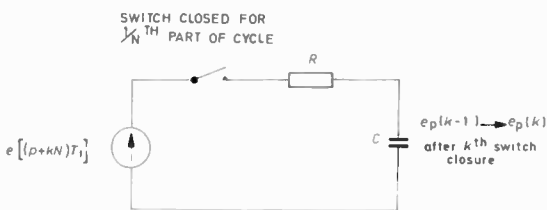
$$e_0\,t) \simeq e_p(k) = E_0 \sin\left[\omega(p+kN)T_i + \alpha + \theta\right] \qquad (1)$$

where k lies in the interval
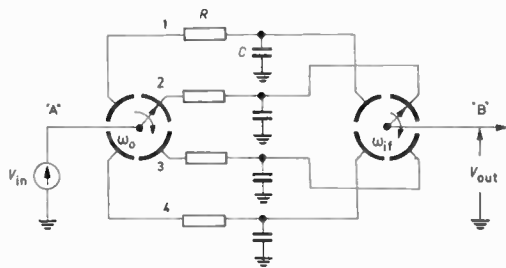
$$(lN + p - \tfrac{1}{2})T_{if} < k < (lN + p + \tfrac{1}{2})T_{if}.$$

and

$$l = \left[\frac{T_i . k}{T_{if}}\right]_{\text{integer part of}}$$

The input and output commutators are initially in-phase at $t = 0$. From equation (1) it can be appreciated that the output will more closely approximate a sinusoid as N is increased. The approximate 3 dB bandwidth of the filter may be found from the equation for E_0 when

$$q^2 = 2(1-q)(1 - \cos \omega N T_i)$$

Let

$$m = \omega N T_i = 2\pi \frac{\omega}{\omega_r}$$

Near resonance $\cos m \to 1$ and hence may be written as:

$$\cos m \simeq 1 - \frac{(\delta m)^2}{2} \quad \text{where} \quad \delta m = \frac{2\pi . \delta\omega}{\omega_r}$$

and $\delta\omega$ is a small deviation from the resonance frequency ω_r.
Thus

$$\frac{2(1-q)}{q^2} \cdot \frac{(\delta m)^2}{2} = 1$$

or

$$\delta m = \frac{q}{\sqrt{(1-q)}}$$

but since $q \ll 1$ under normal operation conditions,

$$\delta m \simeq q \quad \text{and} \quad \delta\omega \simeq \frac{1}{NRC} \text{ rad/s.}$$

At the principal resonance, the phase term θ is clearly zero.

The general theory for N-path filters embodying continuous phased modulators has been developed by Franks and Sandberg and others[6,7], who have shown that the bandpass structure of this type of filter is equivalent to folding the one-sided low-pass channel filter response about the modulation frequency. This exact theory only applies where the input (voltage) to each channel is determinate at any point in time. The 'floating' condition resulting from a simple series switching action reduces the duty cycle to $100/N\%$ and therefore may be considered to increase the effective time-constants to NRC. This is in agreement with the result obtained for a filter composed of single pole low-pass sections. For a four-path filter the bandwidth becomes:

$$B_\omega = \frac{1}{4\pi RC}$$

and the Q factor:

$$Q \simeq 2\omega_r RC$$

The Q factor can become very large ($> 10^4$) as the operating frequency is increased, without the possibilities for instability inherent in the design of a conventional active filter.

The filter output $e_0(t)$ remains substantially constant over that time interval for which the output commutator is in contact with a given segment. Thus, where the filter has only a small number of parallel paths, the subjective appearance of the output waveform shows little resemblance to a sinusoid. The amplitude of the fundamental and harmonics may be calculated for the resonance condition with a specific value of $N(=4)$ by performing a Fourier analysis. Equation (1) becomes

$$e_0(t) = \frac{E.N}{\sqrt{2\pi}} . \sin\left(\frac{p.\pi}{2} + \alpha\right); \quad p = 0, 1, 2, 3.$$

where t lies in the range

$$(LN+p-\tfrac{1}{2})T_{if} < t < (LN+p+\tfrac{1}{2})T_{if}$$

and L is a positive integer.
This analysis is carried out in Appendix 2 where the filter output is given by

$$e_0(t) = \frac{8E}{\pi^2}\left[\sin(\omega_0 t + \alpha) - \tfrac{1}{3}\sin(3\omega_0 t - \alpha) - \right.$$
$$\left. -\tfrac{1}{5}\sin(5\omega_0 t + \alpha) + \tfrac{1}{7}\sin(7\omega_0 t - \alpha) + \dots \text{etc}\right]$$

where

$$\alpha = \tan^{-1}\frac{a}{b} \quad \text{and} \quad a = \frac{2\sqrt{2}}{\pi} E \sin\alpha, \quad b = \frac{2\sqrt{2}}{\pi} E \cos\alpha$$

The value $8/\pi^2 \simeq 0.81$ represents the insertion loss.' Correct phasing of the fundamental is maintained.

As stated above, this particular filter is unusual in that the input and output modulation frequencies are not identical. By maintaining the output modulation frequency constant, the filter also serves the additional function of a constant frequency converter since, in the steady state condition, the fundamental component of the output signal appears at the second commutation frequency. The other significant filter characteristics are in no way affected. The individual channel time-constants RC, being much greater than $1/\omega_r$, ensure that any ripple on the capacitors may be neglected. The second commutator 'looks at' these quasi-d.c. voltages in sequence to generate an output signal which is a crudely quantized version of the input and for which the time scale has been expanded in the ratio ω_r/ω_{if}. The need for this frequency conversion arises directly from the character of the switching N-path filter; the unfiltered output always contains an infinite series of the harmonics of the clock frequency. Equation (7) from Appendix 2 shows these to be of constant amplitude but variable phase relationship with respect to the fundamental. An average responding meter would therefore see a signal whose amplitude was phase dependent. Similarly, since phase difference is defined only for sinusoids, phase readings taken on this quantized output would be almost meaningless. Thus, for the application under consideration, it is essential that all harmonics be heavily attenuated. This can only be satisfactorily achieved by using conventional low-pass or bandpass filters having well-defined frequency and phase response characteristics. Consequently, the secondary commutation frequency must be maintained constant. The actual frequency chosen is purely a matter of practical convenience. Where filtering has reduced the

quantized signal waveform to its fundamental component, the phase may be determined in the conventional manner, except that the reference phase is now derived from the electrical angle of the rotating commutator contact arm.

5.3 The Low-pass Filter

Whereas there is no even harmonic content in the filter output, the relative amplitude of any odd harmonic component is inversely proportional to its order; 33% third harmonic, 20% fifth, etc. Thus, by using low-pass filtering alone, it is sufficient to design for adequate rejection of the lowest-order harmonic. In practice, a rejection ratio of 50 dB (fundamental to third harmonic) is quite adequate since this reduces the total harmonic content to approximately 0.1%. This was realized using a sixth-order Chebyshev filter with 1 dB ripple in the passband and a cut-off frequency of 205 Hz. In this instance, the phase shift at a frequency $\omega_{if}(=125 \text{ Hz})$ is 180°. The constant phase delay in the forward signal path is readily compensated by a simple phase reversal in the reference channel before both signals are applied to the two inputs of the zero crossing phasemeter.

5.4 Frequency Translation of Harmonics

For the present application, where the output is filtered to remove the second and higher-order harmonics, signal harmonics are troublesome only insofar as they may be frequency-translated down to the fundamental (clock) frequency. This particular frequency translation process occurs in all multi-path filters employing switched modulators. Further, since the condition $\omega < \frac{1}{2}N\omega_r$ is not satisfied with a four-path filter for any harmonic other than the fundamental, the unwanted responses are, in general, both phase and frequency dependent. The magnitudes of the responses are therefore most readily obtained by direct computation in the manner described in Appendix 3. The worst case amplitude/phase errors resulting from the presence of 10% distortion at the third, fifth, or seventh harmonic are given in Table 1. To a good approximation, the errors are proportionately reduced with less distorted signals.

Table 1

Harmonic	Amplitude error (%)	Phase error (degrees)
3rd	±3·3	±1·9
5th	±2·0	±1·2
7th	±1·4	±0·8

5.5 Electronic Switching

In the interests of clarity, the discussion on commutating filters has so far presumed that mechanical switching elements are employed. Nevertheless, most forms of electronic switching would prove to be greatly superior both in terms of cost and performance. The first series-switched filter to be constructed incorporated p-channel enhancement mode m.o.s.f.e.t.s as possibly the best replacement for the mechanical commutator. The switching characteristics of these devices are now well known and it is not proposed to elaborate further
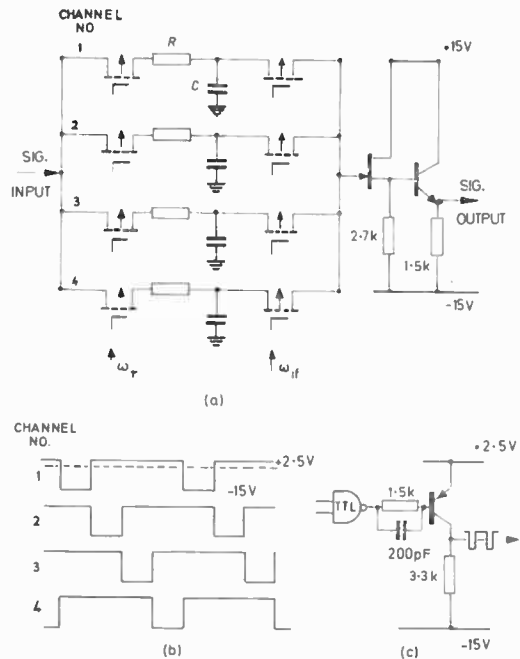
Fig. 4. Filter using m.o.s. f.e.t. switching elements.

here, except to stress that only a very small capacitive current can flow in the gate connexion. This, together with the absence of any source to drain offset voltage, ensures that the device can be accurately represented at medium and low frequencies as an ideal single-pole switch with some added series resistance equal to the small-signal source to drain 'on' resistance $R_{DS(on)}$. To obtain a commutating action using these devices, the circuit may be constructed as in Fig. 4(a) where the four-phase pulse waveforms of Fig. 4(b) determine the duration of each switching interval. Since each device is turned on only when the gate is held negative, it is clear from the Figure that a continuous commutating action is effectively simulated at both input and output. Each channel is connected through an electrical angle of 90° ($\pi/2$ radians). One form of analogue switch driver capable of providing the necessary high speed switching action at the input has been described elsewhere.[8] The output commutating ring, operating at a constant low frequency, will tolerate relatively slow switching waveforms and the simple driver circuit of Fig. 4(c) has proved quite adequate.

5.6 Detector Input

The detector input, which may be either d.c. or a.c. coupled, has an impedance characteristic similar to that of a standard oscilloscope, i.e. 1 MΩ in parallel with approximately 20 pF.

At the detector, the bandwidth of the filter may be varied according to the conditions under which it is anticipated that the instrument will operate. Narrow bandwidths offer the greatest noise rejection, but also increase the delay between measurements since build-up and decay times are directly proportional to the overall system Q factor. A CR product of approximately 80 ms gives a convenient 3 dB bandwidth of 1 Hz. Because of the unusual mode of operation of the filter,

where the input and output commutators operate asynchronously, it is shown in Appendix 4 that an effective phase advance will be introduced into the signal channel when operating at a low Q factor. From this analysis it is apparent that, to maintain phase accuracy within one degree, a Q factor of of at least 20 must be stipulated. Alternatively, a low-frequency phase compensation network may be employed to give a partial cancellation of the phase error.

6 Display Modes

As depicted in the system block diagram of Fig. 2, the instrument provides a display of signal amplitude and phase difference on two front-panel-mounted meters calibrated to an accuracy of 0·5% full-scale. A quadrant switch is provided to increase the phase resolution whilst a polarity switch also allows angles to be read as degrees lag or lead. The amplitude detector is designed using a standard circuit comprising a high-gain operational amplifier with diodes connected in the feedback loop to give a precision rectification. In addition to the above, two coherent detectors operating on the filtered output provide an option for the simultaneous display of the in-phase and quadrature signal components. The reference switching inputs to the coherent detectors are taken directly from the commutation logic and polarities may be reversed if required to give positive meter readings or voltage outputs in any quadrant.

7 Instrument Advantages

The advantages of the instrument may be summarized under the following headings:

(i) *Dynamic Range:* With the elimination of the amplitude quantization processes, the dynamic range is limited only by the switching noise in the filter. Tests on the prototype model suggest a figure approaching 80 dB over the frequency range. This compares very favourably with the all-digital analyser where the quantization noise may be less than 46 dB down on the signal. The importance of dynamic range is most apparent when resolving phase differences on small signals or, alternatively, when taking measurements on signals deeply buried in non-coherent noise.

(ii) *Sweep Capability:* Particular attention to the design of the triangular relaxation oscillator has made possible a voltage programmable linear sweep capability extending over a frequency range in excess of 1000:1. This is particularly valuable where continuous phase-amplitude, Bode and Nyquist plots are to be executed. Low impedance voltage outputs proportional to frequency, phase, and amplitude (modulus or resolved components) are simultaneously available to drive any standard X-Y graph plotter.

(iii) *Cost Advantage:* The hybrid techniques employed throughout the design are intended to give a near optimum reduction in total circuit complexity without sacrificing measurement accuracy or general versatility.

(iv) *Frequency Coverage:* The range of frequencies presently covered by the instrument have extended operation at higher frequencies by at least an order of

magnitude over existing all-digital instruments not employing phase-locked oscillators and sampler units.

8 Accuracy

As a check on the accuracy of the prototype instrument, comparative measurements have been taken on a selection of two-port filter networks, the standard of comparison being a ratio arm bridge with an accuracy of 0·1% of reading. The results obtained indicated excellent phase and amplitude linearity exceeding the calibration accuracy of the indicating meters. The setting-up procedure consists simply of zero and full-scale adjustments. No standard networks need be employed as both the 100% amplitude and the zero phase conditions are established using a direct link between source and detector, whilst the 90° phase condition is set by advancing the quadrant select switch through 90°. The circuit design is such that only the occasional recalibration check should be necessary through the life of the instrument.

Provided that the input amplifier is not overloaded, the accuracy of the instrument is generally independent of the presence of broadband noise or other sources of non-harmonically related interference. However, where the noise level within the acceptance band of the instrument becomes significant, the meter readings fluctuate in a random manner around the true value. The effect of overloading the input amplifier with interfering noise sources is to reduce the apparent signal amplitude. Phase angle is not directly affected.

9 Conclusion

The completed instrument as described in this paper has been developed to operate over the frequency range 10 Hz to 100 kHz with the expectation that developments in the techniques of hybrid digital filtering will extend the upper frequency limit to at least 1 MHz. The low end of the frequency spectrum can also be extended up to the point at which signal averaging times become excessive. With regard to the question of harmonic sensitivity, the instrument is clearly inferior to the Fourier integral analyser. However, this only becomes a material consideration when performing measurements on systems exhibiting gross non-linearities due to causes such as friction, backlash, or quantization effects. With the exception of the latter instance, such systems are normally encountered at the lower frequencies where the all-digital approach is, for other reasons, already acknowledged as superior. In other cases, it is normally quite acceptable to reduce the amplitude of excitation to the point at which the unknown network or system may be treated as piecewise linear. The reduction in cost should make this type of instrument available to a much wider range of users who have hitherto had to combine such items of test equipment as have been available to obtain results of possibly uncertain accuracy and limited scope.

10 Acknowledgment

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12 Appendix 1: Filter Analysis

In practice, the filter operates as a high-Q system, so that the simplification $1 - \exp(-T_i/RC) \simeq T_i/RC$ can be made.

Otherwise, the theory is generalized in respect of an initial phase displacement α and differing commutation frequencies f_r and f_{if}, the latter being dictated by the specific application in which the filter is to be employed.

Consider the potential build-up process (shown in Fig. 3(b)) starting at time $t = 0$ when all capacitors are initially discharged. The general difference equation governing the charging of the pth capacitor during the kth revolution of the commutator arm may be written as[5]

$$e_p(k) = e_p(k-1) + \{e[(p+kN)T_i] - e_p(k-1)\}T_i/RC$$

where the charging process is assumed to be substantially linear due to the high-Q mode of operation and $e[(p+kN)T_i]$ is the (constant) input signal over the time interval $(p+kN-\frac{1}{2})T_i \leqslant t \leqslant (p+kN+\frac{1}{2})T_i$, leading to the terminating series

$$e_p(k) = q \sum_{r=0}^{k} (1-q)^r e[(p+kN-rN)T_i]; \quad q = T_i/RC \quad (2)$$

With a sinusoidal input, the true average value for $e(p+kN-rN)T_i$ is derived as:

$$e[(p+kN-rN)T_i] = \frac{E}{T_i} \int_{(p+kN-rN-\frac{1}{2})T_1}^{(p+kN-rN+\frac{1}{2})T_1} \sin(\omega t + \alpha)\,dt$$

$$= \frac{EN}{2\pi}\Big\{ \{\cos[\omega(p+kN-rN-\frac{1}{2})T_i + \alpha] -$$

$$- \cos[\omega(p+kN-rN+\frac{1}{2})T_i + \alpha]\Big\}$$

$$= \frac{EN}{\pi}\Big\{ \sin[\omega(p+kN-rN)T_i + \alpha] \cdot \sin\left(\frac{\omega T_i}{2}\right)\Big\}$$

This may be written as the imaginary part of

$$\frac{EN}{\pi} \sin\left\{\frac{\omega T_i}{2}\right\} \exp\{j[\omega(p+kN-rN)T_i + \alpha]\}$$

Equation (2) now becomes:

$$e_p(k) = \frac{qEN}{\pi} \sin\frac{\omega T_i}{2} \exp\{j[\omega(p+kN)T_i + \alpha]\} \times$$

$$\times \sum_{r=0}^{k} (1-q)^r \cdot \exp\{-j(\omega N T_i r)\}$$

For the steady state condition let $k \to \infty$ then

$$e_p(k) = \frac{qEN}{\pi} \sin\left[\frac{\omega T_i}{2}\right] \times$$

$$\times \exp\{j[\omega(p+kN)T_i + \alpha]\} \frac{1}{1-(1-q)\exp[-j(\omega N T_i)]}$$

Separating the imaginary component

$$e_p(k) = E_0 \sin(\omega(p+kN)T_i + \alpha + \theta) \quad (3)$$

where

$$E_0 = \frac{EN}{\pi} \left[\frac{\omega T_i}{2}\right] \frac{q}{\sqrt{q^2 + 2(1-q)(1-\cos \omega N T_i)}} \quad (3a)$$

and

$$\theta = -\tan^{-1}\left[\frac{\sin \omega N T_i}{\dfrac{1}{(1-q)} - \cos \omega N T_i}\right] \quad (3b)$$
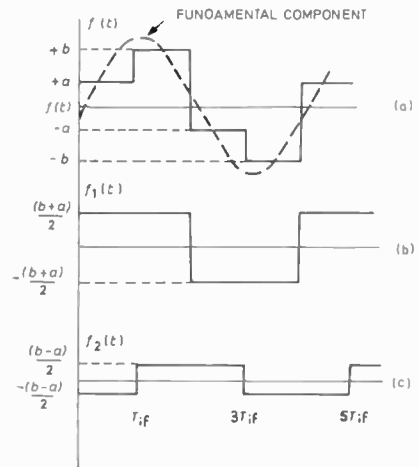

Fig. 5. Components of output waveform.

13 Appendix 2: Harmonic Content

To analyse the output waveform into its fundamental and harmonics, consider the general waveshape shown, as in Fig. 5(a). In one period, the four amplitudes a, b, $-a$, $-b$, follow consecutively and are of equal time duration T_{if}. This waveform may therefore be resolved into the two square waves $f_1(t)$ and $f_2(t)$ of Fig. 5(b) and (c), the amplitudes being $(b + a)/2$ and $(b-a)/2$ respectively. Since the general form of the Fourier series for a square wave of amplitude A is

$$\frac{4A}{\pi}\left[\sin \omega_0 t + \Phi) + \tfrac{1}{3}\sin 3(\omega_0 t + \Phi) + \right.$$

$$\left. + \tfrac{1}{5}\sin 5(\omega_0 t + \Phi) + \ldots\right]$$

where Φ is the phase displacement at the positive-going

edge, it may be shown that

$$f_1(t) = \frac{2(a+b)}{\pi} \sum_{m=0}^{\infty} \frac{1}{(2m+1)} \times$$

$$\times \sin\left[(2m+1)\left(\omega_0 t + \frac{\pi}{4}\right)\right] \quad (4)$$

$$f_2(t) = \frac{2(b-a)}{\pi} \sum_{m=0}^{\infty} \frac{(-1)^{m+1}}{(2m+1)} \times$$

$$\times \cos\left[(2m+1)\left(\omega_0 t + \frac{\pi}{4}\right)\right] \quad (5)$$

Summing the individual harmonics and simplifying, using the standard trigonometrical relationships gives:

$$f(t) = f_1(t) + f_2(t)$$

$$= \frac{2\sqrt{2}}{\pi}(a^2+b^2)^{\frac{1}{4}}[\sin(\omega_0 t + \alpha_0) -$$

$$-\tfrac{1}{3}\sin(3\omega_0 t - \alpha_0) - \tfrac{1}{5}\sin(5\omega_0 t + \alpha_0) +$$

$$+\tfrac{1}{7}\sin(7\omega_0 t - \alpha_0) + \dots \text{etc.}] \quad (6)$$

where

$$\alpha_0 = \tan^{-1}\frac{(a-b)}{(a+b)} + \frac{\pi}{4} = \tan^{-1}\frac{a}{b}$$

However, equation (6) describes the waveform of Fig. 5(a) when

$$\frac{2\sqrt{2}}{\pi}E\sin\alpha = a$$

$$\frac{2\sqrt{2}}{\pi}E\cos\alpha = b$$

and hence

$$\alpha = \tan^{-1}\frac{a}{b} = \alpha_0$$

Therefore

$$e_0(t) = f(t) = \frac{8E}{\pi^2}[\sin(\omega_0 t + \alpha) - \tfrac{1}{3}\sin(3\omega_0 t - \alpha) -$$

$$-\tfrac{1}{5}\sin(5\omega_0 t + \alpha) + \tfrac{1}{7}\sin(7\omega_0 t - \alpha) + \dots \text{etc.}] \quad (7)$$

The filter therefore has an insertion loss of approximately 0·81 at the fundamental resonance but maintains the phase integrity of the signal.

14 Appendix 3: Harmonic Distortion

The 4-path commutating filter gives an unwanted response at the fundamental frequency for all odd harmonics of the clock frequency present at the input. The magnitude of the resulting amplitude/phase error is a function of two phase variables—the harmonic to fundamental and fundamental to reference phase differences. The solution for any combination of these variables may be obtained by computation. Figure 6 shows the two components of a distorted signal

$$\sin(\omega t + \Phi) - \text{fundamental (normalized)}$$

$$V_n \sin(n\omega t + n\Phi + \Phi_n) - \text{harmonic}$$

where the phase displacement Φ_n of the nth order harmonic is measured relative to the fundamental at the positive-going zero crossings. The fundamental itself is
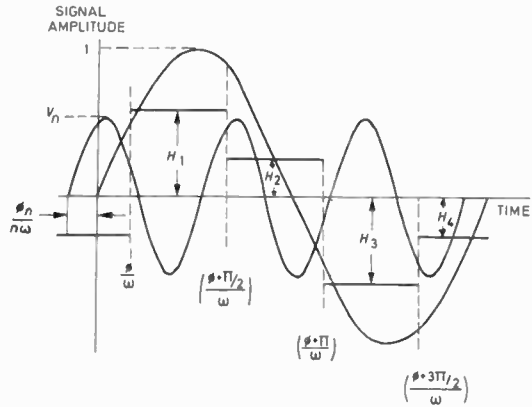
Fig. 6. Harmonic distortion.

phase advanced by Φ radians with respect to the reference. Consequently, the mean voltage H_1 obtained over the first commutation interval may be calculated as:

$$H_1 = \frac{2}{\pi}\left[\sin\Phi + \cos\Phi + V_n\left\{\cos(n\Phi + \Phi_n) - \right.\right.$$

$$\left.\left. - \cos\left(n\Phi + \frac{n\pi}{2} + \Phi_n\right)\right\}\right]$$

H_2, H_3, H_4 may be similarly calculated.

The $+45°$ and $-45°$ fundamental components of this quantized waveform, $V_{f(+45)}$ and $V_{f(-45)}$ respectively may now be conveniently derived as:

$$V_{f(+45)} = \frac{\sqrt{2}}{\pi}(H_1 - H_3)$$

$$V_{f(-45)} = \frac{\sqrt{2}}{\pi}(H_2 - H_4)$$

leading to a phase error of Φ_e degrees.

$$\Phi_e = \frac{180}{\pi}\left[\frac{\pi}{4} - \Phi - \arctan\frac{V_{f(-45)}}{V_{f(+45)}}\right] \text{ degrees}$$

and amplitude error V_e

$$V_e = \frac{100\pi^2}{8}\left[\sqrt{V_{f(+45)}^2 + V_{f(-45)}^2} - \frac{8}{\pi^2}\right] \text{ per cent.}$$

A simple FORTRAN program was used to make the above calculations for each harmonic, Φ and Φ_n being incremented independently over the range 0 to 2π radians.

15 Appendix 4: Low Q Factor Phase Error

Each of the four channels of the filter is active over one quarter cycle, and open circuit or 'floating' the remainder of the time. The voltage on an individual channel capacitor may therefore typically vary in time as shown in Fig. 7(a) where V_N is the mean voltage over the active quadrant and V_{AV} is the effective average over one cycle. Normal high-Q operation ensures that these voltages are very nearly equal and the consequent phase error negligible. As the frequency of operation (hence the Q factor) is reduced, the ratio $V_{AV}:V_N$ will depart
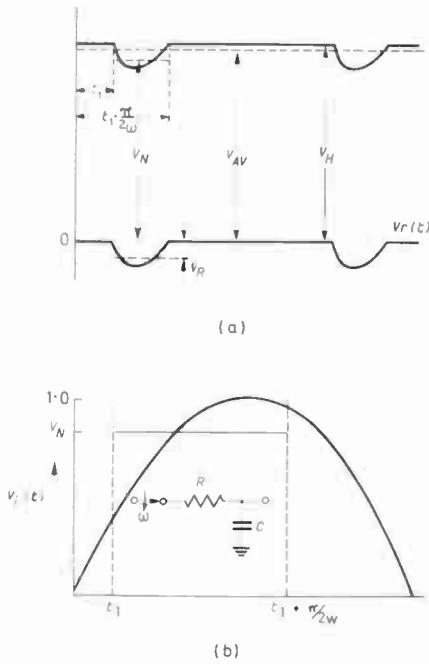
(a)

(b)

Fig. 7. Low Q factor operation.

significantly from unity, and this will be evidenced by an apparent phase error. However, since asynchronous operation of the output commutator in general results in a uniform scanning of all capacitor voltages, an estimate of the mean phase error may be obtained directly from a knowledge of the ratio $V_{AV}:V_N$.

Referring to Fig. 7(b), a single channel may be considered to be active during the time interval $t_1 \rightarrow t_1 + \pi/2\omega$, in which case V_N may be calculated as:

$$V_N = \frac{2}{\pi}(\sin \omega t_1 + \cos \omega t_1)$$

From Fig. 7(a) the instantaneous capacitor voltage may be considered as made up of the 'hold' voltage V_H and a small superimposed ripple component $V_r(t)$. Since the capacitor charging current is approximately $(V_i(t) - V_N)/R$ where V_i is normalized to $\sin \omega t$, the instantaneous ripple voltage is given by the equation

$$V_r(t) = \frac{1}{CR}\int \sin \omega t - V_N)\mathrm{d}t$$

Integrating,

$$V_r(t) = \frac{1}{CR}\left[-\frac{\cos \omega t}{\omega} - V_N t\right] + K \qquad (8)$$

The mean value of this ripple component V_R over the period $t_1 \rightarrow t_1 + \pi/2$ is therefore

$$V_R = \frac{2\omega}{\pi}\int_{t_1}^{t_1+\pi/2} V_r(t)\mathrm{d}t$$

integrating a second time,

$$V_R = \frac{2\omega}{\pi}\left[-\frac{\sin \omega t}{\omega^2 CR} - \frac{V_N t^2}{2CR} + Kt\right]_{t_1}^{t_1+\pi/2} \qquad (9)$$

when $t = t_1$, $V_r(t) = 0$ then, from equation (8),

$$K = \frac{1}{CR}\left[\frac{\cos \omega t_1}{\omega} + V_N t_1\right]$$

Substituting for K in equation (9) and simplifying, gives

$$V_R = \frac{1}{4\omega\pi CR}\left[(4\pi - 8)\cos \omega t_1 + 8 \sin \omega t_1 - \pi^2 . V_N\right]$$

where

$$V_N = 2\frac{(\sin \omega t_1 + \cos \omega t_1)}{\pi}, \quad Q = \frac{\omega NRC}{2} \quad \text{and} \quad N = 4$$

Therefore

$$V_R = \frac{1}{2\pi Q}(2\pi - 8)(\cos \omega t_1 - \sin \omega t_1)$$

The mean error voltage over one complete cycle may be expressed as

$$\Delta V = -\frac{V_R}{V_N} . \frac{3}{4} . 100 = \frac{18.75}{Q} \times$$

$$\times \left[\frac{(2\pi - 8)(\sin \omega t_1 - \cos \omega t_1)}{(\cos \omega t_1 + \sin \omega t_1)}\right] \text{ per cent.}$$

For the particular example of an in-phase signal, t_1 takes values $0, \pi/2\omega, \pi/\omega, 3\pi/2\omega$ for each of the four channels, then

$$\Delta V = +d, -d, -d, +d \text{ respectively,}$$

where

$$d = \frac{32.3}{Q}$$

In this case, ΔV appears as a small quadrature component giving a phase advance $\Delta\Phi$ of

$$\Delta\Phi = \frac{0.323}{Q} \text{ radians}$$

thus, if for example, $Q = 10$

$$\Delta\Phi = 1.84°$$

The theory is accurate only for small phase errors; nevertheless, it does demonstrate how low Q operation degrades performance.

Design and application of active compensation circuits for servo control systems

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Based on part of a paper presented at a meeting of the Southern Section held in Newport, Isle of Wight on 10th November 1972.

SUMMARY

The paper utilizes the phase variable form of the state equation describing a given linear network. The networks that are considered are those most commonly used, namely, phase advance/phase retard and notch networks, to stabilize a servo system. Practical results are given which illustrate the simplicity of tuning the characteristics and which demonstrate that implementation of these filters is a significant improvement in the 'state of the art'.

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List of Principal Symbols

s	Laplace transform frequency variable
ω_0	notch network centre frequency
$\lambda = s/\omega_0$	normalized frequency variable
f_a, f_b	network break frequencies
b	notch network 3 dB bandwidth
d	notch network attenuation at ω_0
J_L	inertia of the load in a servomechanism
J_M	inertia of the servo-motor armature in a servomechanism
N	gear-box ratio

1 Introduction

The design and synthesis of control systems entails the specification of compensation or stabilizing networks. Configurations that are often used are

$$K\left(\frac{s+A}{s+B}\right), \tag{1}$$

$$\frac{s^2+a_1 s+a_2}{s^2+b_1 s+b_2}. \tag{2}$$

The transfer function defining equation (1) represents the form of a phase advance ($A < B$) or a phase lag ($A > B$) network. Basically such networks provide extremely common methods of achieving the desired specification in a control system. Figure 1 shows typical forms of the realization of this circuit.
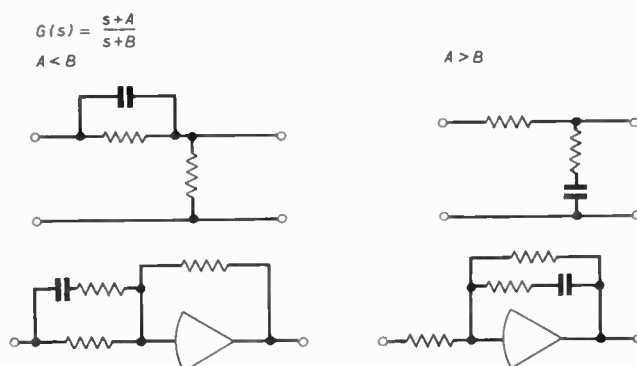
Fig. 1. Simple phase lead and phase lag configuration.

The transfer function defined by equation (2) can be used to describe a network with a frequency response that has a sharply selective characteristic, commonly referred to as a notch network.

In the special case of a narrow frequency selective characteristic of the type in Fig. 2, equation (2) can be expressed[1]

$$\frac{v_0}{v_i} = \frac{s^2+\left(\dfrac{bd}{\sqrt{1-2d^2}}\right)s+\omega_0^2}{s^2+\left(\dfrac{b}{\sqrt{1-2d^2}}\right)s+\omega_0^2}.$$
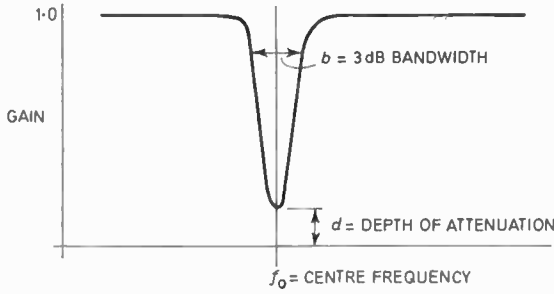
Fig. 2. Definition of notch pa ameters.

It follows that, for a sharply selective network such that $1 \gg d^2$, the notch characteristic is defined by

$$\frac{v_0}{v_i} = \frac{s^2 + bds + \omega_0^2}{s^2 + bs + \omega_0^2}.$$

In circuit theory the Q-factor of a circuit is often used as a design parameter. In terms of the parameters b and ω_0, Q can be written as

$$Q \simeq \left(\frac{b}{\omega_0}\right)^{-1}.$$

Traditionally such networks have been synthesized by bridge and parallel-T network configurations of which there are numerous minor derivatives.[2] A typical circuit is shown in Fig. 3. The problem of all these configurations, particularly where some adjustment is necessary on site, is that changes in the coefficients requires major modification on the components of the network. To change the characteristic of the notch requires a new set of resistors and capacitors, although a limited tuning capability can be introduced by ganging resistors etc., but in general the characteristics are mutually interdependent, consequently only limited tuning is available.

The object therefore of this paper is to present simple active circuits which have independent parameter control and which are ideally suited where on-site tuning is likely.
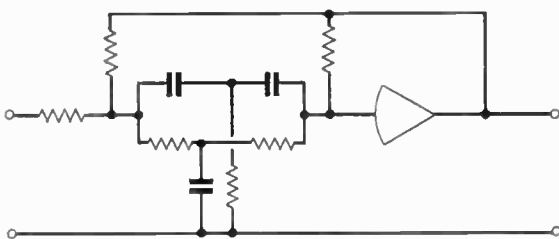
Fig. 3. Typical parallel-T notch network.

2 The State Variable Filter

The transfer function

$$\frac{v_0}{v_i} = G(s),$$

where

$$G(s) = K \frac{(a_n s^n + a_{n-1} s^{n-1} \dots a_0)}{s^n + b_{n-1} s^{n-1} \dots b_0}, \qquad (4)$$

$$= N(s) \cdot \frac{K}{D(s)}, \qquad (5)$$

can be written in terms of the variable ε as

$$\frac{v_0}{\varepsilon} = N(s), \qquad (6)$$

$$\frac{\varepsilon}{v_i} = \frac{K}{D(s)}. \qquad (7)$$

Equation (7) can be expressed in the phase variable form

$$\dot{x} = Ax + Bv_i,$$

$$\varepsilon = Cx,$$

where

$$A = \begin{bmatrix} 0 & 1 & 0 & 0 & \dots & 0 & 0 \\ 0 & 0 & 1 & 0 & \dots & 0 & 0 \\ -b_0 & -b_1 & -b_2 & -b_3 & \dots & & b_{n-1} \end{bmatrix}$$

$$B = \begin{bmatrix} 0 \\ 0 \\ 0 \\ . \\ . \\ . \\ K \end{bmatrix}, \qquad C = \begin{bmatrix} 1 & 0 & 0 & \dots & 0 \end{bmatrix}.$$

This equation can be implemented as shown in Fig. 4. The output, equation (6) is then formed by the combination of the states $x_n, x_{n-1} \dots x_1$ and the derivative of x_n as shown in Fig. 5. In practice the realization of this network is not as difficult as it may appear for high-gain amplifiers are readily available in miniature component size and hence physical realizability is not a particular problem. Further integrator drift can be kept to reasonable proportions provided all the b coefficients are realistically valued.
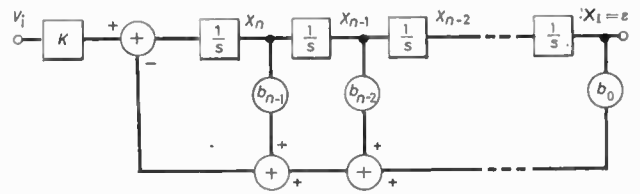

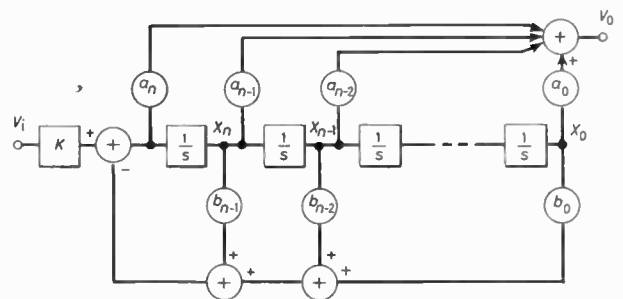Fig. 4. State variable filter for $\varepsilon/V_1 = K/D(s)$.

Fig. 5. State variable filter for $V_0/V_1 = G(s)$.

2.1 *Phase Advance/Phase Lag Networks*

Equation (1) can be implemented as shown in Fig. 6, from which it can be seen that independent control of the break frequencies A, B, can be obtained and range switching introduced as shown. The simplicity of the circuit and the ease of using this instrument particularly for tuning on site commends its application.
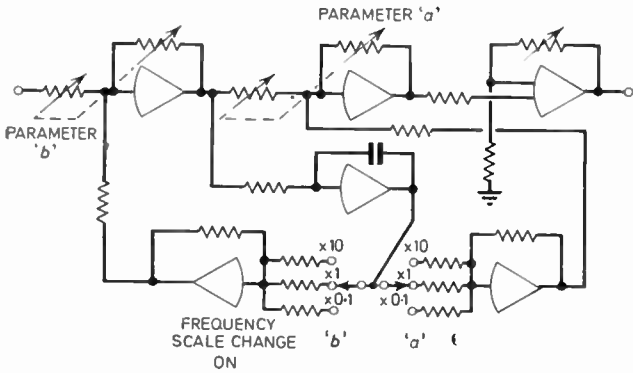

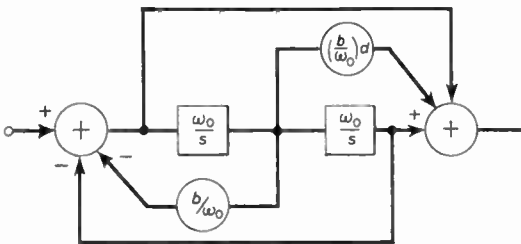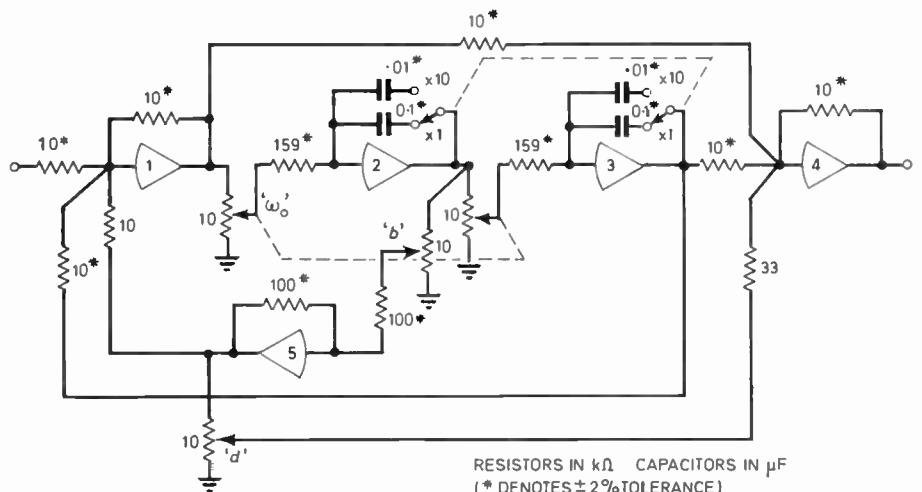Fig. 6. Circuit for phase lead or phase lag filter.

Fig. 7. State variable form of notch filter.

2.2 *Notch Network*

The transfer function defining the notch network can be written in normalized form

$$\frac{\lambda^2 + \lambda\left(\dfrac{b}{\omega_0}\right)d + 1}{\lambda^2 + \lambda\left(\dfrac{b}{\omega_0}\right) + 1},$$

Fig. 8. Circuit for notch filter.

where parameters b, d, ω_0 are defined in Fig. 2 and

$$\lambda = s/\omega_0.$$

Using the state representation the filter can be described as shown in Fig. 7. The parameter (b/ω_0) becomes relative bandwidth $[Q]^{-1}$ and the variable λ becomes the relative frequency variable and hence it can be used as a scale change factor. The state diagram shown in Fig. 7 can be implemented via the circuit shown in Fig. 8 where the parameters ω_0, b/ω_0 and d are independent variables within the limitations that $d^2 \ll 1$.

2.3 *Servo Drive Compensation*

The compensation of resonant and anti-resonant effects which are found in servo drives has been discussed elsewhere.[3] It has been shown that the required compensation transfer function is

$$G(s) = \frac{\lambda^2 + \alpha\lambda + 1 + \beta}{\lambda^2 + \alpha\lambda + 1},$$

where

$$\beta = \frac{J_L}{N^2 J_M},$$

and

α = damping associated with the load inertia.

It follows that the transfer function can be implemented as shown in Fig. 9.
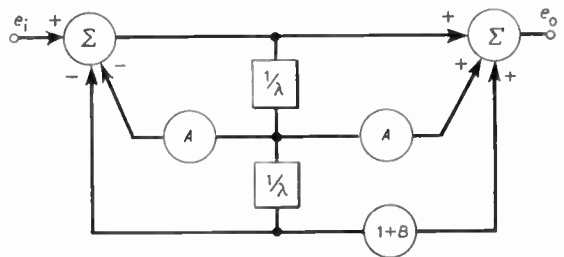

Fig. 9. Servo drive resonance compensation circuit.

3   Practical Results

3.1   *Phase Advance/Phase Retard Filter*

For normal use the ratio of the network break frequencies will lie in the range

$$0 \cdot 1 < \frac{f_a}{f_b} < 10.$$

Figure 10 shows a set of measured results for the filter shown in Fig. 6. Figure 11 shows the effect of changing the network break frequencies f_a, f_b, whilst maintaining the ratio f_a/f_b constant at 10 and $0 \cdot 1$. It is common practice to use networks with break frequencies of these ratios,
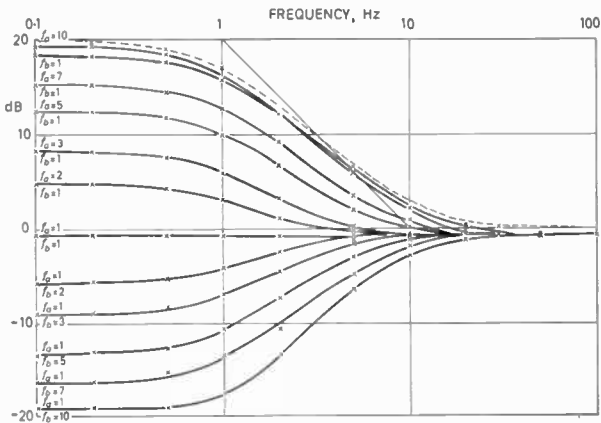

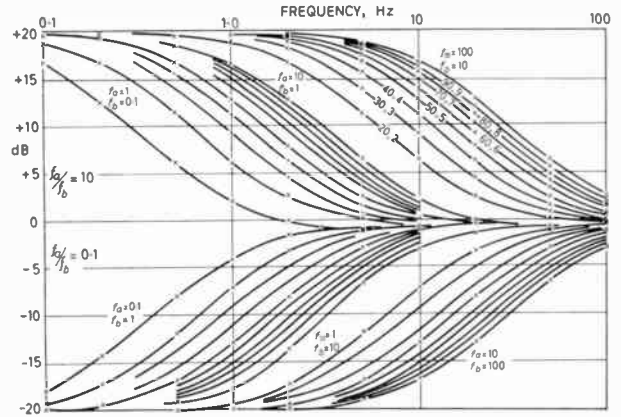**Fig. 11.** Amplitude frequency response for phase advance/phase retard filter for $f_a/f_b = 0 \cdot 1$ and 10.

hence Fig. 11 clearly demonstrates the tuning capability of the filter.

3.2   *Notch Network*

The adaptive capability of the network is shown in Figs 12 and 13 which show a notch set at a nominal 100 Hz and the parameters Q and d varied. Figure 14 shows a series of notches with a Q factor of 10, (or $b/\omega_0 \simeq 0 \cdot 1$) in the range 1 Hz → 100 Hz. Figure 15 compares the repeatability of the notch across the frequency range. The design of notch networks in this frequency range by conventional techniques has not been possible, consequently implication of these networks represents a material improvement in servo filtering techniques.
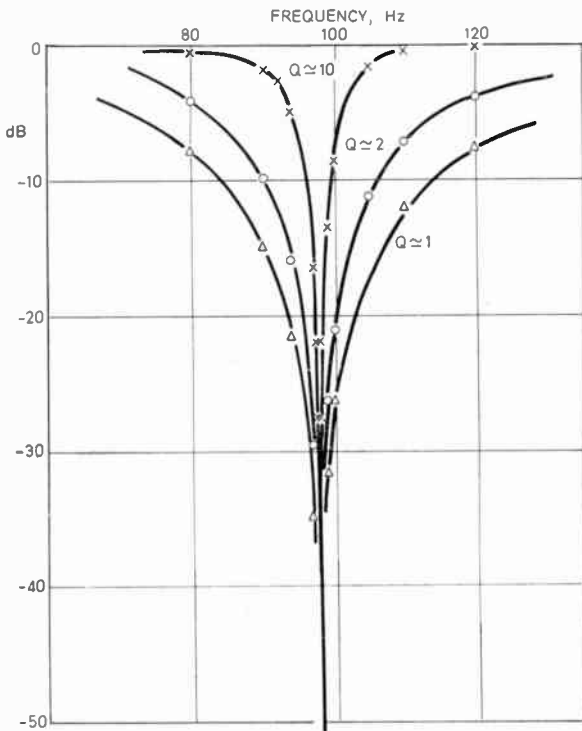

Fig. 10. Amplitude frequency response for phase advance/phase retard filter.

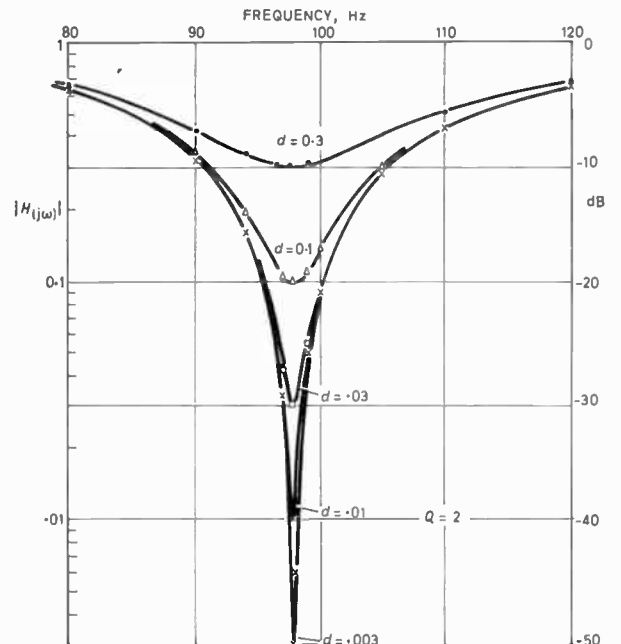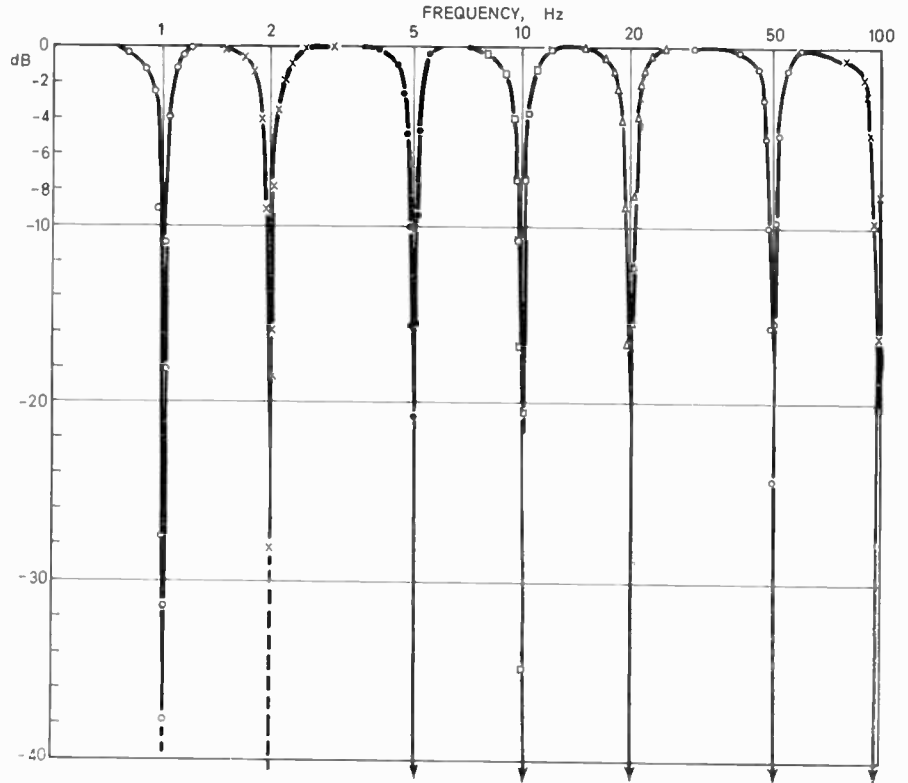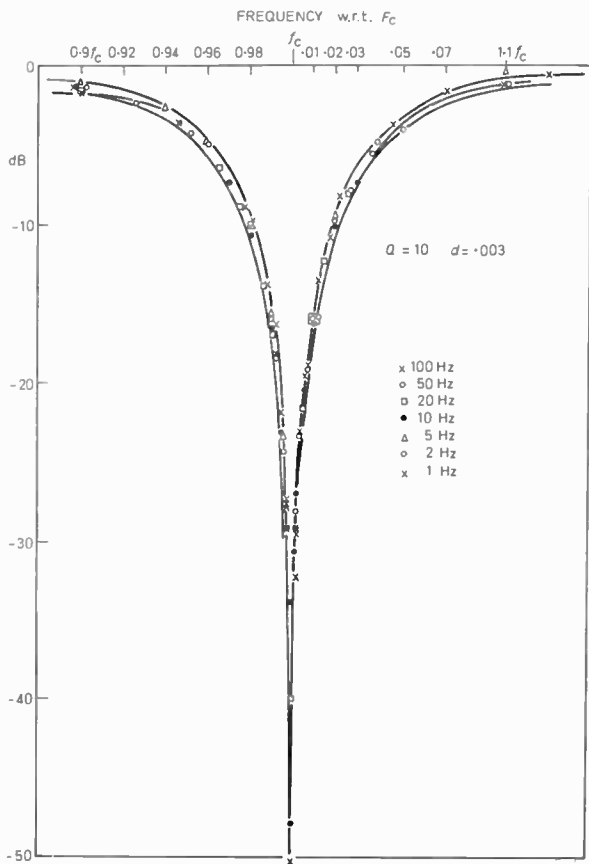**Fig. 12.** Amplitude frequency response for notch filter for fixed f_o and $Q = 1, 2, 10$.

Fig. 13. Amplitude frequency response for notch filter for fixed f_o and Q and $d = 0 \cdot 3, 0 \cdot 1, 0 \cdot 03, 0 \cdot 01, 0 \cdot 003$.

Fig. 14.
Amplitude frequency response of notch filter fixed Q and d, f_0 tuned to 1, 2, 5, 10, 20, 50 and 100 Hz respectively.

Fig. 15. Amplitude against centre frequency, fixed Q and d. The resonance curves have been normalized and superimposed.

4 Conclusions

The network configurations have been discussed which lend themselves to implementation by integrated circuit techniques. The parameters of all the networks can be varied independently and range switching can be simply accommodated. The need for such filters in an experimental and practical system tuning exercise is only too obvious to those engineers who have spent many weary hours 'resistor twiddling' in order to obtain the required frequency characteristics of the compensation filter.

5 References

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GEOS: A European Scientific Satellite

Development and manufacture for ESRO of the first European geostationary scientific satellite—GEOS—will be coordinated by the Electronic and Space Systems group of the British Aircraft Corporation as Prime Contractor on behalf of the STAR Consortium of European electronics and aerospace companies.

The three-years development programme, the contract for which is worth £11 million, will involve 15 companies in 10 European countries. In addition to certain satellite sub-systems, BAC is the coordinating design authority responsible for overall satellite system design and all aspects of the development programme.

GEOS will carry a payload of nine scientific experiments (three of which are integrated to form one unit) into geostationary orbit where from various positions in the magnetosphere the experiments will measure electric and magnetic fields and also particle densities and distributions. In addition the experiment data from GEOS will be correlated with ground based measurements.

As the first scientific geostationary satellite to be undertaken by European Industry, GEOS presents a number of technical challenges, for example:

Deployment of eight booms in orbit ranging in length from 1·5 m to 20 m to isolate sensors from the satellite body.

Measurement by the experiments of much lower field intensities than on previous satellites which impose stringent limitations on the interference generated on board the satellite.

Severe limitations on contamination from the apogee motor (burnt hydrazine) due to the sensitivity of the experiment sensors.

The satellite will be 162 cm in diameter and 110 cm in length and its mass at launch will be 542 kg. The electrical power available for experiments and telemetry will be 90 W derived from solar panels. GEOS is intended to remain in its geostationary orbit (between 15° W and 40° E) for two years from launch date in the summer of 1976.
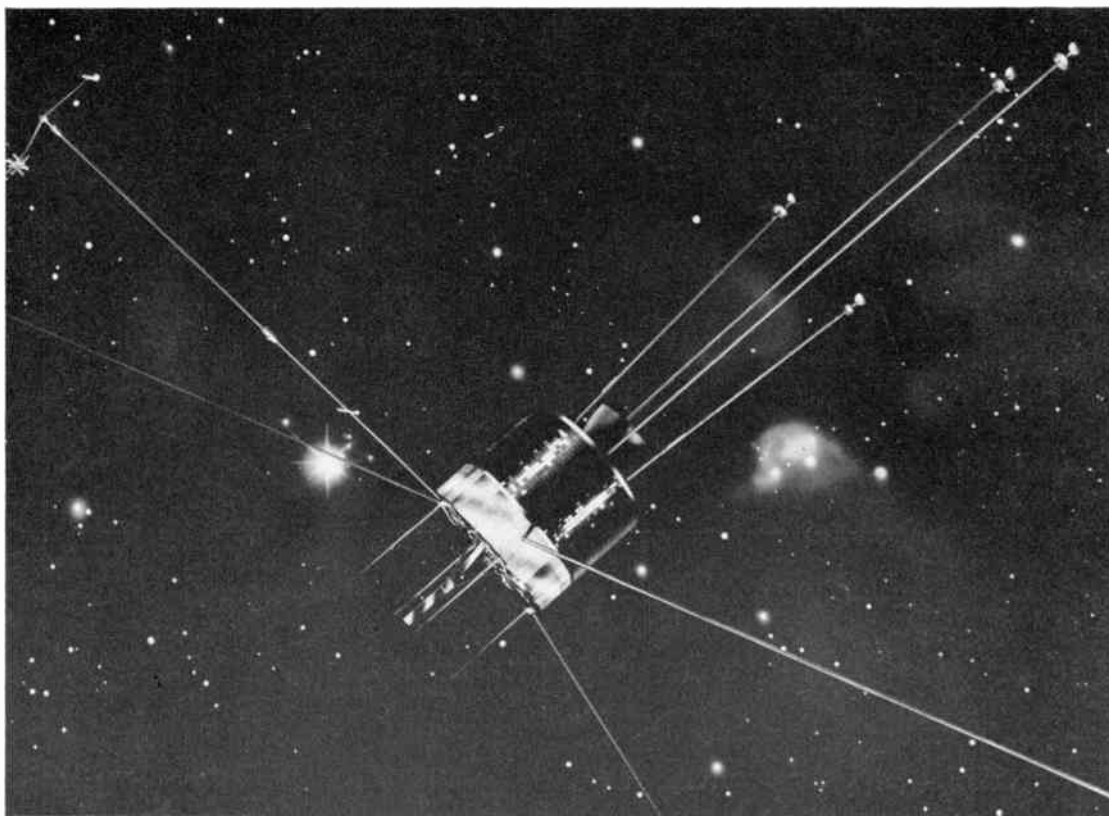
Details of the GEOS experiments are as follows:

RESEARCH INSTITUTE:	EXPERIMENT:
Centre National d'Etudes des Telecommunications, France	Study of thermal plasma through its resonances
France and Danish Space Research Institute, Denmark	Study of low frequency electromagnetic fields
ESTEC Space Science Division, Holland	Measurement of low frequency electric fields
Mullard Space Science Laboratory, London University	Study of the very low energy plasma
Physikalisches Institut der Universitat Bern, Switzerland and Max Planck Institut für Extraterrestrische Physik, Germany	Ion composition, energy spectra and angular distribution of low energy particles and plasma
Kiruna Geophysical Observatory, Sweden	Measurement of the pitch angle distribution of low energy electrons and protons
Max Planck Institut für Aeronomie, Germany	Electron-proton spectrometer for medium energies
Max Planck Institut für Extraterrestrische Physik, Germany	Measurement of d.c. electric field component perpendicular to the magnetic vector
Laboratory for Space Research, CNR, Italy	Measurement of d.c. and u.l.f. magnetic fields

Excitation of dielectric-loaded trough waveguides by an aperture

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and

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SUMMARY

Experimental results and theoretical estimates for the excitation efficiency of the PE_{20} and PM_{11} dielectric-loaded trough waveguide modes by the LSE_{10} and LSM_{11} dielectric-loaded rectangular waveguide modes respectively are presented. The high efficiencies predicted are generally in good agreement with the experimental results.

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List of Symbols

η	excitation efficiency (with various subscripts as defined in the text)
P_T	power transmitted by trough waveguide
P_R	power transmitted by radiation field
S_{11}, S_{12}	scattering coefficients of the excitation system
α	attenuation coefficient of the trough wave mode
l	length of the trough waveguide transmission line
d	thickness of the dielectric in the trough waveguide
k	2π/free-space wavelength

1 Introduction

Dielectric-loaded trough waveguides (herein abbreviated to trough waveguide) have been extensively analysed by Cohn[1,2,3] for use as a low-loss transmission line in the millimetre wave region. Recently, McRitchie and Beal[4] have suggested that trough waveguides can be used to communicate with moving vehicles on railway and other guided ground transport systems.

Because a trough waveguide is not completely enclosed by conductors it is difficult to excite guided modes without loss in the form of radiation. Nevertheless Cohn *et al.*[5,6] have obtained high excitation efficiencies using electric line currents or narrow slots. In this case the maximum efficiency depends fairly critically on both the frequency and the height of the source above the dielectric surface.

An alternative method of excitation is to use an aperture. This method should not be so critical regarding efficiency. The logical method is to form the aperture by truncating a dielectric-loaded rectangular waveguide. The dielectric of this latter waveguide is continuous with the dielectric of the trough waveguide. This report presents theoretical estimates and experimental results for the excitation efficiency of trough waveguide modes excited by such an aperture.

2 General Considerations

Figure 1 shows the general arrangement for exciting the trough waveguide. The length of the dielectric-loaded rectangular waveguide was short compared with the length of the trough waveguide. An air-filled rectangular waveguide was used to feed the dielectric-loaded waveguide. There was a taper in the dielectric to reduce reflexions. Measurements were made in the air-filled waveguide.

Modes in a trough waveguide are either PE-modes or PM-modes. This notation was first introduced by Conlon and Benson[7] and later confirmed by Cohn.[1] In PE-modes the electric field is entirely parallel to the dielectric-air interface. In PM-modes the magnetic field is entirely parallel to the dielectric-air interface. There are two dominant modes. When the trough waveguide is narrow and the dielectric relatively thick the PE_{20} mode is dominant. When the trough waveguide is wide

and the dielectric relatively thin the PM_{11} mode is dominant.

Because of the similarity in mode structure, the dielectric-loaded rectangular waveguide LSE_{10} mode was used to excite the PE_{20} mode and the LSM_{11} mode was used to excite the PM_{11} mode. The dielectric used throughout was p.t.f.e. with a dielectric constant of 2·08.

3 Excitation Efficiency

The excitation efficiency of a trough waveguide is defined in a similar way to the launching efficiency of surface waves.[8] Thus the excitation efficiency

$$\eta = \frac{\text{power in trough waveguide}}{\text{power leaving the source}}$$

$$= \frac{P_T}{P_T + P_R} = \frac{1}{1 + P_R/P_T} \tag{1}$$

This definition relates only to fields to the right of the aperture at A in Fig. 1. If it is assumed that the only loss between the rectangular waveguide and the trough waveguide is due to radiation then[9]

$$\eta = \frac{|S_{12}|^2}{1 - |S_{11}|^2} \tag{2}$$

The scattering coefficients S are readily measured. In any practical device it is desirable to make $|S_{11}|$ zero, but it is not necessary for the application of (2).
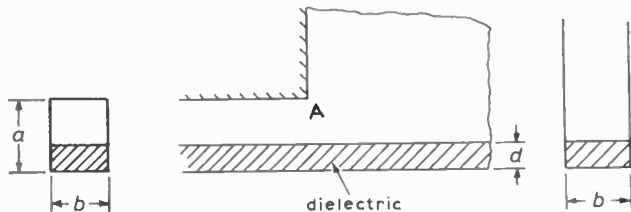

Fig. 1. General arrangement.

4 Theory

The fields excited in the trough waveguide by the aperture were calculated on the assumption that the aperture consisted of a number of electric and magnetic dipoles. The exact distribution of these dipoles is not known so an estimate of this distribution was made using the Field Equivalence Theorems given by Schelkunoff.[10] The induction theorem and the field analogues of Thevenin's and Norton's circuit theorems were all used separately to estimate the dipole distribution. Since no estimate produces the exact answer, these three methods were used to find which of them yielded the best answer when compared with experimental results.

The analytical procedure to find the fields excited by the dipoles is well known. From these field distributions the powers in the trough waveguide and the radiation field were calculated by numerical integration and the excitation efficiency calculated using eqn. (1). The details can be found in Reference 11. The results for the efficiency for various values of kd are shown in Figs. 2 and 3. In the case of the excitation of the PM_{11} mode a very high efficiency is predicted at low values of kd. This

occurs in the range of kd where the PM_{11} mode can propagate but the radiation field (actually a radial waveguide mode[11]) is below cut-off.

5 Experiment

The unit containing the partially filled rectangular waveguide was called a transducer. The various transducers used are shown as insets in Figs. 2 and 3. The first measurements of the scattering coefficients S_{11} and S_{12} were obtained by terminating the trough waveguide by a short circuit and varying the frequency. The efficiencies η_1 obtained using eqn. (2) were substantially below any of the theoretical estimates. An improvement was obtained when the theoretical attenuation of the trough waveguide was used as a correction since the actual efficiency

$$\eta_a = \eta_1 \exp(2\alpha l) \tag{3}$$

The values of η_a from (3) were still below the theoretical estimates, which suggests that the attenuation was higher than estimated. To overcome this error the trough waveguide was terminated by a transducer which was identical to the excitation transducer. The air-filled rectangular waveguide part of this second transducer was then terminated by a variable short circuit. Using this variable short circuit another set of scattering coefficients was obtained, which gave another value for the efficiency η_2. It can be shown[12] that the actual efficiency

$$\eta_a = \eta_2 \exp(\alpha l) \tag{4}$$

Eliminating αl between (3) and (4) yields

$$\eta_a = \eta_2^2/\eta_1 \tag{5}$$

The values of η_a obtained from (5) were in much better agreement with theoretical estimates.

Thus two efficiencies η_1 and η_2 were measured using eqn. (1). The actual efficiency was calculated from eqn. (5).
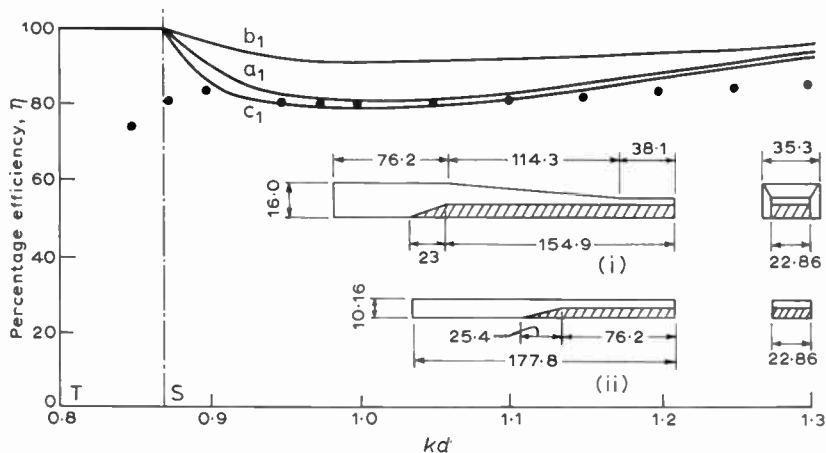
5.1 PE_{20} Mode Excitation

When $b = 10·16$ mm, the PE_{20} mode is dominant in the range $1·51 < kd < 2·40$. Figure 2 (insets) show the transducers used. For the lower values of kd the feed was in WG14 (Fig. 2(i)). For the higher values of kd the feed was in WG16 (Fig. 2(ii)). In these transducers only the LSE_{10} mode could propagate. The measured efficiencies shown in Fig. 2 are in good agreement with those estimated using the field equivalent of Norton's circuit theorem.

The transducer shown in Fig. 2(iii) has a larger aperture. In the range of measurement this transducer can also support the higher order LSE_{20} mode. Assuming that there was no LSE_{20} mode present the efficiency was found to be about 85%. Using a resonance technique[13] it was found that some LSE_{20} mode was present and its power was about 14% of the power in LSE_{10} mode. Now the LSE_{20} does not excite the PE_{20} mode[11] so this 14% must go into radiation. Using this knowledge the previously measured efficiency was corrected. The results are shown in Fig. 2 by the curves.

These results show that if a horn system is used to excite trough waveguides care must be taken not to excite higher-order modes within the horn.

Fig. 3. Graph of the excitation efficiency
of the LSM_{11} mode against kd

●   Experiment

—   Theory, $d/a = 0.625$

a_1 using the induction theorem

b_1 using the field analogue of
     Norton's Theorem

c_1 using the field analogue of
     Thevenin's Theorem

Inset figures: transducers used

(i) $0.85 < kd < 1.05$

(ii) $1.05 < kd < 1.3$

All dimensions in millimetres.

5.2 PM_{11} Mode Excitation

When $b = 22.86$ mm, the PM_{11} mode is dominant in
the range $0.85 < kd < 1.3$. The transducers used are
shown in Fig. 3. For the lower values of kd the feed was
in WG14 (Fig. 3(i)). For the higher values of kd the
feed was WG16 (Fig. 3(ii)). These transducers supported
the LSM_{11} mode only. The experimental results are
shown in Fig. 3. In this case there is not such good
agreement as there is with the theoretical estimates. The
nearest estimate is that obtained using the field analogue
of Thevenin's circuit theorem. At low values of kd the
low measured efficiency is not surprising. In this range,
the guide wavelength is long and the aperture is in a
'supergain' condition. Thus there will be large reactive
fields. These fields induce local wall currents and hence
increase the loss. The experimental results neglect these
wall losses.

6 Conclusions

Theoretical estimates and experimental results have
been presented for the excitation of trough waveguide
PE_{20} and PM_{11} modes by dielectric-loaded rectangular
waveguide LSE_{10} and LSM_{11} modes respectively. For
the excitation of the PE_{20} there was good agreement
between theory and experiment. The best theoretical
estimate was obtained using the field equivalence of
Norton's circuit theorem. This theorem uses the
magnetic field to estimate the dipole distribution in the
aperture. In general LSE modes have one component
of electric field and two components of magnetic field
in the plane of the aperture. The two magnetic fields
have different spatial variations while the electric field
has the same spatial variation as one of the magnetic
fields. Thus Norton's theorem uses the minimum
number of dipoles consistent with all spatial variations.

Theoretical estimates for the excitation of the PM_{11}
mode predict very high efficiencies at both low and high
frequencies with a minimum in between. The experi-
mental results confirm this general trend. The theoretical
estimate based on the field analogue of Thevenin's
circuit theorem yields the best result. This theorem uses
the electric field to estimate the dipole distribution. In
general LSM modes have one component of magnetic
field and two components of electric field. Thus
Thevenin's theorem uses the minimum number of
dipoles consistent with all spacial variations.
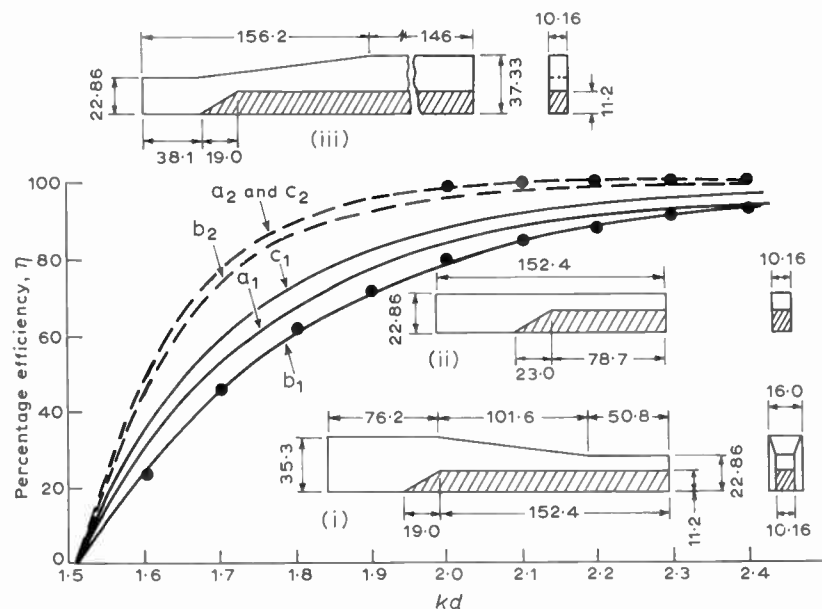
Fig. 2. Graph of the excitation efficiency of the
LSE_{20} mode against kd

●   Experiment

— $d/a = 0.3$ } Theory
--- $d/a = 0.49$

a_1, a_2 using the induction theorem

b_1, b_2 using the field analogue of
       Norton's Theorem

c_1, c_2 using the field analogue of
       Thevenin's Theorem

Inset figures: transducers used

(i) $1.6 < kd < 2.0$ } $d/a = 0.3$
(ii) $2.0 < kd < 2.4$

(iii) $2.0 < kd < 2.4$ $d/a = 0.49$

All dimensions in millimetres.

In the case of the excitation of the PE_{20} mode, higher order modes in the aperture and feed system reduce the efficiency. Thus when very large apertures are used, for example horns, care must be exercised not to have unwanted modes.

7 Acknowledgment

One of the authors (N.B.S.) wishes to thank the University of Salford for a research studentship which made the project possible.

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Dr. J. A. Staniforth graduated with an honours degree in electrical engineering from Sheffield University in 1956. His doctorate was awarded in 1959 by the same University for research on radio surface waves. After a period of about a year at the Admiralty Surface Weapons Establishment, Portsmouth, he joined the staff of the Royal Technical College, Salford in 1960. The College is now the University of Salford and he is at present a lecturer in the Department of Electrical Engineering.

Dr. Staniforth has contributed several papers and articles to learned and scientific journals and he has written a book on Microwave Transmission. His main research activity is wave propagation especially at microwave frequencies.

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For the next year, Dr. Saddler worked as a Research Fellow at the University of Salford, on the propagation of sub-millimetre waves. He returned to Jamaica in September 1969 to take up an appointment as Telecommunications Engineer in the Ministry of Communications and Works, and in October 1972 he joined the Jamaica Telephone Company as a Senior Engineer.

The effect of variations in temperature and in material parameters on the velocity-field characteristic of gallium arsenide

R. J. CLARKE, B.Tech., M.Sc. *

List of principal symbols

D	energy separation between valleys, eV
E	electric field intensity, V/m
E_p	value of E at which $\bar{v}_p$ occurs
E_v	value of E at which $\bar{v}_v$ occurs
E_x	normalizing value of E
k	Boltzmann's constant $= 1\cdot38 \times 10^{-23}$ J K^{-1}
K_1	constant
m_0	free space electron mass, $9\cdot1 \times 10^{-31}$ kg
m_L	lower valley effective electron mass
m_U	upper valley effective electron mass
n_U	upper valley electron concentration, m^{-3}
n_L	lower valley electron concentration, m^{-3}
p	mobility–field exponent
q	electron charge, $1\cdot6 \times 10^{-19}$ C
S	effective density of states ratio
t	relaxation time, seconds
t_{EL}	lower valley energy relaxation time
t_{ML}	carrier momentum relaxation time in lower valley
t_{MU}	carrier momentum relaxation time in upper valley
t_{LU}	lower to upper valley electron transfer time
t_{UL}	upper to lower valley electron transfer time
T	temperature, K
T_L	lower valley electron temperature
T_U	upper valley electron temperature
μ_L	lower valley carrier mobility, m^2/Vs
μ_U	upper valley carrier mobility, m^2/Vs
$\bar{v}$	average carrier velocity, m/s
$\bar{v}_L$	lower valley carrier velocity, m/s
$\bar{v}_p$	peak value of $\bar{v}$
$\bar{v}_v$	valley value of $\bar{v}$

The equations derived in the paper produce consistent values of field and velocity (V/m and m/s respectively). However, due to the widespread use of kV/cm as a unit of field in literature on this subject this unit has been retained in the text and Figures.

SUMMARY

A model for the velocity–field characteristic of gallium arsenide is developed which is based on the assignment of two separate temperature values to the carrier populations of upper and lower conduction-band valleys. The model takes into account two significant factors; (1) the temperature dependence of carrier relaxation time, and (2), the effect of high electric field intensities on the mobility of carriers in the upper conduction-band valley. The resulting characteristic agrees well with those both predicted and observed by other workers whilst the model allows direct insertion of material parameters into the relevant equations.
As the ambient temperature increases, a significant reduction in the ratio of peak-to-valley carrier velocity is shown to occur, together with a small increase in the value of the threshold field.

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1 Introduction

In order to predict the influence of temperature and material parameter variations on the velocity–field characteristic of gallium arsenide it is necessary to employ a model which, as far as possible, incorporates important parameters in a form in which they may easily be varied, in particular according to their temperature dependence. At the same time account must be taken, as far as possible, of all important physical processes occurring within the material. The present paper attempts to do this by suitable modification of a basic model of the velocity–field characteristic.

Initial examination of the well-known model of Mc-Cumber and Chynoweth[1] showed that the form of the velocity–field characteristics did not agree with expectations due to the facts that, firstly, the total carrier distribution was assumed to have a single value of electron temperature, and second, the effect of high field intensities on carrier mobility was not considered. In order to overcome the first of these objections the approach described by Carroll[2] was adopted, in which upper and lower valley populations are each assigned a separate electron temperature. This is necessary since

consideration of the carrier transfer process between valleys and the related values of mobility shows that the lower valley electrons will be much 'hotter' (have much greater energy) than those in the upper valley. There are two main reasons for this. First, the much smaller value of mobility applicable to the upper valley means that carriers in that valley gain energy much less readily than do lower valley carriers. Second, the electron population of the upper valley is derived from that of the lower valley through a collision process involving significant energy loss. The model thus incorporates two different valley electron temperatures, that of the lower valley which may rise to very high values (around 6000 K) as the applied field becomes very large and that of the upper valley which, as the carriers have much lower energies, is taken as the temperature of the crystal lattice.

The resulting characteristic still has a definitive positive slope at high field intensities, however, since the reduction in carrier mobility associated with large values of field has not been allowed for. This objection is overcome by allowing the upper valley carrier mobility to become a function of the applied field. The resulting characteristic then follows that observed experimentally[3,4] at moderate values of field (up to approximately 20 kV/cm) and thereafter the average velocity remains approximately constant as the field is increased. That this behaviour is correct is supported by the large values of electric field shown to exist within the travelling domain by Gunn[5] which require the absence of a pronounced positive slope of the velocity–field characteristic at high field values.

In addition to the application of the model to determination of the effect of temperature on the velocity–field characteristic it is used to examine the effect on the characteristic of variations in various material parameters. This makes possible for example, the detection of those parameters to which the velocity–field characteristic is particularly sensitive.

2 The Basic Model

The relevant defining equations for electric field intensity and average carrier velocity are derived in Appendix 1. Appropriate values for the parameters are specified in Appendix 2 and may be directly substituted in equations (1) and (4). A range of values of T_L is then selected starting just above ambient temperature and extending to some 6000 K and the corresponding values of N and E are then calculated. From these values of N and E the average carrier velocity $\bar{v}$ may be found from equation (9).

3 Modifications to the Basic Model

To make the model a useful analytical one it is now necessary to examine both its behaviour at high values of electric field and its response to changes in ambient temperature.

3.1 High Field Behaviour

At high values of field the velocity–field characteristic will have a positive slope given by $\bar{v}/E = \mu_U$, occupation
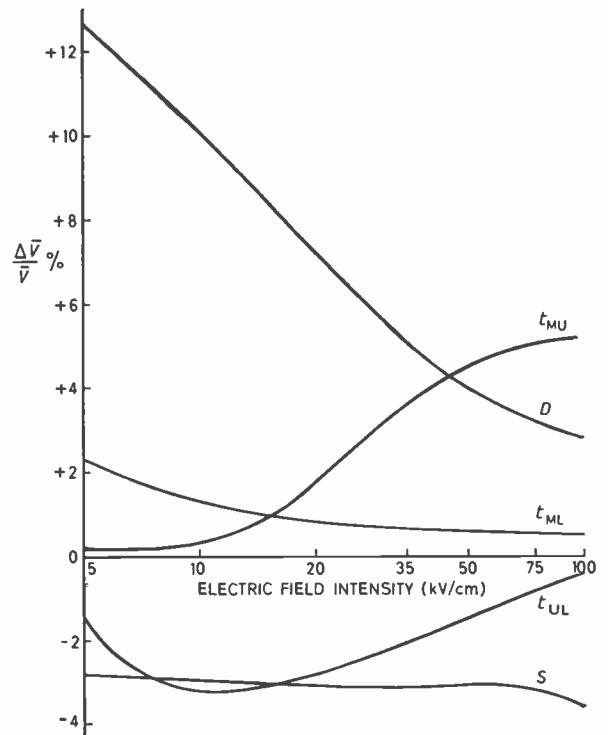

Fig. 1. Percentage change in carrier velocity in response to a $+10\%$ change in the parameter specified (ambient temperature $= 400$ K).

of the upper valley in preference to the lower being essentially complete. This behaviour is not in accordance with expectations,[5] however, and the model must be modified to take account of the decreasing carrier mobility which occurs at high electric fields. Conwell and Vassell[6] find that the field dependence is such as to produce a reduction in upper valley mobility from $0.016 \, \text{m}^2/\text{Vs}$ at low fields to $0.01 \, \text{m}^2/\text{Vs}$ at fields of about $50 \, \text{kV/cm}$. Accordingly their data have been fitted with a relationship of the form

$$\mu_U = \frac{\mu_{U(\text{low field})}}{1 + (E/E_x)^p}$$

where, in this case, $E_x = 70 \, \text{kV/cm}$ and $p = 1.6$. This relationship is applied to the relaxation time t_{MU} which defines the upper valley carrier mobility.

3.2 Temperature Dependence

To produce an adequate temperature dependent model the variations in material parameters with ambient temperature must be specified. As these are not all known with great accuracy an initial investigation is made to ascertain those parameters which, when changed, have most effect on the velocity–field relation. Figure 1 shows the percentage change in velocity in response to a 10% increase in individual specified material parameters and is commented on below (ambient temperature = 400 K).

(i) The energy gap between main and satellite valleys is significant at low field values, its influence falling sharply with increase in field. Its temperature dependence is relatively small however, according to James[7] who estimates it at $-1.1 \times 10^{-4} \, \text{eV K}^{-1}$ and although it has

been included in the calculations it may be neglected without great error.

(ii) A change in the density of states ratio S has an almost constant effect for values of field up to 100 kV/cm. Its temperature dependence is likely to be small as it is a function of the ratio of effective electron masses, whose temperature dependences may be expected to be similar, at least over a limited range of ambient temperature.

(iii) The effect of a change in intervalley relaxation time t_{UL} is greatest at fields in the region of 10 kV/cm, falling to a negligible value at high fields. At present its temperature dependence remains unknown.

(iv) Relaxation times t_{ML}, t_{MU} control lower and upper valley carrier mobilities respectively in accordance with equations (7) and (8). Their effect upon the velocity–field characteristic reflects the changing carrier populations in upper and lower valleys. At low fields the lower valley contains the electron population almost exclusively whilst at higher fields the rapid filling of the upper valley brings the effect of changes in upper valley relaxation time into prominence. The temperature dependence of these parameters is derived from theoretical and experimental work reported by Ehrenreich.[8] Over the range 300–500 K Ehrenreich finds that the variation of mobility with temperature follows the relationship:

$$\mu \propto T^{-K_1}$$

where K_1 is a constant lying in the range 0·8–1·0. Relaxation times t_{ML}, t_{MU} are therefore allowed to vary with temperature in a similar fashion, the value at 300 K being taken as a reference point, so that,

$$\frac{t}{_{(300)}} = \left(\frac{300}{T}\right)^{K_1}$$

where $t_{(300)}$ is the value of the relaxation time at 300 K and K_1 may be selected to fit a known mobility–temperature dependence. In the present analysis, in view of the range of values resulting from Ehrenreich's work, K_1 is set at unity.

4   Results

When the modifications detailed in the previous Section are included in the basic velocity and field equations of Section 2 the results are as shown in Fig. 2,
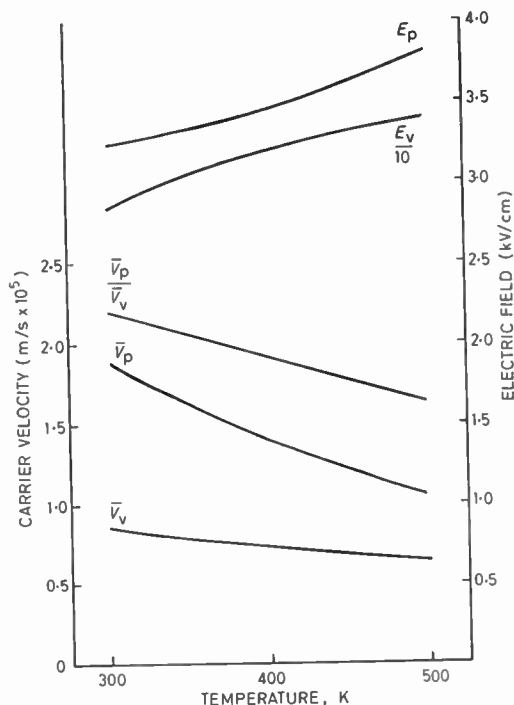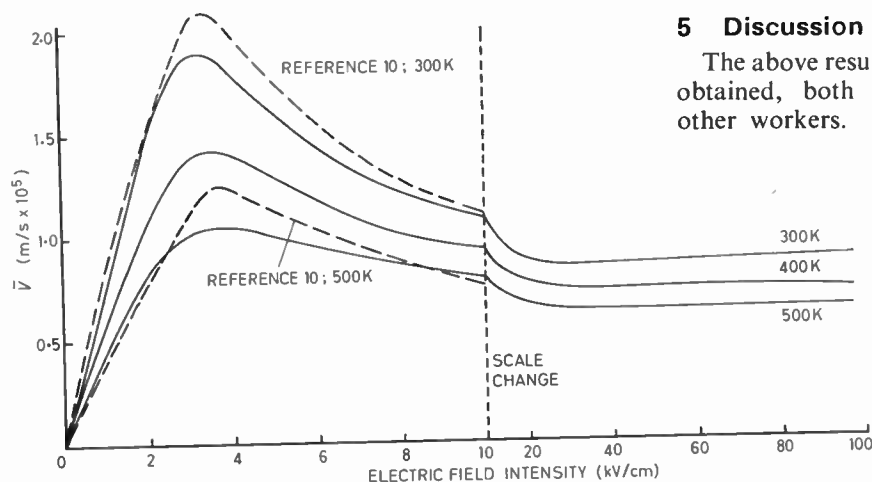
Fig. 3. Temperature variation of the parameters of Table 1.

material parameters being as detailed in Appendix 2. Curves of average carrier velocity against electric field intensity have been plotted for temperatures of 300, 400 and 500 K, together with the results of Ruch and Fawcett.[10] The main points of interest deriving therefrom are included in Table 1 and are displayed in Fig. 3.

Table 1

Temperature °K	$\bar{v}_p$, m/s	$\bar{v}_v$, m/s	E_p, kV/cm	E_v, kV/cm	$\bar{v}_p/\bar{v}_v$
300	$1·90 \times 10^5$	$0·86 \times 10^5$	3·23	28·5	2·21
400	$1·41 \times 10^5$	$0·74 \times 10^5$	3·45	32·0	1·91
500	$1·05 \times 10^5$	$0·64 \times 10^5$	3·82	34·0	1·64

5   Discussion

The above results may usefully be compared with those obtained, both theoretically and experimentally, by other workers. Of the four sets of results shown in

Fig. 2. Velocity–field characteristic for $T = 300$, 400, 500 K with results from Reference 10 shown for comparison.

Fig. 4. Comparison of velocity–field characteristics.

(a) REFERENCE 3
(b) REFERENCE 4
(c) REFERENCE 9
(d) AUTHOR'S VALUES

Table 2 and Figs. 2 and 4 those of Braslau and Hauge[3] and Ruch and Kino[4] are experimental, those of Fawcett et al.[9] and Ruch and Fawce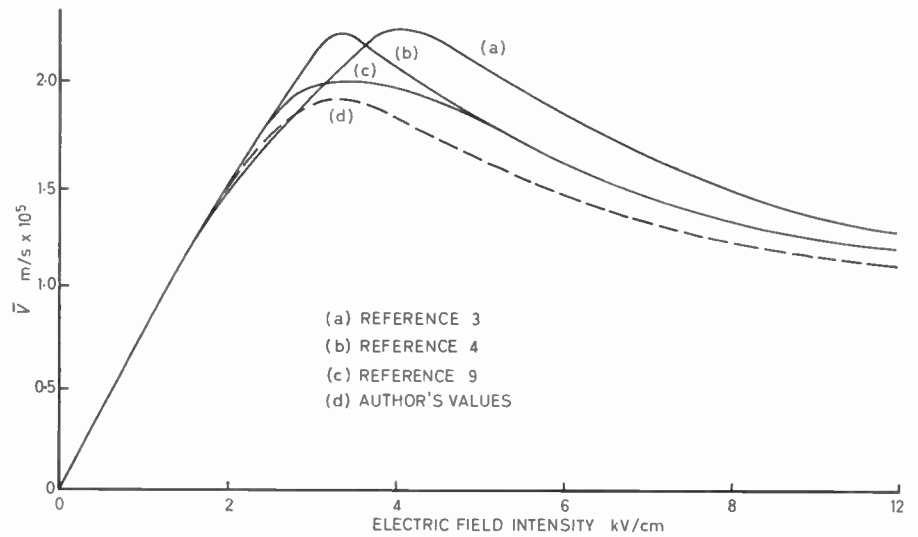tt[10] are theoretical, the latter being of most interest as they include variation of the velocity–field characteristic with material temperature.

Table 2

Refer-ence	$\bar{v}_p$, m/s	$\bar{v}_v$, m/s	E_p, kV/cm	E_v, kV/cm	$\bar{v}_p/\bar{v}_v$
3	2.2×10^5	1.2×10^5	4.0	> 10	1.8
4	2.2×10^5	1.1×10^5	3.3	> 14	2.0
9	2.0×10^5	0.9×10^5	3.4	20–25	2.2
10 (300 K)	2.1×10^5	1.0×10^5	3.4	> 12.5	2.1
10 (500 K)	1.2×10^5	0.7×10^5	3.7	> 12.5	1.7

Braslau and Hauge[3] use a microwave technique and find a relatively high value of peak field, 4.0 kV/cm, all other authors obtaining values (at room temperature) of less than 3.5 kV/cm. It should be noted that values of $\bar{v}_v$ and thus of $\bar{v}_p/\bar{v}_v$ in Table 2 are only approximate, there being a large degree of uncertainty as to the exact location of the minimum. This is reflected in the varying values quoted for E_v, which in several cases lie outside the range of field values employed by the various authors. Fawcett

et al.[9] quote a value for E_v of 20–25 kV/cm which is in good agreement with that of 28.5 kV/cm produced by the present work as indeed are all of the other values determined by them.

The work of Ruch and Fawcett[10] affords a comparison of results obtained as the sample temperature is varied (Fig. 2). Here, especially, an exact comparison will depend on relevant material properties, in particular the variation of carrier mobility with temperature. Nevertheless, both analyses predict a significant decrease in the peak to valley carrier velocity ratio and also a slight increase in threshold field, as the temperature is increased. It is felt that, in view of the inevitable differences in analytical approach and in material parameters employed, overall agreement between all the results quoted is good.

The model may also be used to generate other information about the transferred electron effect in gallium arsenide, for instance, that depicted in Fig. 5 where the variation in lower valley carrier population with electric field intensity is shown for ambient temperatures of 300 and 500 K. Two main features are noticeable. First, the expected rapid electron transfer from lower to upper valley near the threshold field, and second, the assistance given to the transfer process by an increase in ambient temperature at low values of electric field and consequently electron temperature, ambient temperature variations having little effect on the population ratio at high fields.

6 Conclusions

A model of the velocity field characteristic of gallium arsenide has been developed which is in much closer agreement with the observed and expected characteristic than the thermal equilibrium model, whilst still allowing direct employment of the relevant material parameters. It accounts for the effect of high values of electric field on carrier mobility and also of parameter temperature dependence and predicts both a decrease in peak to valley carrier velocity ratio and an increase in threshold
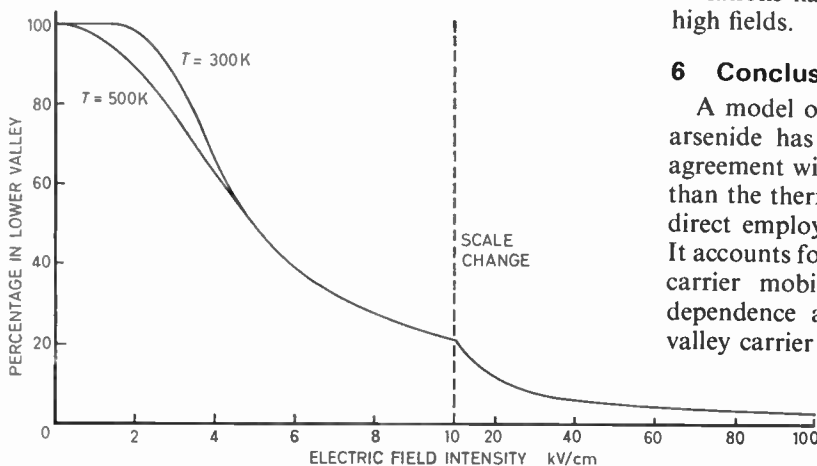

Fig. 5. Field dependence of lower valley carrier population.

field as ambient temperature is increased. The power dissipation in a c.w. transferred electron oscillator results in a material temperature of about 400 K. It is obvious, therefore, that employment of room temperature values of material parameters in oscillator design will lead, in particular, to over-optimistic values of efficiency, which is strongly dependent on the peak to valley velocity ratio.

In addition to the overall effect of temperature on the velocity–field characteristic, is has been shown that, at low fields,the characteristic is very sensitive to changes in the energy gap between upper and lower conduction band valleys, whilst at the highest values of field considered, the single most important factor in determining the form of the characteristic is the upper valley momentum relaxation time (directly related to upper valley mobility).

7 References

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8 Appendix 1:

The equations for field and average carrier velocity

Writing the ratio of electron concentration in upper and lower conduction band valleys as

$$\frac{n_U}{n_L} = S \exp(-D/kT_L) = N \tag{1}$$

The lower valley momentum balance equation[2] leads to an expression for the carrier velocity in the lower valley.

$$v_L = \frac{qE}{m_L[(1/t_{ML})+(N/t_{UL})]} \tag{2}$$

Combining this with the energy balance equation

$$qn_L v_L E = \tfrac{3}{2}k(T_L - T_U)\left(\frac{n_L}{t_{EL}} + \frac{n_U}{t_{UL}}\right) \tag{3}$$

yields an expression for the electric field–lower valley electron temperature relationship (after calculating the

numerical constant and writing the relaxation times for convenience in multiples of 10^{-14} seconds),

$$E^2 = 58 \cdot 8 \times 10^{10}(T_L - T_U)\left(\frac{1}{t_{EL}} + \frac{N}{t_{UL}}\right)\left(\frac{1}{t_{ML}} + \frac{N}{t_{UL}}\right) \tag{4}$$

where an effective electron mass $m_L = 0 \cdot 08\, m_0$ has been assumed.

The average carrier velocity–field relationship

$$\bar{v} = E\left(\frac{\mu_L n_L + \mu_U n_U}{n_L + n_U}\right) \tag{5}$$

is written as

$$\bar{v} = \frac{E}{1+N}\left(\mu_L + N\mu_U\right) \tag{6}$$

where

$$\mu_L = \frac{q}{m_L}\frac{1}{\left(\dfrac{1}{t_{ML}} + \dfrac{N}{t_{UL}}\right)} \tag{7}$$

and

$$\mu_U = \frac{q}{m_U}\frac{1}{\left(\dfrac{1}{t_{MU}} + \dfrac{1}{Nt_{LU}}\right)}. \tag{8}$$

Thus

$$\bar{v} = \frac{qE}{1+N}\left[\frac{t_{ML}}{m_L\left(1 + N\dfrac{t_{ML}}{t_{UL}}\right)} + \frac{Nt_{MU}}{m_U\left(1 + \dfrac{t_{MU}}{Nt_{LU}}\right)}\right]$$

or

$$\bar{v} = \frac{1 \cdot 76 \times 10^{-3}E}{1+N}\left[\frac{12 \cdot 5\, t_{ML}}{\left(1 + N\dfrac{t_{ML}}{t_{UL}}\right)} + \frac{2 \cdot 7\, Nt_{MU}}{\left(1 + \dfrac{t_{MU}}{t_{UL}}\right)}\right] \tag{9}$$

where relaxation times are again specified in multiples of 10^{-14} s, $m_U = 0 \cdot 37\, m_0$ and $Nt_{LU} = t_{UL}$ under the assumption that, in equilibrium, upward and downward electron transfer rates between valleys are equal.

9 Appendix 2:

Values of parameters used in the calculations

S	60
D	$0 \cdot 35$ eV
m_L	$0 \cdot 08\, m_0$
m_U	$0 \cdot 37\, m_0$
T_U	ambient temperature, K
t_{EL}	80×10^{-14} s
t_{ML}	34×10^{-14} s
t_{MU}	$3 \cdot 1 \times 10^{-14}$ s
t_{UL}	200×10^{-14} s.

The corresponding values of low field lower and upper valley carrier mobilities at 300 K are $0 \cdot 75$ and $0 \cdot 015$ m^2/Vs respectively.

IERE News and Commentary

Survey of Professional Engineers

From the President of the Institution

A further Survey of Professional Engineers is being undertaken by CEI this year following up those carried out in 1966, 1968 and 1971. Questionnaires are being distributed in June and July through the Journals of Constituent Members and are inserted in copies of this issue of *The Radio and Electronic Engineer* sent to Corporate Members and Graduates in the United Kingdom.

Completed questionnaires should be returned as early as possible to CEI as it is hoped to complete the analysis of results and publish the Report by the end of the year. In addition to the information on the profession as a whole and data on the various disciplines published in the Report, each Institution receives detailed statistics relating to their own membership. In order that a comprehensive sample may be obtained with the engineering disciplines fairly represented, I would particularly ask for your co-operation in filling in the form and putting it in the post.

A. A. DYSON

Election of the 1973–74 Council of the Institution

In accordance with the Bye-Laws the following vacancies for election to the Council for 1973–74 will arise:

4 Vice-Presidents, 5 Fellows, 3 Members, 1 Associate,

1 Associate Member.

Council's nominations to fill these vacancies are enclosed with this issue.

At its meeting on April 18th, the Council unanimously adopted the recommendation of the Executive Committee that Dr. Ieuan Maddock, C.B., O.B.E., F.R.S. should be nominated for election as President of the Institution for 1973–74.

New Council has its First Meeting

The Council for Environmental Science and Engineering (CESE) sponsored jointly by CEI (the Council of Engineering Institutions) and CSTI (Council of Scientific and Technological Institutes) has had its formation meeting under its Chairman, Sir Kingsley Dunham, S.D., F.R.S., C.Eng., Director of the Institute of Geological Sciences.

The Council, which is composed of senior engineers and scientists representing the 20 constituent institutions of CEI and CSTI, will confine its activities to engineering and scientific aspects of the environment and the problems related to preservation of both natural and man-made resources. The Council believes that everything has its price but that it is possible to achieve a balance between conservation and natural and technological development. Although it is the public which ultimately makes the decisions, it is the responsibility of the engineer and the scientist not only to put the facts to those who take the decisions so as to reconcile amenities and economic consequences but also to participate in decision making itself. Obviously the Council will be much involved in the study of those engineering and scientific environmental problems that could be hazardous to the public.

The formal Terms of Reference of CESE are:

To anticipate and evaluate environmental problems and to represent, coordinate and promote the activities of engineers and scientists in resolving these problems for the benefit of mankind.

To seek and to promote the adoption of realistic solutions to environmental problems, and in particular where technological works are concerned to minimize any adverse consequences.

To encourage and coordinate initiatives by member bodies in the arranging of conferences and in undertaking the study of problems in environmental science and engineering.

The IERE representative on CESE is Professor H. M. Barlow, F.R.S., C.Eng. (Honorary Fellow).

Transfers and Elections to Corporate Membership and 'C.Eng.' Registration

Engineers who do not possess a recognized degree, but who met the academic requirements of the Institution under the IERE Graduateship Examination regulations by the end of 1970, are required by CEI Bye-Laws to register as Chartered Engineers on or before 31st December 1973, if this status is to be achieved without a further academic test. Those concerned will normally be in Graduate membership of the Institution and hold such qualifications as, for example, HNC or HND and endorsements.

It is advisable that such members should apply *now* for a proposal form (Form P) and submit this to the Institution without delay when it has been completed. Those who consider they will not meet the professional engineering experience requirements for some months, but will certainly do so before the end of 1973, should not wait until that time before preparing their applications for election or transfer.

A further announcement will be made concerning the final date for receipt of applications.

Leeds Electronics Exhibition

Mr. A. A. Dyson, O.B.E., President of the Institution of Electronic and Radio Engineers, will open the Tenth Leeds Electronics Exhibition on 3rd July.

The Exhibition takes place in the Department of Electrical and Electronic Engineering at Leeds University from 3rd to 5th July and is open from 10.00 a.m.–6.00 p.m. each day.

The Yorkshire Section of the IERE has arranged programmes of lectures on each of the three days as follows:

Tuesday, 3rd July

2.00 p.m.   'P.c.m. as applied to broadcasting.'
               C. J. DALTON.

3.30 p.m.   'P.c.m. and the telephone network.'
               J. R. CANNON.

Wednesday, 4th July

10.30 a.m.   'Controlled power supplies using high voltage switching transistors.'
               L. E. JANSSON.

2.00 p.m.   'Electronics for the control of small a.c. motors.'
               P. BOWLER.

3.30 p.m. 'Harmonic distortion on supply systems.'
          B. OGDEN

Thursday, 5th July
10.30 a.m. 'European communication satellites.'
          A. DICKINSON.

2.00 p.m. 'Optical fibre communication systems.'
          F. F. ROBERTS.

3.30 p.m. 'Surface acoustic waves in radio communications.'
          R. A. WALDRON.

Apply for tickets to the Exhibition Director, Department of Electrical and Electronic Engineering, University of Leeds, Leeds LS2 9JT.

IEE Conference Publications

Members of the IERE may purchase from the Institution of Electrical Engineers at special reduced rates the Publications for those Conferences of which the IERE was a co-sponsor.

The following IEE Conference Publications have recently become available:

No. 96: Computer Aided Control System Design. Special members' rate: £5.40 (Public price £8.30)

No. 97: Software Engineering for Telecommunication Switching Systems. Special members' rate: £6.70 (Public price £10.20)

No. 98: Propagation of Radio Waves at Frequencies above 10 GHz. Special members' rate: £5.70 (Public price £8.70)

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

Design Council's 1974 Engineering Awards

The Design Council is now inviting applications for its 1974 Engineering Products and Components Awards. The closing date for companies to submit applications is 31st July 1973.

Initiated in 1967, the Design Council Engineering Awards are open to all engineering companies and embrace every field of engineering design from electronics to excavators. In 1973 ten award-winners were chosen from nearly 200 entries, covering such products as a hydrostatic muck-shifter, a solid state relay and an ultrasonic diagnostic scanner. (Electronic products will be described in the next Journal.)

In 1974, for the first time, the Awards will be divided into two sections—Engineering Products and Engineering Components—each with its own panel of judges. In both cases the judges will be looking for an amalgam of engineering excellence, sound ergonomics and value for money. Although products entered for awards should normally have been in use with customers for a year, the judges may consider products of proven performance which have been in use for a slightly shorter period.

Companies wishing to enter for the Design Council's 1974 Awards or requiring more information should contact Mr. Tony Key, Engineering Awards Office, Design Council, 28 Haymarket, London SW1Y 4SU. (Telephone: 01-839 8000, extension 43.)

Correction

The following amendment should be made to the paper 'Impulsive Noise Reduction in Radio Receivers' which was published in the May 1973 issue of *The Radio and Electronic Engineer*:

The block diagrams of Figs. 1 and 6 should be interchanged.

Engineering Unemployment in Canada Increases

Notwithstanding a recent increase in employment opportunities for engineers in Canada, the number of engineers unemployed or temporarily employed in unsuitable other occupations and actively seeking suitable employment has increased by approximately 50% since early 1972.

In a follow-up to the survey of employment carried out in January 1972 by the Canadian Council of Professional Engineers, as part of the annual membership salary survey, the members of the provincial associations were requested to indicate their employment status on the salary survey questionnaire.

On the basis of a 36% return, it was established with a reasonable degree of accuracy that as at 1st December 1972 some 1400 Canadian engineers were either unemployed or temporarily employed in other occupations while actively seeking suitable employment. This is a 49% increase over 1971 and represents 2% of the active engineering labour force in Canada. In addition, some 1950 employed engineers are seriously seeking new employment because their present position appears to be in danger while another 2550 are seriously seeking new employment for other reasons, bringing the total number of Canadian engineers available for employment to well over 5500.

During the last 12 months it is estimated that 4200 new graduates at the bachelor level were produced in Canadian universities and 1600 foreign engineers were admitted to Canada. Taking into consideration deaths, retirements and new graduates entering graduate schools, it can be estimated that the net increase of engineers in the labour market during the year was at least 4000. (The total number of registered engineers and recorded engineers-in-training has increased by approximately 3400 during the same period.)

The rate of unemployment varies with the branch of engineering and the geographical location, the highest rate occurring in British Columbia at 3·4% followed by Ontario and Quebec at 1·9% down to zero unemployment in Prince Edward Island.

New Educational Films

Five new 16 mm educational films in colour have recently been announced by Mullard Ltd. The first two form a series on magnetism, the other three deal with television.

These films are notable in the Mullard catalogue in that they have been made in collaboration with the Company's overseas associates. Either films of overseas origin have been 'anglicized' both in regard to captions etc. and educational approach, or they are to be adapted for use in Europe or elsewhere. In general the level of approach is appropriate to a sixth form in a grammar school or the early years of, say, an ONC course.

Details of the films are as follows (running times are from 16 to 22 minutes):

Magnetism Part I. The Lorentz Force and Hall Effect

Magnetism Part II. The Magnetic Field of a Current

The Television Picture Tube

The Television Carrier Wave

The Television Camera Tube

All five films are available on loan from the Mullard Film Library, 269 Kingston Road, Merton Park, London, S.W.19. Alternatively, they may be purchased from Gateway Educational Films Ltd., St. Lawrence House, 29–31 Broad Street, Bristol BS1 2HF. (0272-25351).

Forthcoming Conferences

Semiconductor Memory Device and Circuits Summer Course

The Microelectronics Laboratory of the Catholic University of Louvain, Belgium, is organizing a summer course on Semiconductor Memory Devices and Circuits, to be given at Louvain-la-Neuve during the week of August 27th–31st, 1973. The purpose of this course is to offer a tutorial survey of modern semiconductor memories, and it will cover the operation of semiconductor memory devices, characteristics of memory cells, and the operation of memory integrated circuits. The official language of the course is English and lecturers are from Europe and the United States.

The advance registration fee is $75. Further information on the programme and accommodation may be obtained by writing to: Prof. E. Van de Wiele, Université Catholique de Louvain, Batiment Maxwell, 1348 Louvain-la-Neuve, Belgium.

R.F. Measurements Vacation School

A comprehensive series of lectures delivered by men well known in the measurements field, tutorials, discussion groups and demonstrations all figure in the second residential Vacation School on 'R.F. Electrical Measurement Practice'. Organized by the Science, Education and Management Division of the Institution of Electrical Engineers and sponsored by the British Calibration Service, the Institution of Electronic and Radio Engineers and the Institute of Electrical and Electronics Engineers (UKRI Section), the School will be held at the University of Surrey, Guildford from 9th to 21st September 1973.

Both the theoretical background and the practical techniques of measurement using modern instruments will be covered at the School. Tutorials and discussion groups will be related to the lectures which will include such subjects as 'R.f. bridges' by Professor D. Woods of the University of Surrey, 'Dielectric measurements' by Arnold Lynch (Post Office Research), 'Estimation of uncertainties' by Laurie Hinton (British Calibration Service) and 'Do's and don't's for r.f. measurements' by Norman Abbott (Electrical Quality Assurance Directorate of the Ministry of Defence).

Laboratory management will also be discussed at the School following an introductory lecture 'Management of measurement laboratories' by Brian Murphy (Marconi Instruments).

Companies who send employees to this vacation school may record the time spent as days of training for General Grant claim from the Engineering Industries Training Board.

The provisional programme and registration form for the Vacation School is now available from the Secretary, Institution of Electrical Engineers, Savoy Place, London WC2R 0BL, quoting the reference LS(S).

European Microwave Conference

The 1973 European Microwave Conference, the third in the series, incorporates the former Microwave and Optical Generation and Amplification Conference. It will be held at the Free University of Brussels from 4th to 7th September 1973. As well as a large number of papers to be presented in three parallel sessions, there will be 12 invited papers which are representative of the scope of the conference.

Registration forms and the preliminary programme are now available from Dr. G. Hoffman, Secretary General, 1973 European Microwave Conference, St. Pietersnieuwstraat 41, B—9000, Gent, Belgium.

Conference on Precision Electromagnetic Measurements

The 1974 Conference on Precision Electromagnetic Measurements will be held from 1st–5th July 1974, at the Institution of Electrical Engineers, London, England.

This will be the 9th Conference in the biennial series which began in 1958 and is the first to take place outside the United States. The change of venue on this occasion is a recognition of the increasingly international character of the Conference, reflected in the theme of the 1974 meeting which is 'the advancement and application of precision electromagnetic measurements against a background of international scientific co-operation'.

The Conference is being organized by the Royal Society and the Institution of Electrical Engineers in conjunction with the following co-operating sponsors: Union Radio-Scientifique Internationale (URSI), the Institution of Electronic and Radio Engineers, the National Physical Laboratory and the Scientific Instrument Manufacturers' Association. The Conference also has the benefit of the advice and support of the permanent sponsors of the CPEM in the United States. The programme will continue to cover the fields traditional to CPEM of d.c. and a.c. electrical measurements, time and frequency and r.f. and microwave measurements, although in the light of the growing diversification in applications of electrical measurement techniques a number of additional topics have been proposed for the 1974 Conference.

Contributions of original papers in the following subjects are invited: d.c. and l.f. measurements; r.f. and microwave measurements; time and frequency; time domain measurements; measurements at millimetre and sub-millimetre wavelengths; laser measurements; and automated and computer-controlled measurements.

Contributions on the following areas of topical interest are also invited; the speed of light; electromagnetic measurements of length and position; measurements on dielectric materials; health and safety; r.f. interference and the electromagnetic environment; precision industrial measurements; management and economic factors in measurement; and international co-operation in the field of precision electromagnetic measurements.

Papers should be submitted to the 1974 CPEM Secretariat, c/o the Conference Department, Institution of Electrical Engineers, Savoy Place, London, WC2R 0BL. Authors should contact the IEE 1974 CPEM Secretariat concerning the submission of abstracts for consideration. Special manuscript paper and guide notes will then be supplied, sufficient for an abstract not exceeding 3 A4 pages (approximately 1,500 words of text if diagrams are not included) which should be submitted by 31st October 1973.

Circuit Theory and Design

A call for papers has been made for the 1974 European Conference on Circuit Theory and Design to be held at the Institution of Electrical Engineers, Savoy Place, London WC2R 0BL from 23rd to 26th July 1974.

The Conference will embrace all aspects of the analysis, synthesis and design of lumped-element, distributed and digital networks but papers on the application of modern theoretical developments to engineering design, either of a research or tutorial nature, would be particularly welcome. It is intended that the Conference will provide a forum for the exchange of ideas between circuit theorists and circuit designers.

Offers of contributions not exceeding 3,000 words (a maximum of six A4 pages including typescript and diagrams) for consideration for inclusion in the Conference programme

Obituary

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

Gerald Francis Larner Beresford (Member 1962) died suddenly on 8th February after a heart attack at the BBC Television Centre. He was fifty-six years of age. After studying at Woolwich Polytechnic Mr. Beresford worked with Burndept Ltd., as a Junior Research Engineer until 1938 when he joined the BBC at the London Television Station. At the outbreak of the War he transferred to Brookmans Park Transmitting Station and then to Daventry and he spent some time in Italy during the War installing and operating an Allied Forces transmitter. He returned to BBC service on Television Outside Broadcasts in 1946 and in the late 1950s transferred to Designs Department, working in particular on experimental equipment for the 'Roving Eye' Unit. Mr. Beresford returned to Outside Broadcasts in 1961 and during the latter years of his service he worked on a number of O.B. projects and on power distribution at the Television Centre.

Daniel Eduardo Frias (Fellow 1971, Member 1960) died suddenly on 30th November 1972 aged fifty-eight. He leaves a widow.

Born and educated in Argentina, Mr. Frias studied radio and electrical engineering at the Radio Institute of Buenos Aires and he subsequently entered the Ministry of National Communication as a telecommunications engineer. In the forties and early fifties he held various appointments concerned with both telecommunications and broadcasting and also held part-time posts at the National University of Tucuman. At the time of his election to Membership of the Institution, Mr. Frias was in charge of technical services for the Ministry of Aeronautics at Tucuman Airport. Subsequently he was appointed Director of Communications in the Province of Tucuman, the post he held at the time of his death.

Cyril Vernon Jarvis (Member 1957, Graduate 1951) died on 20th March last aged fifty-seven years. He leaves a widow.

Educated at Southend High School Mr. Jarvis joined the Marconi Company in 1936 as a Junior Technical Assistant in the Instrument Test Division. Promoted in 1940 to Technical Assistant and then in 1946 to Senior grade, he worked from 1940 to 1952 in the Test Division, first in the Receiver Section on d.f., s.s.b., and field strength measuring sets and from 1949 in the Standards and Equipment Section.

From 1952 until his death Mr. Jarvis was with AWRE at Foulness, being promoted from experimental Officer to Senior Experimental Officer in 1955. From 1952–55 he worked on blast research and subsequently on design and development of high speed pulse measurement and recording apparatus.

Wing Commander Kuahiraman Meethal Kunnummal Nambiar, IAF (Member 1958, Associate 1953) died on 3rd April 1973 aged sixty years; he leaves two sons and a daughter.

After obtaining a B.A. degree in Mathematics at Madras University in 1936, Wing Commander Nambiar joined the Air Force in India in 1943, serving as a Radar Operator and later as a Radio Mechanic. He was commissioned in the Technical Signals Branch of the Indian Air Force in 1948 and subsequent appointments were with Hindustan Aircraft Ltd., Bangalore, and with mobile Signals Units. While at Operational Command Headquarters he was concerned with installation, repair and maintenance policies and also on instructional duties.

On his promotion to Squadron Leader in 1956 he was appointed an Assistant Director of Signals at Air HQ, New Delhi, subsequently serving in the Joint Cipher Bureau. In 1963 he was promoted Wing Commander and appointed to the IAF Technical College, Bangalore. From 1965 he was with Hindustan Aeronautics Ltd., Nasik, as Chief Communications Officer.

Charles Lionel Philps (Fellow 1958) died on 30th March 1973 at the age of sixty-six years. He leaves a widow, son and daughter.

Mr. Philps was born at Leigh-on-Sea and attended Felsted School, Essex. He joined the RAF in 1940 and was commissioned as a Signals Officer in 1942. He was mainly concerned with radar countermeasures and received a Mention in Despatches in 1945 after commanding a Mobile Unit in Europe. After demobilization in 1946 he joined Atkins, Robertson and Whiteford Ltd. in Glasgow, first as Sales Manager and then in 1948 became Managing Director in charge of development and production. During this period Mr. Philps was an active supporter of the Scottish Section.

In 1960 he was appointed Commercial Manager responsible for instrumentation and control equipment with STC at Enfield. In 1963 he moved to Harlow as Marketing Manager for rectifier equipments and in 1966 was appointed Manager of the Company's North Eastern regional office at Horsforth, Leeds. After a brief period in 1971 at Basildon, he became Resident Manager in Baghdad in January 1972. He was taken ill there and was flown home in March 1973.

Forthcoming Conferences *(cont.)*

are invited. Those intending to offer a contribution should submit a synopsis of approximately 100 words to the Conference Department, IEE, Savoy Place, London WC2R 0BL before 1st October 1973. Typescripts will be required for assessment by 18th February 1974.

The 1974 European Conference on Circuit Theory and Design is being organized by the Electronics Division of the Institution of Electrical Engineers in association with the Institute of Physics, the Institute of Electrical and Electronic Engineers (UKRI Section) and the Institution of Electronic and Radio Engineers. It is envisaged that this Conference will be the first of a biennial series on the same topic and that future conferences will be held in other European centres.

Further details are available from the Manager, Conference Department, IEE, Savoy Place, London WC2R 0BL.

IBC '74 Date Announced

The fifth International Broadcasting Convention—IBC '74—will be held in London at the Grosvenor House, Park Lane, from 23rd to 27th September 1974. Technological advances in all aspects of broadcasting will be highlighted both in the technical sessions and in the exhibition of all types of broadcasting equipment, both radio and television.

Among the items to be covered in the technical sessions are: automation broadcasting; training and management; future maintenance philosophy; propagation and service planning; receivers; recording, storage and replay; satellites in broadcasting; signal distribution systems; signal sources; sound systems—stereophonic and quadraphonic; technical aspects of international programme exchange; and transmitters, transposers and aerials.

The sponsors of the IBC are the Electronic Engineering Association, the Institution of Electrical Engineers, the Institute of Electronic and Radio Engineers, the Royal Television Society and the Society of Motion Picture and Television Engineers.

IERE members serving on the Management Committee are Mr. P. L. Mothersole (Fellow), who is also Chairman of the Technical Programme Committee, and Mr. R. S. Roberts (Fellow).

All enquiries in connexion with IBC '74 should be addressed to the Secretariat, International Broadcasting Convention, IEE, Savoy Place, London WC2R 0BL.

Members' Appointments

BIRTHDAY HONOURS

The Council has congratulated the following member of the Institution whose name appears in Her Majesty's Birthday Honours List published on 2nd June 1973:

TO BE AN ORDINARY OFFICER OF THE MILITARY DIVISION OF THE MOST EXCELLENT ORDER OF THE BRITISH EMPIRE (O.B.E.)

Lieutenant-Colonel Malcolm Hudson MacKenzie-Orr, RAOC (Graduate 1965).

Col. MacKenzie-Orr was from 1970 to the end of last year Chief Ammunition Technical Officer (Guided Weapons) at the Headquarters of the Director of Ordnance Services, Didcot. Since January he has been Chief Ammunition Technical Officer responsible for bomb disposal operations in Northern Ireland.

CORPORATE MEMBERS

Brigadier R. Knowles, C.B.E. (Fellow 1962) has been appointed to HQ Technical Group REME, as Project Manager responsible for co-ordinating all activities involved in the development and implementation of future REME management and engineering information systems. Brigadier Knowles, who is a member of the Institution's Council and serves on the Professional Activities Committee and on Group and Conference Organizing Committees, previously held a staff appointment at the Ministry of Defence in which he was responsible for studying equipment standardization.

Mr. P. H. Reynolds (Fellow 1960, Member 1965, Graduate 1950) has been promoted to Vice-President of Engineering of James G. Biddle Co., Plymouth Meeting, Pennsylvania. He joined the company, which manufactures electrical instruments, as Chief Engineer in 1967, following appointments with various U.S. and Canadian firms. Before going to Canada in 1957 Mr. Reynolds was a Senior Development Engineer with Elliott Bros. Ltd., of Lewisham.

Mr. K. R. Barham (Member 1972, Graduate 1967) is now Engineer-in-Charge of Studio Operations with Rediffusion Cablevision, Bristol. He was previously with Rediffusion Industrial Services Ltd., as a Commissioning and Project Engineer for three years and with Rediffusion Vision Services Ltd., as a Development Engineer from 1966 to 1970.

Captain P. C. Coderre (Member 1971, Graduate 1967) has been appointed Staff Officer Engineering and Maintenance 2 at Headquarters Canadian Forces Communication Command in Ottawa. Captain Coderre was formerly Deputy Commanding Officer, Signals Squadron at Headquarters 4 Canadian Mechanized Battle Group based in Lahr, West Germany.

Mr. B. N. Cotton (Member 1971), who is in the Department of Electronic and Radio Engineering at Riversdale College of Technology, Liverpool, has been promoted to Principal Lecturer in Marine Electronics.

Mr. G. J. Gibbs (Member 1973, Graduate 1968) has joined the Canadian Marconi Company in Montreal as an engineer concerned with the applications of the Omega Navigation System. He was previously concerned with this work with Marconi-Elliott Avionic Systems Limited at Basildon Essex.

Mr. W. S. Hodgson (Member 1969) has been promoted to Senior Electrical Engineer in the Traffic Control and Communications Division of the Department of the Environment. Before joining the D.o.E. three years ago he was an Executive Engineer with the Post Office.

Mr. H. G. Kendall (Member 1972, Graduate 1963) has been appointed Vice-Principal and Head of the Department of Engineering and Science at East Monmouthshire College of Further Education, Pontypool. For the past four years, Mr. Kendall was Senior Lecturer in Electronic and Electrical Engineering at the College which he joined in 1958 after working in industry.

Mr. P. N. Kirwan (Member 1972, Graduate 1967) is now a Consulting Engineer with USM Corporation, Beverly, Mass. U.S.A. For the past three years he was a Project Engineer with Fairbanks Morse Inc., which he joined following experience with the Marconi Company and in the Royal Air Force.

Mr. S. C. Lintern (Member 1972, Graduate 1968) is now a Main Grade Engineer with the South East Metropolitan Regional Hospital Board having special responsibility for electronic bio-medical equipment. Until February of this year Mr. Lintern was a Senior Electronics Engineer with Marconi Space and Defence Systems Limited, Rochester.

Mr. John Roberts (Member 1970) has been appointed Senior Engineer (Planning) —Telecommunications with West Midlands Gas. Before returning to the UK in 1971, Mr. Roberts was Telecommunications Engineer for some five years with the Zambian General Post Office.

Mr. J. Sears (Member 1957, Graduate 1950) is now a Senior Scientific Officer in charge of the Electronics Section of Met. O. 15 of the Meteorological Office, Bracknell. Mr. Sears was previously with the Science Research Council, London, following a number of years with the Atomic Energy Research Establishment.

Mr. D. R. Staples (Member 1972, Graduate 1966) is now a Principal Electronic Development Engineer in the Electro-Optical Systems Division of Marconi-Elliott Avionic Systems Limited, Basildon. He was previously at the Braxted Park Laboratories of the Plessey Company, Witham.

Mr. P. Sumner (Member 1970, Graduate 1962) has been appointed Engineering Manager, Crossbar Switching, with Standard Telephones and Cables Limited, at East Kilbride. He has worked on telephone exchange development with the Company for the past seven years.

Mr. T. C. Wright (Member 1969, Graduate 1963) is now a Professional and Technology Officer II at the National Gas Turbine Establishment. He was previously at Headquarters Signals Command, RAF Medmenham.

NON-CORPORATE MEMBERS

Mr. K. E. Barnett (Graduate 1970) has been promoted to Plant Automation Engineer with the British Steel Corporation, Strip Mills Division, Llanwern. He joined BSC as an electrical apprentice in 1961.

Mr. W. D. Brunt (Graduate 1953) is now Engineering Manager, Electronics, with the Sperry Gyroscope Division of Sperry Rand Limited, Bracknell.

Mr. A. F. Cross (Graduate 1972) has joined Thames Television Limited, Teddington as a Digital Design Engineer. He was previously a Development Engineer with Data Dynamics Limited of Hayes.

Mr. J. N. Elven (Associate 1966) has recently been appointed Divisional Sales Manager of the Analytical Division of Rank Precision Industries, Margate. He was previously for a number of years with Standard Telephones and Cables Limited, latterly as Head of Market and Product Planning with the Data Equipment and Services Division.

Mr. S. H. Gruszka (Graduate 1971) has joined Fairchild Semiconductor Limited as an Internal Technical Sales Engineer following a number of years with the Metal Box Company as a Development Engineer.

Mr. H. Javid (Graduate 1964) is now Managing Director of Flange Industrial and Commercial Company, Tehran. He was previously Central Project Engineer with TAI Inc.

Mr. J. M. Myler (Graduate 1969) has been appointed Computer Systems Engineer with Allied Business Systems Limited; he was previously a Customer Engineer with Control Data Limited.

Mr. R. R. Sücker (Graduate 1969) has been promoted to Senior Transmitter Engineer at the British Broadcasting Corporation's Station at Skelton, Cumberland.

Mr. A. R. Wood, B.A. (Graduate 1973) has joined Marconi Space and Defence Systems Limited, Camberley, as a Junior Engineer in the Radar Control Division following the completion of a Bosworth Course in radar technology at Birmingham University.

Government and Parliament

Industrial Technologies Committee Structure Complete

The six chairmen of the Committees being set up to advise the Department of Trade and Industry's Committee for Industrial Technologies have now all been appointed.

They are:

Committee on Tribology

Dr. H. Peter Jost, Managing Director, K.S. Paul Products Ltd.

Committee on Terotechnology

Major General Sir Leonard Atkinson (Past President of the IERE), Director, Weir Engineering Industries.

Committee on Education and Training

Dr. A. H. Chilver, Vice Chancellor, Cranfield Institute of Technology.

Committee on Corrosion

Mr. J. B. Cotton, Part-time Consultant, Imperial Metal Industries.

Committee on Materials Handling

Mr. H. Surtees, Managing Director, EASAMS.

Committee on Information and Publicity

Mr. N. Shepherd, Chief Information Officer, Department of Trade and Industry.

The Committee for Industrial Technologies, set up in 1972 under the Chairmanship of Mr. Gilbert A. Hunt, Managing Director of Chrysler UK, is responsible for promoting a greater awareness in industry of the importance of these multi-disciplinary technologies. In addition to industrial and academic members, it includes the Chairman of the Science Research Council, Sir Brian Flowers, and Mr. M. W. Leonard, Secretary of CEI.

It has been estimated that industry could save as much as £1500 million a year by planning the results of technological research and existing knowledge in these disciplines.

Technical Harmonization within the EEC

The European Economic Commission has proposed to the Council of Ministers a new series of harmonization proposals on which it proposes to submit directives by January 1974 for Council decision by July 1974. These would deal with the elimination of technical trade barriers in respect of aircraft, railway stock, motor cycles, fuels, fire fighting equipment, firearms, telecommunications equipment, toys, school equipment, packaging, farm tractors and motor vehicles.

Engineers' Registration Board

At question time in the Commons on May 14th, Mr. E. Bishop (Lab., Newark) asked the Secretary of State for Trade and Industry what action he had taken in the past year to resolve some of the outstanding problems arising from the setting up of the Engineers' Registration Board and what discussions were currently taking place with the Council of Engineering Institutions and the professional institutions.

Mr. M. Heseltine, Minister for Aerospace and Shipping, replied that the setting up of the Board was of direct concern to the profession. His concern was that, while raising standards generally, this should not discriminate against a significant minority and affect their opportunities at home and in Europe. Discussions between his officials and those of other interested Departments and the profession had been continuing during recent months. A range of possibilities had been considered, but he could not yet forecast the solution. A solution had to be found before EEC directives were determined.

EEC and British Patents

Answering questions in the Lords, Lord Drumalbyn, for the Government, said that the European and the Community Patent Conventions were expected to enter into force not later than 1976, but the British patent system would continue to exist after that date. An applicant would therefore have the alternative of a European patent or a purely British patent. It was intended to amend British patent law so that it would be strengthened in accordance with the main recommendations of the Report of the Banks Committee and to bring it into line with European Patent Law. It was expected that some European patent work would for some time be done in the British Patent Office on an agency basis. As the European system got under way the number of British patent applications files would diminish, so there would be problems of adjustment; but they intended to maintain an efficient British Patent Office operating side by side with the European patents system.

He added that it was expected that the European Patent Convention would be concluded at a diplomatic conference in Munich in September 1973, and the Community Patent Convention would be concluded at a further conference in May 1974.

New Diploma in Finance and Accounting

DAVID EAGLE*

With the increasing size and complexity of modern industrial and commercial organizations, the task of planning the deployment of their capital resources and the control of their day-to-day activities depends on the availability and use of an effective management information system. Modern management accounting techniques are an essential link in this chain of control—facilitating the forecasting of trading strategy and future capital requirements, the budgeting of planned revenue and expenditure, and the evaluation of capital investment projects.

However, the effective use of such a system depends on the individual manager having a basic knowledge of the elements of business finance and accounts and the ability to understand and correctly interpret the financial information he is given. The manager who has been trained in the use of management accounting information is obviously in an advantageous position as he can apply his knowledge to make better decisions and to suggest better solutions to business problems. His personal career prospects are thus enhanced

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* Association of Certified Accountants. 22 Bedford Square, London WC1B 3HS.

by his greater all round knowledge of the financial background to industry and commerce.

However, despite the need for engineers in management to gain this knowledge of finance and accounts, few can entertain the possibility of obtaining a full professional accounting qualification because the examinations are extensive and no professional accountancy body will admit them to membership until they have acquired considerable practical experience.

To meet this need and indeed to bring about greater understanding between the accountant and his managerial colleagues, the Association of Certified Accountants has introduced a scheme for the award of the Certified Diploma in Finance and Accounting which demonstrates that candidates have obtained a sound appreciation of the principles of accounting, taxation, management accounting, financial planning and finance.

Enrolment on the scheme is open to professional engineers and the recommended period of study leading to an examination of four papers is one year. Tuition can be obtained either by part-time evening or afternoon and evening courses run by polytechnics and colleges of further education or by the correspondence course prepared by the Association, part of which is full-time study during a period of five days held in London on a non-residential basis.

After passing the examinations a candidate becomes a subscriber to the Association, is awarded the Diploma and may use the designatory letters C.Dip.F.A. The professional engineer thus has a sense of association with the accounting world, receiving regularly the journal of the Association and being free to use the Association's library.

Further details can be obtained from the Secretary of the Diploma Scheme, Association of Certified Accountants, 22 Bedford Square, London WC1B 3HS.

Letter to the Editor

The Institution's Council does not necessarily agree with views expressed by correspondents.

Correspondence of a technical nature, or on any matter of interest to electronic and radio engineers, is welcomed.

From: Mr. L. S. Cornish, B.Sc.(Eng.), C.Eng., M.I.E.E.

Insulation Faults revealed by Voltage Surge Tests

Recent voltage surge tests carried out on electronic equipment* revealed a number of potentially dangerous faults. Although the equipments tested were designed for airborne applications, the faults described could just as easily have occurred in laboratory equipment.

The two equipments in question were designed to operate from a $+28$ V d.c. supply. When during testing, a $+80$ V surge of 60 milliseconds duration was applied to the equipment supply terminals (as required by the equipments' specifications), both equipments presented a short-circuit to the supply which remained after the surge was removed. Both equipments had been functioning normally before the application of the surge. Fortunately in both cases protecting fuses were blown which prevented serious damage occurring.

In the first equipment, the fault was caused by a badly finished heat sink on which the power transistor of the series voltage regulator was mounted. The holes in the heat sink had not been de-burred and the mica insulating washer between the transistor case (which was connected to the transistor collector and the positive supply terminal) and the heat sink (which was earthed) was punctured. This caused a short circuit between the positive supply and earth.

In the second equipment there was insufficient clearance between the circuit components and the chassis, which by coincidence also produced a short circuit between the positive supply terminal and earth.

In both cases the remedy was straightforward. In the first case the quality of the heat sink was improved, with proper de-burring of rough edges and greater care was taken in handling the mica washer which is quite fragile. In the second case it was a simple matter to increase the clearances. Additional protection could have been provided by sticking a piece of insulating material to the chassis adjacent to the circuit components.

The problem of clearances between surfaces of differing electrical potentials should be especially carefully considered when the equipments concerned are to be used in environments where there are high levels of vibration. Clearances can be drastically reduced in the presence of an unfavourable mode of vibration.

L. S. CORNISH

Electrical Engineering Department,
University of Hong Kong.
5th March 1973.

Editorial note.

Mr. Cornish's letter was written in response to an invitation made to readers of the Journal in an article on Electrical Accidents in Factories (December 1972) to submit brief accounts of ways in which hazards have arisen (or could arise) in laboratories or factories and have been dealt with.

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* Cornish, L. S. 'Transient-testing vital to airborne systems reliability,' *Electronic Engineering*, 44, pp. 61–3, September 1972.

Industrial News

Safety Information Sheets

A great many of the expensive and tragic accidents caused by fire every year could be prevented—by simple common-sense precautions, claims the Fire Protection Association. To promote this basic approach, the FPA is producing a series of fire safety sheets, intended to put over simple safety rules concerning a variety of common industrial products, practices and hazards.

Among the first four sheets is 'Electric Soldering Irons' (S1). These ubiquitous tools of the laboratory often set fire to neighbouring combustible materials and the sheet explains, with the help of illustrations, such points as hazards to contend with, precautions to avoid fire and simple instructions for correct use.

The sheets (which also include 'Fire Blankets', 'Waste Incinerators' and 'Oil soaked floors') are for use by works fire and safety officers, for distribution to supervisory and other staff and may also be displayed as 'mini-posters' in appropriate parts of the premises. Sheets S2–S4 cost 5p each or 12p for any four and are available, along with details of the FPA's membership scheme, from the Fire Prevention Information and Publications Centre, Aldermary House, Queen Street, London EC4N 1TJ.

Among other publications by the FPA, Data sheet FS 6005, 'Automatic Fire Alarm Systems' is the latest addition to a series that deals with specific aspects of fire technology relating to industry and commerce. This costs 10p each, from the same address.

British Standard for Insulated Screwdrivers

The use of insulated screwdrivers is, or should be, second nature to engineers when working with electrical apparatus. What is perhaps not generally realized is that insulated tools are implicitly required by law in such circumstances as a result of electricity regulations associated with the Factories Act, which also demand their periodical examination by an authorized person.

To provide an authoritative specification for the adequate design of such screwdrivers—and for the benefit of purchasers —the British Standards Institution has published BS 2559 Screwdrivers: Part 3: 1972, Insulated screwdrivers. The Standard covers the materials and thickness of the insulation suitable for the tools specified in other parts of BS 2559, and gives a test which utilizes a potential of 5kV. Manufacturers of screwdrivers complying with this standard may apply to BSI to obtain a licence to use the Kitemark, which provides an independent assurance from BSI that the product does comply with the standard as claimed. Copies of BS 2559 Part 3 cost 55p each including postage (plus 20p for orders under £2.00), from BSI Sales Branch, 101 Pentonville Road, London N1 9ND.

A Record Year for Consumer Electronics in Great Britain

Total television receiver deliveries to the UK market during 1972 reached a record 3,607,000, representing a rise of 34% over 1971 (2,690,000) according to the latest figures compiled by the British Radio Equipment Manufacturers' Association. Most of this increase was due to the great increase in colour television deliveries which for 1972 exceeded those for mono-chrome receivers for 1971. Deliveries of colour television in December of 192,000 gave a total for the year of 1,775,000, a rise of 93% over 1971 (921,000). Monochrome television deliveries of 161,000 sets for December resulted in the total for 1972 reaching 1,832,000, a rise of 4% over 1971 (1,769,000).

Radio receiver deliveries of 520,000 for the month brought a 1972 total of a record 6,369,000, compared with 4,895,000 for 1971—a rise of 30%. Deliveries of record players reached 568,000 for 1972, compared with 587,000 in 1971, a fall of 3%; in addition to these figures, deliveries of an estimated 1,050,000 stereo systems were made during the year, against approximately 640,000 in 1971. Radiogram deliveries reached 267,000 for the year, compared with 233,000 for 1971 (a rise of 14%). These figures are estimates of UK made and imported deliveries to the home market, including those to specialist rental and relay companies.

Explaining BS 9000

A major revolution in the specification and stocking of electronic components is described in 'Profit from BS 9000', an information booklet published for the British Standards Institution by the Department of Trade and Industry and now available. Intended for electronic equipment manufacturers and users, the booklet explains, using mainly a 'question and answer' presentation, the BS 9000 System for Electronic Components of Assessed Quality, which offers a single system of specifications for all kinds of electronic components. This cannot be other than welcome to designers, purchasers, stock control managers and others in the industry who, for years, have had to contend with a great multiplicity of specification systems.

Since the standards in the BS 9000 system all contain quality assessment data presented in the same way, it will be much easier for users to buy the right components and to ensure that their equipment has the desired levels of quality and reliability in its intended environment. The system was pioneered in this country by BSI, and a new corresponding system has been set up in Western Europe with which the BSI system is aligned while a world-wide system is also under consideration.

The booklet is aimed at purchasers of electronic components to persuade them to adopt the System and to ensure that British companies get the full benefit from this important advance. The BS 9000 System was featured on the BSI stand at the recent London Electronic Components Exhibition.

EEA and CAMA

The Electronic Engineering Association and the Control and Automation Manufacturers' Association have announced that agreement has been reached on arrangements to improve co-operation in the fields of Industrial Control and Auto-mation, with the object of enhancing the effectiveness of the two Associations in these fields.

To this end the two Associations intend to initiate certain joint activities and share some services and, by this means, improve communications and cost effectiveness, reduce over-lap and advance the introduction of Automation into manu-facturing industries.

Germany Equipment Safety Law

With exports from the UK amounting to over £539M in 1972, the Federal Republic of Germany ranks second only to the USA in terms of importance as a British export market. For manufacturers interested in its highly attractive export potential, a thorough understanding of the effects of the German Law on technical equipment is of vital importance in ensuring a successful venture.

The Law (generally known as the Equipment Safety Law) has wide applications which can affect the majority of firms intending to export to the Federal Republic of Germany. Under this Law all manufacturers of both industrial and consumer goods are required to ensure before displaying for sale, or before delivery, that equipment is safe from hazards to life and health (particularly accident hazards). Those products affected include tools, instruments, machines, household equipment, personal safety equipment, toys, games and sports goods.

The Technical Help to Exporters (THE) service of the British Standards Institution has now produced a new publication (*General P*.106). *Federal Republic of Germany—Equipment Safety Law*. It consists of a translation of the full text of the framework law and the subsequent administrative regulations with detailed explanatory notes. It also included lists of standards and special regulations governing technical safety which have been nominated as setting compliance requirements within the meaning of the Law, and a comprehensive list of other relevant publications.

This THE publication is available from Technical Help to Exporters, Maylands Avenue, Hemel Hempstead, Herts HP2 4SQ, price £4.00 to THE members and £5.00 to non-members, postage included.

The Indian Computer Industry

Over 160 computer systems are in use in India today. The first digital computer in the country, a British Tabulating Machines *HEC 2*, was installed in 1956 at the Indian Statistical Institute (ISI), Calcutta; the Institute subsequently acquired a Russian *Ural* computer.

It was in 1961 that the first indigenous electronic digital computer TIFRAC was built and commissioned by the Tata Institute of Fundamental Research, who later developed an on-line data precesser. The Jadavpur University in Calcutta were successful in developing their computer ISIJU in 1965 in collaboration with the ISI.

The Electronics Group at the Bhabha Atomic Research Centre (BARC) gave to the nation their first high-speed real-time digital computer, the *TDC 12* which can perform 250 000 additions or subtractions per second. In 1968, the commercial manufacture of the *TDC 12* system was taken up by Electronics Corporation of India (ECIL), a department of Atomic Energy Undertaking in the public sector, and the first *TDC 12* was completed by ECIL's Computer Division in March 1971. The Corporation have an ambitious programme to manufacture and install over 130 computers by 1975–76: the target of ten computer installations in 1972–73 is to be increased five-fold for 1975–76, and capital expenditure by that year is expected to reach a level of Rs. 43 M providing employment to over 2000 engineers, scientists, technicians and supporting staff.

ECIL have plans to market a larger third-generation system, the technology of which is expected to be on par with similar systems elsewhere in the world, both in terms of performance and reliability. Applications already envisaged for ECIL's third-generation computers are the control of nuclear reactors, space research, defence and scientific and data processing.

India has had the benefit of the experience of international companies including IBM and ICL. IBM have been active in the country since 1950 and started manufacturing there in 1963. Their plant in Bombay was set up to make key-punches and recondition and assemble 1401 central processors, unit-record and peripheral equipment. Interestingly, the key-punches manufactured in India were even exported to IBM's overseas branches and subsidiaries.

ICL also started its manufacturing operations in India in 1963. Their area of specialization relates to the assembly of off-line card preparation equipment in their key plant at Poona. Non-printing key-punches, verifiers, duplicators, interpreters, etc., are also among the products manufactured indigenously. This factory is also engaged in the import and assembly of peripheral equipment as a supplementing measure to the current manufacturing activity in India.

Recently, ICL have started manufacturing 1901A series of computers in collaboration with Bharat Electronics Limited of Bangalore. The indigenous effort of BEL is to be in back panel wiring assembly, mainframe assembly and final testing.
Based on an article in 'India Weekly' for 25th January 1973.

Japanese Television Production Statistics

In a recent financial newsletter issued by the Fuji Bank of Tokyo, industrial trends in Japan for household electrical appliances (which includes television receivers) were reviewed as follows:

'Production of household electrical appliances, which declined by 2·1% in 1971, will rise by 8·6% in 1972 to an estimated total of 2 170 000 yen (£3000 M). Shipments of colour television sets, which increased by 29·2% in 1971 to a total of 7 470 000 units, are expected to level off because the domestic market is saturated and exports to Europe and the USA are unfavourable. This year's colour television output is estimated at 7 500 000 units. Exports of appliances showed increases of 17·0% in 1970 and 19·0% in 1971 but the rate for this year is forecast at only about 10%, reflecting stagnation of exports in both colour and black-and-white television sets.'

Confirmation of these estimates, prepared before the end of 1972 has not yet been received.

Conference of the Electronics Industry —BEAMA Joint Working Party

Discussions have proceeded for some time past on the possibility and desirability of achieving closer working arrangements between the CLI and the BEAMA. A joint statement points out that the problem is a complex one requiring careful analysis and consideration. It has therefore been decided as a practical step to set up a small joint CLI-BEAMA Working Party to examine the problem how best to achieve closer liaison. The Chairman of the Working Party is Sir Leslie Robinson, K.B.E., C.B.

The Working Party held its initial meeting in April and is planning to report its findings to the Councils of the CLI and of the BEAMA in the early autumn.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

IN ACCORDANCE with Bye-law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details. These proposals for the election and transfer of 55 candidates to Corporate Membership of the Institution and the election and transfer of 7 candidates to Graduateship and Associateship arise from recommendations made by the Executive and Membership Committees at their April and May meetings.

Membership Approval List No. 158

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

BRADNAM, Ronald Hastings. *Hartley Wintney, Hampshire.*
DANN, Paul Charles David, B.Sc.(Eng.). *Chelmsford, Essex.*

Transfer from Companion to Fellow

H.R.H. The Duke of Kent, G.C.M.G., G.C.V.O. *London, S.W.1.*

Direct Election to Fellow

PARKER, Bernhard Dollman. *Morden, Surrey.*

Transfer from Graduate to Member

ADAMSON, Simeon Ronald, B.Sc.(Eng.), M.Phil. *Wokingham, Berkshire.*
AIRD, Robert Alexander. *Crowthorne, Berkshire.*
APPLEBY. John David, B.Sc. *Thatcham, Berkshire.*
ASPINALL, Malcolm. *Wigan, Lancashire.*
BAKER, Peter John. *Woolston, Hampshire.*
BAYLY, Anthony Edward, B.Sc.(Eng.). Flight Lieutenant, RAF. *Nanstallon, Cornwall.*
BOOTH, David George. *Upper Skelmorlie, Ayrshire.*
CHRISTIE, Stanley. *Blackrock, County Dublin.*
CLARK, Tony. *Lincoln.*
COOMBES, Alan, B.Sc. *Stevenage, Hertfordshire.*
COOMBES, Michael. *Whyteleafe, Surrey.*
COUSINS, James Henry. *Woking, Surrey.*
DAY, Peter William. *Reading, Berkshire.*
DELGAN, Neville David. *Higham, Kent.*
DICKINSON, David Charles. *Bradford, Yorkshire.*
DYER, Peter Ernest Llewellyn. *Maidenhead, Berkshire.*
FINDLAY, Ian Michael. *Windsor, Berkshire.*
GAY, John Richard. *Danbury, Essex.*

GAZI, Cawasji Framroze. *Maldon, Essex.*
HAGGITT, Robert Edward, B.Sc. (Eng.). *Thornbury, Bristol.*
HALLIWELL, Raymond. *Albrighton, Staffordshire.*
HALLOWS, Peter James. *Ilford, Essex.*
HARLIN, Eric Alan. *Hazel Grove, Cheshire.*
HARRIS. Paul Graeme. *Ashchurch, Gloucestershire.*
HOLLAND, Geoffrey John, B.Sc. *Havant, Hampshire.*
KIRBY, Michael John. *London, S.W.19.*
MANT, Anthony Roger. *London, W.4.*
MOLYNEAUX. Martyn William. *Writtle, Essex.*
MOONEY, Michael Kevin. *London, S.E.12.*
MORLEY, Dennis Henry. *Long Ditton, Surrey.*
ROBERTS, Paul Anscombe. *Surbiton, Surrey.*
SCOTT, Barry Shepley. *Witham, Essex.*
TARBAT, David Fairlie. *Welling, Kent.*
WATSON. David, Squadron Leader, RAF. *Tilehurst, Berkshire.*
WHEATLEY, Henry. *Edinburgh.*
YEATES, David. *Southam, Warwickshire.*

Transfer from Associate to Member

OXLEY, Terence Hunter. *High Wycombe, Buckinghamshire.*

Direct Election to Member

ALLEN, William Francis. *St. Neots, Huntingdonshire.*
CANTON, Bryan Victor, Flight Lieutenant, RAF. *Cranwell, Lincolnshire.*
PATEL, Dahyabhai Chhitabhai, B.Sc., Ph.D. *Horndean, Hampshire.*
RILEY, Henry Rimmer, Group Captain, RAF. *Sealand, Flintshire.*
SHORT. James Adrian. *Ashford, Middlesex.*
TAYLOR, Allen, B.Sc., M.Sc., Squadron Leader, RAF. *Manby, Lincolnshire.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

RADHA KRISHNAN. Patchay. *Birmingham 6.*

Direct Election to Graduate

POPE, George Brian. *Milkwall, Gloucestershire.*
SIMISTER, Frederick. *Macclesfield, Cheshire.*

Direct Election to Associate

COOPER, James Arthur. *London, E.3.*

STUDENTS REGISTERED

CASPERD, Alan Norman. *Stocksfield, Northumberland.*
JEWELL, Timothy Robert. *Stocksfield, Northumberland.*
O'SULLIVAN, Gearoid Michael Charles. *Dublin 6.*
PATEL, Suresh. *London, W.9.*
SMITH, Ian George. *Chesterfield, Derbyshire.*
WATSON, Robert Victor. *Penylan, Cardiff.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

BEECH, Albert Lloyd. *Wellington 4, New Zealand.*
GLENDENNING, John Stuart Irving. *Kuala Belait, Brunei.*
MANGRU, Soney James. *Teheran, Iran.*
RWADEBA-RWLSE, Sam, B.Sc.(Eng.). *Nairobi.*
WAN, Seng Kong. *Singapore 21.*

Transfer from Associate to Member

DAVIS, Jack Haskel. *Ramat-Gan, Israel.*

Direct Election to Member

GHOSH Dipak Kumar, B.Sc.(Eng.). *Calcutta.*
HASAN, Mian Saleem, B.Sc. *Lahore, Pakistan.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

YOUNG, Neville. *Canberra, Australia.*

Direct Election to Associate

HARRIS, John, Warrant Officer, REME. *B.F.P.O. 51.*
YUN, Victor Khyam Yung. *Kota Kinabalu, Sabah.*

STUDENTS REGISTERED

BOROLE, Popat Digambar. *Poona, India.*
CHIN. Wou Chau. *Kuala Lumpur, Malaysia.*
KUMAR, Kadayam Parameswaran. *India.*
LA I. Yew Fatt. *Singapore 3.*

Notice is hereby given that the elections and transfers shown on List 155 have now been confirmed by the Council.

New Books Received

Microwave Transmission

J. A. STANIFORTH. English Universities Press Ltd., London 1972. 21·5 cm × 14 cm. 243 pp. £4·25 (hardback); £2·35 (unibook—limpcover).*

CONTENTS: Transmission Line Theory. Transmission Line Charts. Fields and Plane Waves. Waveguides. Two-conductor Lines. Other Waveguides. Microwave Circuits. Principles of Matching. Cavity Resonators. Propagation. Microwave Components. Appendices: General equation of a circle. General solutions for cylindrical waveguides. Table of rectangular waveguide sizes.

University texts are usually created to satisfy various well-established requirements and a textbook such as this one which falls short in certain areas can only be recommended with reservations. For instance both students and lecturers find examples and exercises of great assistance and this book contains only a few isolated examples. Another requirement is that a text should to some extent enlighten the student and supplement lecture material by injecting informative views on broader and perhaps fringe issues; here the text deviates little from a familiar stereotyped theme and somewhat resembles a set of lecture notes itself; in particular a final discussion chapter relating the fundamental issues to the microwave applications listed in the Editor's foreword would not be out of place. Measurement techniques are scattered around the text and are vital enough to justify a separate chapter. Personally I would have preferred an extended Bibliography and also some source references on industrial and communication applications. Appendix 1 seems a very trivial issue which could have been tucked into the relevant chapter and Appendix 2 is probably too fragmentary to help a student who wants this detail which in any case is well covered elsewhere.

On the credit side the book has an easy going pace, is easy to read, deals with the subject at an introductory level and at this reasonable price may be helpful to students who are facing the topic for the very first time.

J. R. JAMES

(*Dr. Staniforth is a Lecturer in Electrical Engineering at the University of Salford—see page* 388.)

The Scanning Electron Microscope.

Part 1. The Instrument.

C. W. OATLEY. Cambridge University Press, London 1972. 21·5 × 13·3 cm. 194 pp. £5·30.

CONTENTS: Historical Introduction. Fundamental Limitations and Design. The Component Parts of the Scanning Microscope. The Interactions of Electrons with a Solid. Contrast and Resolution.

The scanning electron microscope has provided a tool of immense practical value and many s.c.m. micrographs have been published in this *Journal* in connexion with the examination of the structure of integrated circuits and other components in which the effects of surface configuration are critical. This book is a useful guide to the user of the s.e.m., enabling him to understand its principles and limitations and thereby achieve the best results.

(*Professor Oatley, who is Emeritus Professor of Electrical Engineering of the University of Cambridge, developed the scanning electron microscope as a practical instrument*).

Terrestrial Propagation of Long Electromagnetic Waves

JANIS GALEJS. Pergamon Press, Oxford 1972. 25 cm × 17·5 cm. 362 pp. £14·00.

CONTENTS: Boundaries of the Terrestrial Waveguide. Natural Sources of Radiation. Waves in a Spherical Guide. Waves in a Cylindrical Guide. Fields in Stratified and Anisotropic media. E.L.F. Propagation and Schumann Resonances. V.L.F. Propagation. L.F. Propagation.

This monograph treats wave propagation in the space between the Earth and the ionosphere which is excited either by lightning discharges or by artificial sources. It is primarily a reference text for research workers in this field and it also provides a detailed basis for engineering applications as well as for geophysical applications.

(*The author, who is at present with the Naval Research Laboratory, Washington, D.C., carried out much of the work described in the monograph while in the Applied Research Laboratory of Sylvania Electronic Systems*.)

* These books may be ordered from the Institution at list price plus 25p to cover postage and packing. (See panel on page 351 of May issue.)

STANDARD FREQUENCY TRANSMISSIONS—April 1973

(*Communication from the National Physical Laboratory*)

April 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)		April 1973	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	0	0	0	714	617·2	17	+0·1	+0·1	+0·1	714	623·5
2	−0·1	0	−0·1	715	617·0	18	+0·1	+0·1	0	713	622·8
3	0	0	0	715	616·8	19	0	+0·1	+0·1	713	621·9
4	0	−0·1	0	715	617·4	20	0	−0·1	0	713	623·3
5	0	0	0	715	617·6	21	0	0	0	713	623·3
6	0	0	0	715	618·6	22	−0·1	0	0	714	624·2
7	−0·1	0	0	716	620·0	23	−0·1	−0·2	0	715	625·7
8	−0·1	0	0	717	621·4	24	0	0	0	715	625·3
9	−0·1	−0·1	0	718	622·1	25	0	−0·1	0	715	625·9
10	+0·1	+0·1	0	717	621·0	26	0	0	+0·1	715	626·2
11	0	+0·1	0	717	620·1	27	−0·1	−0·1	0	716	626·9
12	0	0	0	717	620·1	28	−0·1	0	0	717	627·2
13	0	0	0	717	620·1	29	0	0	0	717	627·6
14	+0·1	+0·1	0	716	619·5	30	−0·1	−0·1	0	718	628·4
15	+0·1	+0·1	+0·1	715	624·9	31					
16	0	+0·1	0	715	624·1						

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

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

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