

Founded 1925
Incorporated
by Royal Charter 1961

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

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

— And now, the next 25 years!

At any notable anniversary, such as the Silver Jubilee of Her Majesty Queen Elizabeth II, it is usually felt appropriate to look back over the period that is being commemorated. As members know, however, in 1975 the 50th anniversary of the Institution's foundation was duly commemorated in the Journal in this time-honoured way: while it would be presumptuous to take the view that all that needed to be said or recorded was in the editorial and in the papers of that October 1975 issue as well as in that of March 1977 where the President reviewed the 40 years of Mr. G. D. Clifford's secretaryship, a different approach should perhaps be adopted on this occasion.

The twenty-fifth anniversary of Her Majesty's accession therefore presents an opportunity to look ahead along the route which the Institution is to take in the coming years. We should consider how it will build on the foundations which have been laid in the past quarter century (and earlier) and respond to the challenges that will arise in the technical field as well as through the complexities of life in Britain in the last decades of the twentieth century.

The challenges set to the Institution by technological advances will surely call above all for flexibility in recognizing the changing role of the professional electronic engineer. Already he has to adopt a 'systems' approach in dealing with many projects, and frequently needs to incorporate special-purpose computer techniques, typified by the now ubiquitous microprocessor. Closer links between the systems engineer and the device engineer must be encouraged by the Institution through its liaison with the universities and colleges in providing tomorrow's engineers with a realistic technological base—not forgetting the need for re-training those who qualified in 'pre-l.s.i.' days. There are many new application areas, associated, for instance, with medical electronics, with intelligent machines for industry and commerce, as well as developments in communications generally, which the Institutions' future 'learned society' plans will need to cover.

An Institution concerned with a science as live and progressive as electronics has in fact great incentives continually to introduce forward-looking policies. As well as widening its coverage of technological affairs, the geographical distribution of members is broadening and this will call for more Local Sections. Collaboration with other organizations is increasingly necessary, whether these are from other disciplines, engineering or scientific, or the technician and technician engineer bodies. In these and in other new ventures, the IERE must go forward and to provide it with a base consistent with changing demands a move to a new headquarters in London is to be made soon.

But what about the electronic engineer as a member of the engineering profession? The supplemental charter of the Council of Engineering Institutions provides an opportunity, which must not be missed, of achieving a more integrated and cohesive profession, able to play a full part in the challenges of the modern world: the energy problem that is with us already and which will not be solved without the engineer's voice being heard is but one of these. Is it too much to expect that the coming years will see a greater measure of public recognition of the engineer's contribution to prosperity? It is the importance of this contribution that justifies referring to its place in this prospect of the coming years.

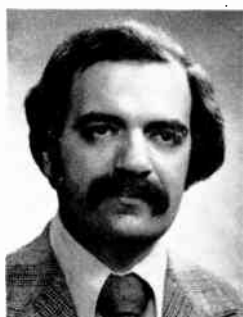
F. W. S.

Contributors to this issue*

Kamilo Feher, Associate Professor of Electrical Engineering at Concordia University for several years and a consultant to RCA Limited, has recently taken up an Associate Professorship in the Department of Electrical Engineering, University of Ottawa. A fuller biography was published in the April 1977 Journal.



Mohamed El-Torky received the B.Sc. degree in electrical engineering from Alexandria University, Egypt, and the B.Sc. degree in mathematics from the same University in 1971 and 1974 respectively. From 1971 to 1974 he was a Teaching Assistant at Alexandria University and currently he is working towards a doctorate at Concordia University, specializing in digital communications.



R. De Cristofaro received the Electronic Technician Diploma in 1966 from E.T.F.G. at Rio de Janeiro, Brazil and from 1965 to 1970 he worked as a Communications Engineering Technician. He then went to Canada and received the degree of Bachelor in Vocational Teaching (Electrotechnology) from the University of Quebec in Montreal. In 1976, he received a Master of Engineering Degree from Concordia

University for his work in the design and evaluation of a binary transversal filter for satellite communications. Since June 1975, he has been employed by Bell Canada in the Analog & Digital Transmission Design Group.



M. N. S. Swamy graduated in mathematics from Mysore University in 1954, and received the Diploma in electrical communication engineering from the Indian Institute of Science, Bangalore in 1957 and the M.Sc. and Ph.D. degrees in electrical engineering from the University of Saskatchewan in 1960 and 1963 respectively. He worked as a Senior Research Assistant at Bangalore until 1959, when he

began graduate study at the University of Saskatchewan, returning to India in 1963 to work for a year at the Indian

Institute of Technology, Madras. From 1964 to 1965 he was an Assistant Professor of Mathematics at the University of Saskatchewan and subsequently taught as Professor of Electrical Engineering at Nova Scotia Technical College, Halifax and the University of Calgary. He is now Chairman of the Department of Electrical Engineering, Concordia University, Montreal. Professor Swamy has published a number of papers on number theory, semiconductor circuits, control systems and network theory.



Marcel Bouthéon graduated from ENSIAM (France) in 1962 and the Polytechnic Institute of Grenoble in 1963, specializing in automatic control. Between 1963-65 he was a research assistant at the University of Grenoble and received his doctorate in engineering from this University in 1966. He was also a research officer for one and a half years with the French Navy between 1965-66. Since 1967, he

has been working at the European Organization for Nuclear Research (CERN), Geneva, Switzerland, as an engineer and supervisor for the computer control and operation of the CERN Proton Synchrotron Complex. His main interests are in the areas of electronics and machine physics specially relevant for the computer control of particle accelerators.



V. C. Chohan obtained a first class honours degree in electrical and electronic engineering at the University College of South Wales, Cardiff in 1971. This was followed by research through a scholarship donated by I.T.T. Components Group Europe, Harlow, for computer-aided-design, development and fabrication of optimum monolithic crystal-filter configurations used in f.s.k.-type applications. The

work was carried out partly at the University of Essex and partly in industry, resulting in a Ph.D. degree. Since January 1975, Dr. Chohan has been at CERN working for the group controlling and operating the proton synchrotron via a computer.



Jean-Pierre Potier graduated with a Master of Science degree in physics from the University of Poitiers in 1965. He continued his studies at the same University, carrying out research and development work in mass spectrometry which culminated in a doctorate in 1968. He joined CERN in 1969 and has been working since then in the controls and operations group of the Proton Synchrotron Division. His main

interests and responsibilities are in the areas of accelerator physics and computer software, especially related to the development and implementation of techniques for the computer control and operation of the CERN PS complex.

* See also pages 268, 273, 276 and 282

Optimum pulse shaping application of binary transversal filters used in satellite communications

K. FEHER, Ph.D., Eng.*

M. EL-TORKY, B.Sc.*

R. DE CRISTOFARO, M.Eng.†

and

Professor M. N. S. SWAMY,

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

SUMMARY

The implementation of binary transversal filters (b.t.f.) is gaining wide acceptance, due to their operational adaptability to variable bit rates, their achievable high out-of-band attenuation and linear phase performance and to the recent availability of medium-scale integrated components such as charged-coupled devices.

The design and evaluation results of a b.t.f. meeting the single channel per carrier *Intelsat* satellite communications specifications are presented, which are based on Mueller's synthesis method. The performance evaluation problems of these type of filters are also described and, in particular, emphasis is given to the following: out-of-band to in-band power ratio inaccuracies when wideband measurement techniques are employed, and the evaluation of the r.m.s. jitter of b.t.f. based on the jitter probability density function.

1 Introduction

In data communications it is essential to minimize the co-channel, the adjacent channel and the intersymbol interference (i.s.i.). A major contribution to these signal impairments is the 'out-of-band' energy of the digital signal. 'Out-of-band' is defined to be the band of frequencies above or below a specified signal bandwidth, which in turn is given in terms of the theoretical minimum Nyquist bandwidth.

In order to conserve radio bandwidth, it is essential to approach the Nyquist channel. With analogue passive or active filter design techniques, it is hard to achieve a minimum bandwidth channel which has low intersymbol interference. High-order filters with a large number of analogue group delay equalizer sections are required; eleventh-order filters have been reported in the literature.¹ Another disadvantage of analogue filters is that they are not easily adaptable for variable data rates. This is a drawback in digital systems if a flexibility in operating conditions is required.

A class of transversal filters called Binary Transversal Filters (b.t.f.) is studied. It is easy to implement these filters with integrated circuitry and with weighting resistors. For this reason they are easily adaptable to medium and large scale integration. The performance evaluation problems of these type of filters and the introduced measurement inaccuracies are discussed.

2 Binary Transversal Filters

The basic structure of a transversal filter, also known as a non-recursive digital filter, is shown in Fig. 1. It consists of delay elements D , multipliers c_i and an adder Σ .

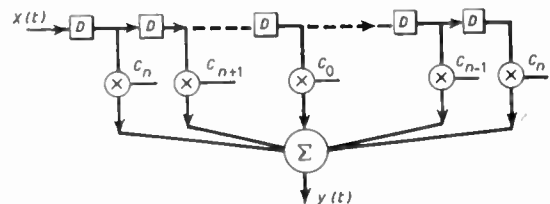


Fig. 1. Transversal filter structure.

The frequency response of this filter is given by:

$$H(\omega) = \exp(-j\omega a) \sum_{i=-n}^{+n} c_i \exp(-j\omega D) \quad (1 a)$$

or expressed in a vector form as:

$$H(\omega) = \mathbf{c}^T \mathbf{x} \quad (1 b)$$

where

$$\mathbf{c}^T = (c_{-n}, \dots, c_0, \dots, c_n) \quad (1 c)$$

and

$$\mathbf{x}^T = (x_{-n}, \dots, x_0, \dots, x_n), \quad \text{with} \quad x_i = \exp(-j_i \omega D) \exp(-j\omega a) \quad (1 d)$$

* Department of Electrical Engineering, Concordia University, 1455 de Maisonneuve Boulevard West, Montreal, Quebec H3G 1M8, Canada.

† Bell Company of Canada Ltd., Montreal, Quebec, Canada.

in equation (1 a) the multipliers c_i are the coefficients of the complex Fourier series expansion of the $H(\omega)$ function [Ref. 2, chap. 5]. The term $\exp(-j\omega a)$ represents a constant delay which can, in most cases, be compensated for at the receiver end. Different approaches have been used for the synthesis of transversal filters. The goal of all of these techniques has been to approximate the transfer function $H(\omega)$. Least squares technique,³ Chebyshev approximation⁴ or simple Fourier series expansion,⁵ were among the techniques frequently employed.

If we recall that in data transmission filtering is applied frequently only to binary waveforms, that is, ones and zeros, and that the intersymbol interference has to be minimized, then it becomes evident that the above mentioned techniques are far too complex.

A technique that takes into consideration these specifications and at the same time expresses the spectrum roll-off in terms of the excess Nyquist bandwidth has been proposed by Mueller.⁶ Using Mueller's technique of time domain filtering, the transversal filter coefficients c_i are optimized in order to maximize the transmitted energy in a specified bandwidth. This bandwidth is specified as an excess Nyquist bandwidth.

Using shift registers as delay elements (DS_i), i.e. 'Delay Shifts', resistors for multipliers (c_i) and an operational amplifier as a summing network, it is possible to implement a binary transversal filter (Fig. 2) which responds to only binary pulses, and thus is applicable for digital transmission.

We will assume the input, $x(t)$, is a pulse sequence and is of the return-to-zero type (r.z.)

$$x(t) = \sum_{i=0}^{\infty} a_i p_i(t) \tag{2}$$

where

$$p_i(t) = 1 \text{ for } (i-1)D < t \leq iD$$

$$p_i(t) = 0 \text{ elsewhere}$$

and

$$a_i = 0 \text{ or } 1$$

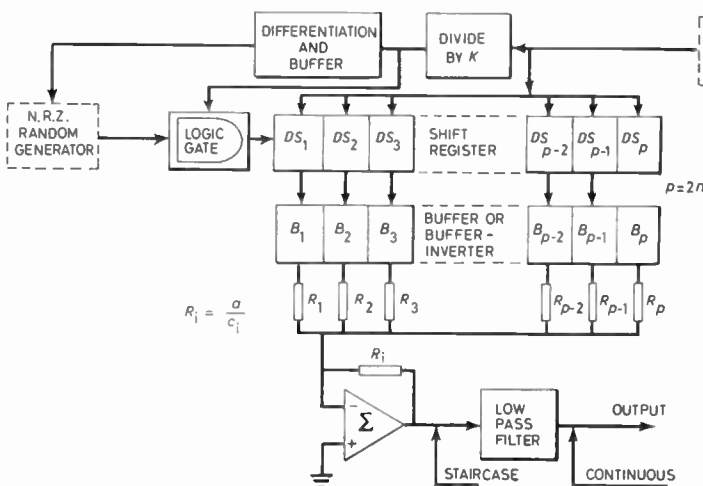


Fig. 2. Binary transversal filter—general block diagram.

with $\text{prob}(0) = \text{prob}(1) = \frac{1}{2}$.

It is noted that this definition of the r.z. pulse implies a 100% duration cycle. The time domain response $y(t)$ of a Nyquist type b.t.f. excited by a single pulse representing a binary 'one' is shown in Fig. 3. If we define T as the unit baud interval, then the number of samples per baud, K , is given by:

$$K = T/D, \quad K = 1, 2, 3, \dots \tag{3}$$

Since K is an integer it follows from (3) that D has to be an integer submultiple of T .

To assure negligible i.s.i. degradation, the Nyquist criterion requires that $y(t)$ crosses the zero axis at regularly spaced intervals T , except for the central peak (Fig. 3). From these requirements the following equations were established:⁶

$$c_0 = \max \{c_i\} \tag{4 a}$$

$$c_{-i} = c_i \tag{4 b}$$

$$c_{NK} = 0, \text{ for } N = \pm 1, \pm 2, \dots, \pm L \tag{4 c}$$

$$c_i = 0, \text{ for } |i| \geq KL \tag{4 d}$$

Note that the impulse response $h(t)$ of the b.t.f. is limited to L number of lobes (eqn. (4 c)), and that phase linearity is guaranteed by equation (4 b). The symbol L in equations (4 c) and (4 d) represents the number of lobes in the impulse response, and for convenience the sampled impulse response has been truncated at an integer number of lobes (eqn. (4 d)). To obtain the optimum b.t.f. coefficients which maximize the transmitted energy in the excess Nyquist bandwidth, a program has been written.⁷

The block diagram of a binary transversal filter is shown in Fig. 2. An external clock (dashed lines) generates sharp pulses. These pulses are K times faster than the corresponding bit (data) rate ($1/T$). This clock is used for two purposes:

- (1) The shift towards right the content of the shift register DS_i . At each clock pulse the content of DS_i is transferred to DS_{i+1} .

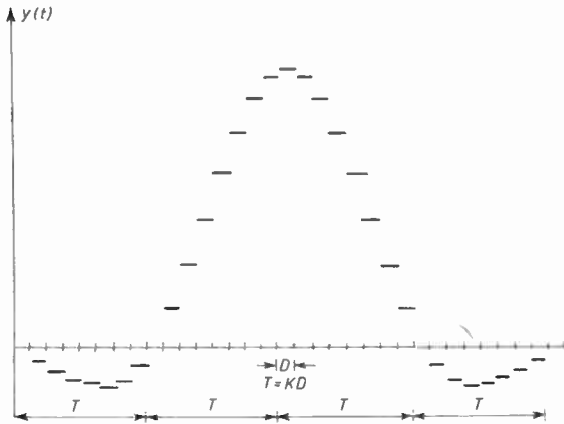


Fig. 3. Transversal filter impulse response.

(2) To drive the 'divider by K' module which in turn generates pulses having a width of D units of time.

The next stage is a differentiator and a buffer. The D width pulses are converted into sharp positive-going impulses. These bit rate spikes synchronize the external random data generator (dashed line) whose output is of the non-return-to-zero (n.r.z.) format. An AND logic gate combines these n.r.z. pulses with pulses having a width D arriving from the 'divider by K' module. This leads to r.z. type pulses. If a logic one is present at the output of the random generator and a pulse of width D at the output of the divider then a pulse $p(t)$ will be present at the input of the shift register. Depending on the sign of the coefficients c_i , a buffer or a buffer inverter is used at the output of the corresponding flip-flop. These buffers are required to avoid the voltage drop due to the loading resistors R_i . At the output of each buffer/inverter a resistor

$$R_i = a/c_i \quad (5)$$

is employed to approximate a current proportional to c_i . These resistors are the physical realization of multiplicative constants. All resistors are connected to the input of an operational amplifier used in its inverting configuration. An inverted staircase approximation of the Nyquist-type pulses is present at this point. Afterwards, an active low-pass filter follows which smoothes and inverts again the staircase response of the b.t.f. The final output of the filter is a continuous wave truncated to $\pm L$ lobes (see eqn. (4 d)).

3 B.T.F. Design and Evaluation

The described theory and design techniques of the previous Sections will now be applied to satellite communications. In particular we will focus attention on the design of a premodulation filter which is employed in the q.p.s.k. modulator of the single channel per carrier (s.c.p.c.) p.c.m. multiple access demand assignment equipment (SPADE). SPADE is gaining a wide popularity and as it was shown most recently by M. Asahara *et al.*,⁸

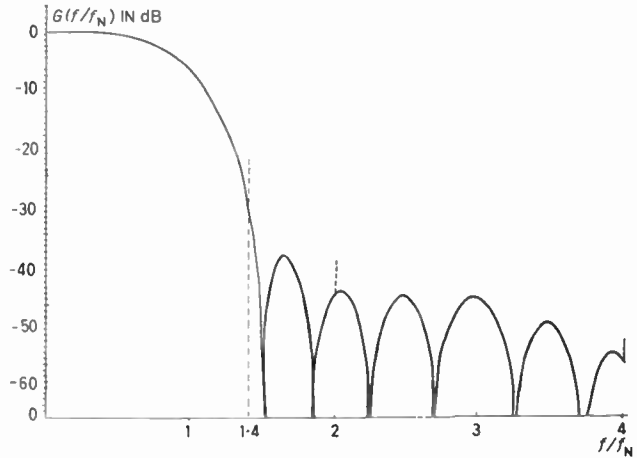


Fig. 4. B.t.f. power spectrum density.

there is an intensive ongoing research in this field. De Cristofaro⁷ points that his research results are very similar to the recent results of Asahara. The specifications for the 64 kbit/s s.c.p.c. premodulation filter are: (1) data rates 32 kbaud; (2) the ratio of the in-band to out-of-band energy will be greater than 26 dB, where all frequencies higher than 22.5 kHz are considered out of band. This corresponds to 1.406 times the Nyquist frequency.

Using the computer program developed by the authors it was found that for 40.6% excess Nyquist the 26 dB energy ratio could be marginally met using 8 samples per band interval and 2 lobes in the impulse response (4 c-d). Since the first and the last b.t.f.s coefficients are zero (i.s.i. constraint) we considered only the last one and the b.t.f. was implemented with a 32-stage shift register. To achieve a shift register of this length we used 4 c.o.s.m.o.s. integrated circuits having 8 stages each.

A sensitivity study of the filter transfer function $H(\omega)$ with respect to the c_i coefficients variations showed that the maximum allowable tolerance of the coefficients c_i is $\pm 2.5\%$. The details are presented in Section 6.

A computer simulation of the filter output power spectrum density for random input data is shown in

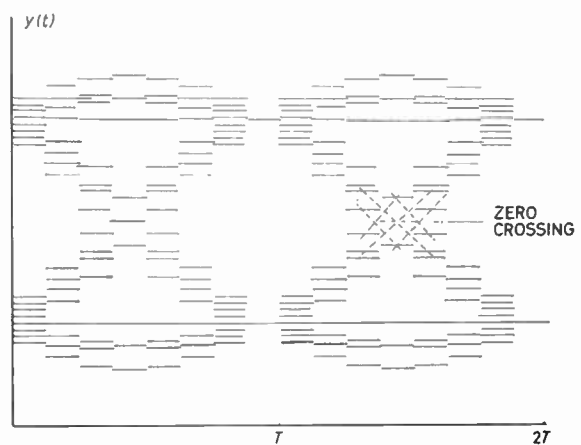


Fig. 5. B.t.f. eye diagram simulation ($K = 8, L = 2, \alpha = 0.4$).

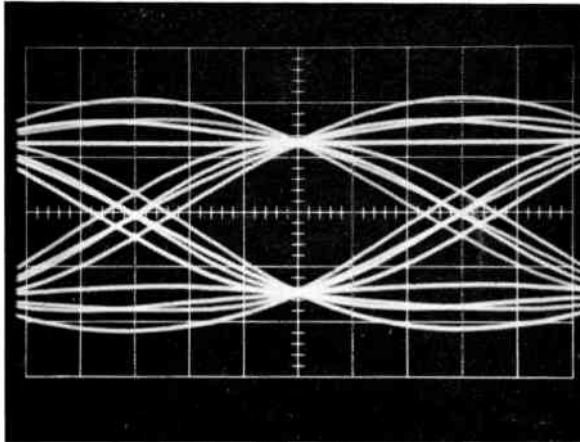


Fig. 6. Eye diagram of the b.t.f. followed by a low-pass filter. Vertical: 0.1 volts/div. Horizontal: 5 μs/div.

Fig. 4. In this simulation it was found that if a second order low-pass filter is employed after the b.t.f. then an in-band to out-of-band energy rejection ratio of greater than 35 dB is achieved.

It was shown by simulation that the b.t.f. has a negligible i.s.i. contribution. The computer simulation result is shown in Fig. 5 and the practical laboratory measurements on the designed hardware are shown in Fig. 6. The measured output power spectrum density of this filter for a very long pseudo-random input is shown in Fig. 7. As expected, the output power spectral density has a much faster spectral roll-off than a conventional unfiltered random n.r.z. data stream. The discrete tones at the bit rate multiples are due to the r.z. signal processing step occurring within the filter which was described in Section 2.

Finally, the in-band to out-of-band energy ratio was measured to be equal to 32 dB. It is evident from the measured results that the combined b.t.f. and the second-order l.p.f. shown in Fig. 8 satisfy the SPADE requirements.

4 Introduced Measurement Inaccuracies

This Section describes introduced measurement inaccuracies when wideband measurement techniques are

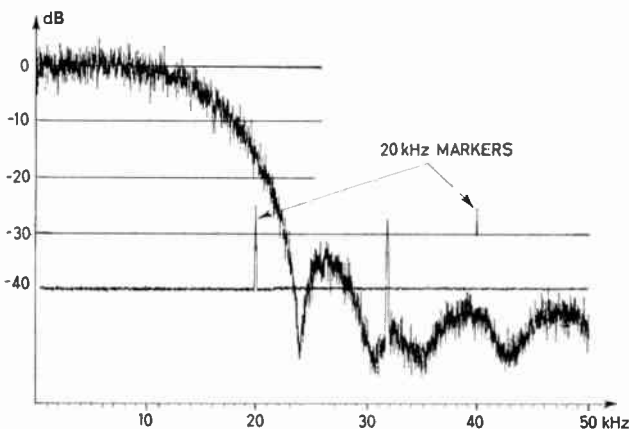


Fig. 7. Measured power spectrum density of b.t.f. (α=0.41).

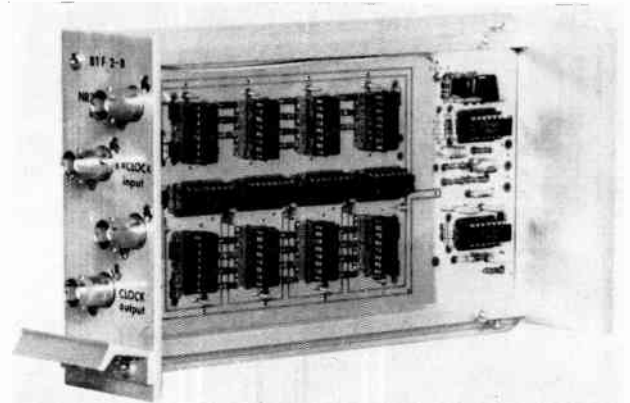


Fig. 8. Hybrid binary transversal filter.

used to find the out-of-band to in-band power ratio (Fig. 9). We recall that ‘out-of-band’ is defined to be the band of frequencies above or below a specified signal bandwidth, which in turn is given in terms of the theoretical minimum Nyquist bandwidth.

The power spectrum density $G_D(f)$ at the output of the digital b.t.f. is given by:⁹

$$G_D(f) = f_N |Y(f)|^2 + 2f_N^2 \sum_{m=1}^{\infty} |Y(2mf_N)|^2 \delta(f - 2mf_N) \quad (6)$$

where the first term represents a continuous function and the second term represents the power concentrated at the harmonics of $(2f_N)$, $Y(f)$ is the Fourier transform of $y(t)$, the signal representing binary one, the signal representing binary zero is represented by absence of information, binary ones and binary zeros are assumed to have equal probability of being transmitted, finally f_N is the Nyquist frequency.

$$Y(f) = D \text{ sinc}(\pi f D) \times \exp(-j2\pi f a) \sum_{i=-n}^n c_i \exp(-j2\pi f i D) \quad (7)$$

$$T = \frac{1}{f_s} = \frac{1}{2f_N} = KD \quad D = (2f_N K)^{-1} \quad (8)$$

where T is the bit interval, f_s is the bit rate, K is the number of samples per baud interval and L is the number of lobes in the impulse response. Replacing D by $(2f_N K)^{-1}$, a by nD , n by KL in equation (7) and recalling that in our application, the b.t.f. has linear phase, guaranteed by the even symmetry of the coefficients c_i , i.e. $c_i = c_{-i}$ equation (7) becomes:

$$Y(f) = \frac{1}{2Kf_N} \text{ sinc}\left(\frac{\pi f}{2Kf_N}\right) \times \exp\left(-j\frac{\pi f L}{f_N}\right) \left\{ c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i f}{f_N K}\right) \right\} \quad (9)$$

Now substituting this last equation into (6) and setting $f = x f_N$, we obtain:

$$G_D(x) = \frac{1}{4K^2 f_N} \times \left[\text{sinc}^2\left(\frac{\pi x}{2K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi x i}{K}\right) \right|^2 + \frac{1}{2K^2} \sum_{m=1}^{\infty} \left\{ \text{sinc}^2\left(\frac{\pi m}{K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{2\pi i m}{K}\right) \right|^2 \right\} \delta(x-2m) \right] \quad (10)$$

To attenuate further the higher-order components of the digital b.t.f. spectrum, thus achieving a higher in-band to out-of-band power ratio, the digital b.t.f. is followed by a second-order low pass filter having a Butterworth response. The cut-off frequency of the filter required to minimize the i.s.i. of the hybrid network was determined experimentally to be approximately $2.8 f_N$ (45 kHz), where f_N is the Nyquist frequency (16 kHz). It is

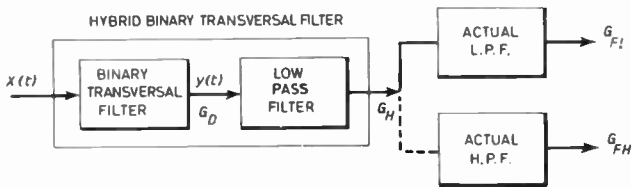


Fig. 9. Set-up for measuring out-of-band to in-band power ratio.

important to notice that a hybrid network will limit the original digital b.t.f.s self-adaptability to different bit rates, since in our case the b.t.f. is designed for a specific bit rate, adaptability is not an important factor. The power spectrum density at the output of the hybrid b.t.f. is given by:

$$G_H(x) = \frac{1}{4K^2 f_N} \text{sinc}^2\left(\frac{\pi x}{2K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i x}{K}\right) \right|^2 \left\{ 1 + \left(\frac{x}{2.8}\right)^4 \right\}^{-1} + \frac{1}{2K^2} \sum_{m=1}^{\infty} \text{sinc}^2\left(\frac{\pi m}{K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i 2m}{K}\right) \right|^2 \left\{ 1 + \left(\frac{m}{1.4}\right)^4 \right\}^{-1} \delta(x-2m) \quad (11)$$

If α is the excess Nyquist bandwidth then the excess Nyquist frequency is $f_N(1+\alpha)$. A comparison of the in-band power (i.b.p.) and out-of-band power (o.b.p.) using ideal infinite slope and realizable finite slope filters follows:

$$(i.b.p.)_I = \int_0^{1+\alpha} G_H(x) dx \quad (12)$$

$$(o.b.p.)_I = \int_{1+\alpha}^{\beta} G_H(x) dx \quad (13)$$

In the computer simulation and in measurements the upper ∞ bound is replaced by β where $\beta \geq 20$, this implies that in equation (11) m has the values $1, 2, \dots, \beta/2$. If the h.b.t.f. is followed by a realizable finite slope l.p.f. to measure the in-band power and since the available filter is of the N th order Butterworth type, the power spectrum density at the output of this filter is given by:

$$G_{FL}(x) = \frac{1}{4K^2 f_N} \text{sinc}^2\left(\frac{\pi x}{2K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i x}{K}\right) \right|^2 \left\{ 1 + \left(\frac{x}{2.8}\right)^4 \right\}^{-1} \times \left\{ 1 + \left(\frac{x}{1+\alpha}\right)^{2N} \right\}^{-1} + \frac{1}{2K^2} \sum_{m=1}^{\infty} \text{sinc}^2\left(\frac{\pi m}{K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i 2m}{K}\right) \right|^2 \left\{ 1 + \left(\frac{m}{1.4}\right)^4 \right\}^{-1} \times \left\{ 1 + \left(\frac{2m}{1+\alpha}\right)^{2N} \right\}^{-1} \delta(x-2m) \quad (14)$$

If we replace the realizable l.p.f. by an N th order Butterworth h.p.f. in order to measure the out-of-band wide-band power equation (14) becomes:

$$G_{FH}(x) = \frac{1}{4K^2 f_N} \text{sinc}^2\left(\frac{\pi x}{2K}\right) \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i x}{K}\right) \right|^2 \times \left\{ 1 + \left(\frac{x}{2.8}\right)^4 \right\}^{-1} \left\{ 1 + \left(\frac{1+\alpha}{x}\right)^{2N} \right\}^{-1} + \frac{1}{2K^2} \sum_{m=1}^{\infty} \text{sinc}^2\left(\frac{\pi m}{K}\right) \times \left| c_0 + 2 \sum_{i=1}^n c_i \cos\left(\frac{\pi i 2m}{K}\right) \right|^2 \left\{ 1 + \left(\frac{m}{1.4}\right)^4 \right\}^{-1} \left\{ 1 + \left(\frac{1+\alpha}{2m}\right)^{2N} \right\}^{-1} \delta(x-2m) \quad (15)$$

The in-band correction factor (i.b.c.f.) and the out-of-band correction factor (o.b.c.f.) are given by:

$$i.b.c.f. = \int_0^{1+\alpha} G_H(x) dx - \int_0^{\beta} G_{FL}(x) dx \quad (16)$$

and

$$o.b.c.f. = \int_{1+\alpha}^{\beta} G_H(x) dx - \int_0^{\beta} G_{FH}(x) dx \quad (17)$$

Adding the i.b.c.f. to the measured in-band power and the o.b.c.f. to the measured out-of-band power we are able to predict the in-band to out-of-band power ratio with a considerably increased accuracy. With the above correc-

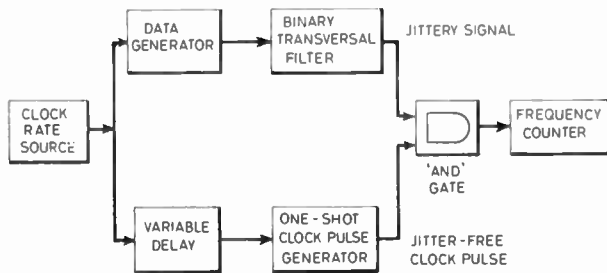


Fig. 10. Jitter evaluation set-up.

tion factors it is possible to predict wideband continuous power ratios if practical finite slope filters are used.

5 Root Mean Square Jitter Evaluation of the B.T.F.

In this Section the r.m.s. jitters of b.t.f.s based on the jitter probability density function measurements is discussed. The set-up shown in Fig. 10 is used to detect and count the number of zero crossings of the jittery signal for different positions of the jitter-free clock pulse (Fig. 11).

In Fig. 11 when the sampling clock pulse is at a position x_j less than x_0 , then there is no simultaneous high logic input to the AND gate (Fig. 10). Thus for this delay setting there is no response on the frequency counter. On the other hand, when the sampling is set to x_j greater than x_1 , then the sampling clock and the jittery signal act simultaneously. Thus for this delay setting the frequency counter begins to count all clock pulses, i.e., the pulse rate.

The peak-to-peak jitter is equal to $(x_1 - x_0)$. This corresponds to the time delay difference of the two time instants in which the frequency counter begins to count and when it begins to count all clock pulses that is the pulse rate.

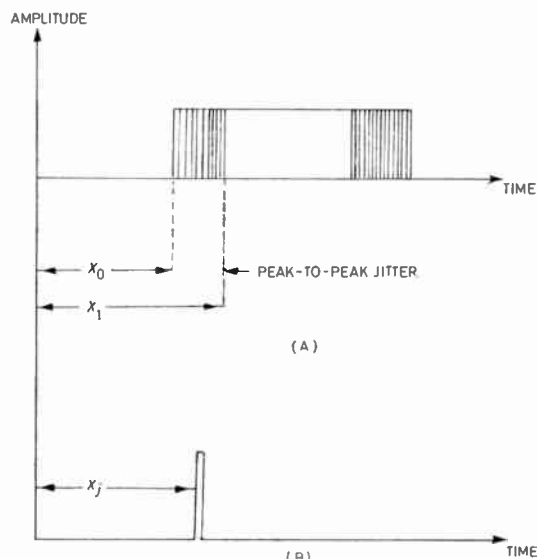


Fig. 11. Illustration of the jittery and of the reference signals. (a) Jittery signal; (b) Jitter-free clock pulse.

Dividing the accumulated number of zero crossings by the clock rate we obtain $F(x_j)$ the jitter probability distribution function. The jitter probability density function $f(x_j)$ (Fig. 12) is obtained from $F(x_j)$ by:

$$f(x_j) = \frac{F(x_{j+1}) - F(x_j)}{x_{j+1} - x_j} \quad (18)$$

In this equation x_j is the time instant of the corresponding zero crossing. The r.m.s. jitter J_{RMS} is calculated using equation (19):

$$J_{RMS} = \sqrt{\sum_j (x_j - \bar{x})^2 f(x_j)} \quad (19)$$

where \bar{x} in equation (19) is given by:

$$\bar{x} = \sum_j x_j f(x_j) \quad (20)$$

In the above method it was shown how to measure and compute with the aid of an improvised laboratory set-up the peak to peak jitter, probability distribution and density functions of the jittery signal and the r.m.s. jitter. This method applies to a large class of randomly varying signals.

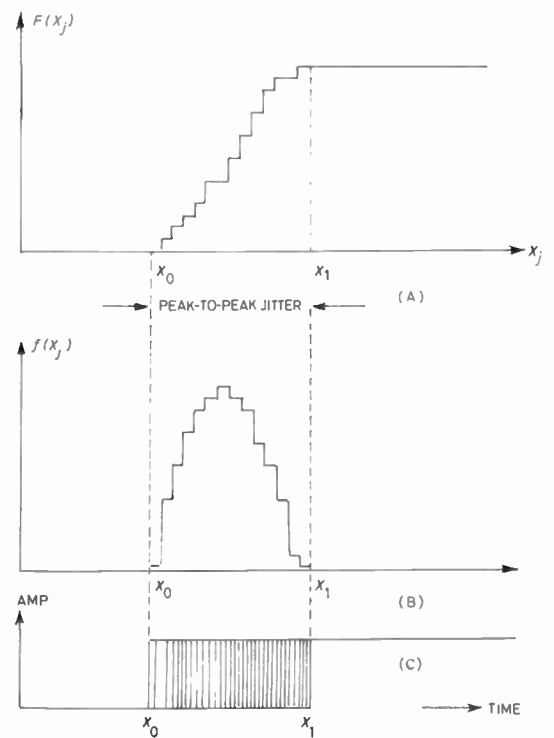


Fig. 12. Jitter distribution and density functions. (a) Jitter probability distribution function; (b) Jitter probability density function; (c) Jittery signal.

6 Sensitivity of the B.T.F. Energy Ratio to Changes in the Coefficients

The objective of this Section is to evaluate the degradation in the energy ratio performance considering real coefficients, i.e. coefficients which are accurate only within some specified limits.

The problem was approached using a statistical method which involved computing the energy ratio of filter coefficients which have been perturbed at random. It is evident that to obtain meaningful results, the perturbation has to be within a specified range. This process is repeated for a large number of times (e.g. 100) until enough samples of the energy ratio are obtained. The mean and the standard deviation of 50 and 100 samples were computed, and are given in Tables 1 and 2. Since the results were very similar, we have concluded that one hundred samples were sufficient for statistical purposes.

The out-of-band to in-band energy ratio for a transversal filter is given by:

$$\frac{E(h)}{E(\omega_0)} = \frac{E(\infty)}{E(\omega_0)} - 1 \quad (21)$$

where $E(\infty)$ is the total output energy and $E(\omega_0)$ is the energy below and $E(h)$ the energy above the specified excess Nyquist frequency, where $\omega_0 = 2\pi\{(1 + \alpha)f_N\}$.

To calculate the total energy, we use equation (22):

$$E(\infty) = \hat{c}^T \hat{c} T / K \quad (22)$$

where (T/K) is the normalized energy of a single pulse. Now expressing T , the baud interval in terms of the Nyquist frequency f_N , we have:

$$E(\infty) = \hat{c}^T \hat{c} (2f_N K)^{-1} \quad (23)$$

To evaluate the energy below the specified excess

Nyquist frequency, we have:

$$E(\omega_0) = \frac{1}{\pi} \int_0^{\omega_0} |W(\omega)|^2 |H(\omega)|^2 d\omega \quad (24)$$

The filter transfer function $H(\omega)$ is then replaced by $H(\omega) = \hat{c}^T \hat{x}$ and, the weighting function $W(\omega)$ by $D \text{sinc}(\omega D/2)$, allowing the evaluation of the energy using numerical integration.

From equation (23), we can deduce that the in-band energy depends not only on $H(\omega)$ but also on $W(\omega)$. Although the sensitivity study here concentrated exclusively on $H(\omega)$, it is important to note that perturbation or imperfections of the weighting function $W(\omega)$ will also degrade the energy ratio performance. Further studies should be conducted to evaluate if these imperfections in $W(\omega)$ are a dominant factor. Some possible examples are: pulse leading and trailing edges, top tilt and/or overshoots.

The coefficients c_i of the binary transversal filter were perturbed using the following relation:

$$\tilde{c}_i = c_i \{1 + \text{RANDU} \times \text{PER}\} \quad (25)$$

where \tilde{c}_i are the perturbed coefficients, RANDU is a random variable uniformly distributed between -1 and $+1$ and PER is the specified coefficients accuracy.

Assuming the coefficients c_i are statistically independent, it can be shown that equation (24) accurately models the physical realization of the b.t.f.

Table 1. Expected value of the b.t.f. energy ratio for $\pm 1\%$ coefficients accuracy.

B.t.f. characteristics			Optimum coefficients		$\pm 1\%$ coefficients accuracy						Mean energy ratio (dB) degrad.
					50 samples			100 samples			
					$E(h)/E(\omega_0)$			$E(h)/E(\omega_0)$			
L	K	Excess Nyquist	$\frac{E(\omega_0)}{E(\infty)}$	$\frac{E(h)}{E(\omega_0)}$ dB	Mean	Standard deviation	dB mean	Mean	Standard deviation	dB mean	
1	8	30%	0.9918395	-20.85	0.0082554	0.0000106	-20.83	0.0082548	0.0000097	-20.83	0.02
1	8	41%	0.99402229	-22.21	0.0060405	0.0000109	-22.19	0.0060400	0.0000099	-22.19	0.02
2	8	30%	0.99588559	-23.84	0.0041588	0.0000096	-23.81	0.00415877	0.0000092	-23.81	0.03
2	8	40%	0.99624374	-24.24	0.0037974	0.0000097	-24.21	0.0037975	0.0000093	-24.21	0.03

Table 2. Expected value of the b.t.f. energy ratio for $\pm 5\%$ coefficients accuracy.

B.t.f. characteristics			Optimum coefficients		$\pm 5\%$ coefficients accuracy						Aver. energy ratio (dB) degrad.
					50 samples			100 samples			
					$E(h)/E(\omega_0)$			$E(h)/E(\omega_0)$			
L	K	Excess Nyquist	$\frac{E(\omega_0)}{E(\infty)}$	$\frac{E(h)}{E(\omega_0)}$ dB	Mean	Standard deviation	dB mean	Mean	Standard deviation	dB mean	
1	8	30%	0.99183905	-20.85	0.0089115	0.0002663	-20.50	0.0089978	0.0002434	-20.51	0.34
1	8	41%	0.99402229	-22.21	0.0066849	0.0002721	-21.75	0.0066719	0.0002482	-21.76	0.45
2	8	30%	0.99488559	-23.84	0.0048164	0.0002400	-23.17	0.0048149	0.0002308	-23.17	0.67
2	8	40%	0.99624374	-24.24	0.0044449	0.0002428	-23.52	0.0044456	0.0002337	-23.52	0.72

In our study, values of 1%, 5%, 10% were given to PER.

The mean value (\bar{X}) and the standard derivation (σ) of the energy ratio were computed using the following relations:

$$\bar{X} = \frac{1}{N} \sum_{i=1}^N X_i \quad (26)$$

$$\sigma = \sqrt{\left[\frac{\sum X_i^2 - (\sum X_i)^2 / N}{N-1} \right]} \quad (27)$$

where N is the total number of samples and X_i the sample value.

A computer program called SENST was implemented in order to evaluate the energy ratio and its statistics. The mean value of the energy ratio was expressed in dB. In order to check the accuracy of the program SENST in calculating the energy ratio the value of λ or $E(\omega_0)/E(\infty)$ was computed initially using non-perturbed optimum coefficients. The value of λ (see Tables 1 and 2) was then compared with the maximum eigenvalue computed using another program called BFT1 and no discrepancies were found for up to eight significant digits, proving the validity of the program SENST.

The out-of-band to in-band energy ratio $[E(h)/E(\omega_0)]$ was then evaluated for different types of filters. Several of these filters were implemented.

From Tables 1 and 2, the first conclusion that can be drawn is that, contrary to what were our expectations before this study, the binary transversal filter's coefficients do not have to be extremely accurate to keep the energy ratio within practical limits. Tables 1 and 2 also show that the standard deviations are fairly narrow indicating uncomplicated filter repeatability.

In order to force the eye diagram of a b.t.f. to be maximally open at its centre, an empirical rule should be followed, requiring that the coefficients' accuracy be at least one half the minimum difference existing between two adjacent coefficients. In this manner, two steps of the eye (see Fig. 5) can in the worst case be of equal amplitude without ever exceeding each other.

Observing this rule, we arrive at the conclusion that the coefficients accuracy should be kept below the $\pm 2.5\%$ value.

7 Conclusion

Based on Mueller's theory⁶ a c.o.s.m.o.s. binary transversal filter meeting s.c.p.c. transmit filter requirements was designed. In this design the tap multipliers were chosen so as to minimize intersymbol interference and maximize the in-band to out-of-band power ratio. The introduced measurement inaccuracies of the out-of-band to in-band power ratio when using wideband measurement techniques is described. It was shown that with the in-band and out-of-band correction factors it is possible to accurately predict the filter power ratio making usage of available finite slope filters. The r.m.s. jitter evaluation and finally a sensitivity analysis of the multipliers is presented. Theoretical studies and experimental results have demonstrated that the coefficient accuracy is not, as was supposed, a major controlling factor for the power ratio performance.

8 References

- 1 Spaulding, D. A., 'Synthesis of pulse-shaping networks in the time domain', *Bell Syst. Tech. J.*, **48**, pp. 2425-44, September 1969.
- 2 Huelsman, L. P., 'Active Filters, Lumped, Distributed, Integrated, Digital and Parametric' (McGraw-Hill, New York, 1970).
- 3 Vincent-Carrefour, A., Boeri, F., Dubus, F. and Esteban, D., 'Synthesis of a transversal digital filter defined in the complex plane', *Ann. Telecommun. (France)*, **28**, pp. 297-303, July-August 1973.
- 4 Parks, T. W. and McClellan, J. H., 'Chebyshev approximation for non-recursive digital filters with linear phase', *IEEE Trans. on Circuit Theory*, CT-19, No. 2, pp. 189-94, March 1972.
- 5 Danilov, B. S., 'Binary shift registers for generating a given signal spectrum in data transmission equipment', *Telecommun. Radio Engng (USSR)*, **23**, No. 2, pp. 26-31, 1969 (English translation).
- 6 Mueller, K. H., 'A new approach to optimum pulse shaping in sampled systems using time-domain filtering', *Bell Syst. Tech. J.*, **52**, No. 5, pp. 723-9, May-June 1973.
- 7 De Cristofaro, R., 'Transversal filter design and application in satellite communication', M.Eng. Thesis, Concordia University, Montreal, April 1976.
- 8 Asahara, M., Toyonaga, N., Sasaki, S. and Miyo, T., 'A 4-phase psk modem for an scpc satellite communications system', International Conference on Communications, ICC-75, San Francisco, (Vol. 1), 16th-18th June 1975.
- 9 Bennett, W. R. and Davey, Y. R., 'Data Transmission', chap. 19 (McGraw-Hill, New York, 1965).

Manuscript first received by the Institution on 4th August 1976 and in final form on 21st December 1976. (Paper No. 1767/Com. 150.)

© The Institution of Electronic and Radio Engineers, 1977

A novel least-square Fourier algorithm for decomposition of discrete, non-equidistant acquisition data

M. BOUTHÉON, Dr.Ing., A.M.Ing.,*

V. CHOCHAN, B.Sc., Ph.D.*

and

J. P. POTIER, M.Sc., Doct.*

Based on a paper presented at the Digital Equipment Computer User Society's European Symposium, held in Munich in September 1976.

SUMMARY

A novel procedure for evaluating directly the Fourier series coefficients of a function described by unequally spaced but symmetrically disposed interval discrete points is given and an example illustrated. The procedure's simplicity enables it to be used for harmonic analyses of non-equidistant interval data without using the intermediate curve-fitting techniques.

* PS Division, CERN, 1211 Geneva 23, Switzerland

1 Introduction

The representation of a periodic signal or a function often takes the form of a series of discrete samples with the sampling carried out according to the Nyquist criterion and the sample points spaced apart at equal intervals. The Fourier decomposition of such a function is readily carried out by the direct application of derived expressions given in the references.^{1,2} However, in instances where the sampling is not carried out at equal intervals or where the value at a sample point is totally erroneous (and hence, has to be rejected) or is unavailable due to faulty acquisition electronics, the solution is no longer a straightforward application of such expressions. One form of solution would be to fit a continuous curve to the available non-equidistant interval data and then perform the equidistant interval sampling and analysis on this curve according to the known expressions. For a remote acquisition, analysis and control problem as in the CERN Proton Synchrotron (PS), this would imply a long, time-consuming process and hence an alternative solution based on a simple computer implementation had to be formulated.

In this contribution, a versatile least-square approximation algorithm is developed to solve this problem and a typical application is illustrated. The method shown has been successfully applied to the harmonic analysis of the closed orbit of the proton beam in the CERN PS and the subsequent orbit minimization by applying the necessary harmonic-related currents in the correction magnets. By using a simple technique based on the odd and even separation of the function, the method circumvents the problem of insoluble equations yielded by the least-square approximation of the complete function.

2 The Algorithm

Let the unequally spaced M points along the abscissa be represented by $\theta_1, \theta_2, \dots, \theta_M$, where θ is within the range $-\pi < \theta < +\pi$ and let the value of the discrete data at a point θ_i be represented by $f(\theta_i)$.

We postulate that $f(\theta_i)$ is periodic and therefore may be approximated over all i by a n -term Fourier series ($n < M$) of the form:

$$\left\{ F(\theta_i) = \sum_{j=0}^n A_j \cos j\theta_i + \sum_{j=1}^n B_j \sin j\theta_i \right\} \quad (1)$$

A direct least-square approximation of $f(\theta_i)$ by $F(\theta_i)$ yields equations in which the summation in the sin-cos cross-product terms over the complete period is non-zero due to the unequal interval spacing and hence, they may not be readily solved for the unknown coefficients by a simple matrix inversion. However, it is permissible to separate $f(\theta_i)$ into its odd and even parts, i.e.:

$$g(\theta_i) = \frac{1}{2}\{f(\theta_i) + f(-\theta_i)\} \dots \text{even} \quad (2)$$

$$h(\theta_i) = \frac{1}{2}\{f(\theta_i) - f(-\theta_i)\} \dots \text{odd} \quad (3)$$

By applying the summed least-square error principle

over all M points separately for both odd and even functions approximated respectively by sine and cosine terms, a simple solution results.

For the even function—The error η is given by

$$\eta = \sum_{i=1}^M \left[g(\theta_i) - \sum_{j=0}^n A_j \cos j\theta_i \right]^2 \quad (4)$$

Differentiating with respect to coefficient A_k where $0 \leq k \leq n$,

$$\frac{d\eta}{dA_k} = -2 \sum_{i=1}^M \left[g(\theta_i) - \sum_{j=0}^n A_j \cos j\theta_i \right] \cos k\theta_i \quad (5)$$

For a minimum summed-square error, eqn. (5) equates to zero, therefore:

$$\sum_{i=1}^M g(\theta_i) \cos k\theta_i = \sum_{i=1}^M \sum_{j=0}^n A_j \cos j\theta_i \cos k\theta_i \quad (6)$$

Equation (6) yields $(n+1)$ equations for values of k between 0 and n for all summations over the discrete points 1 to M . The problem therefore reduces to finding the $(n+1)$ coefficients A_k from these equations.

Letting

$$G_k = \sum_{i=1}^M g(\theta_i) \cos k\theta_i \quad (7)$$

$$C'_{kj} = \sum_{i=1}^M \cos k\theta_i \cos j\theta_i \quad (8)$$

Therefore

$$G_k = \sum_{j=0}^n C'_{kj} A_j \quad (9)$$

i.e.

$$[G] = [C'] [A] \quad (10)$$

where $[G]$ and $[A]$ are $(n+1) \times 1$ matrices, while $[C']$ is a $(n+1) \times (n+1)$ matrix. Inverting the cosine product matrix $[C']$ would then yield the coefficients directly, i.e.

$$[A] = [C']^{-1} [G] \quad (11)$$

Letting

$$[C']^{-1} = [C] \quad (12)$$

gives

$$[A] = [C] [G]. \quad (13)$$

Hence a matrix inversion of the cosine product matrix whose elements are described by eqn. (8) yields the solution given by eqn. (13). An interesting aspect of this solution is that the cosine product matrix is entirely independent of the value of the function and is solely dependent on the unequally spaced points θ_i along the abscissa (e.g. the beam position monitors in the CERN PS ring). Hence this matrix evaluation and inversion may be carried out separately depending only on the available data sampling

points along the abscissa. Similarly, the matrix G is readily computed once the odd and even separation is made from eqns. (2) and (3).

For the odd function—Applying the same procedure as in (9) above yields the sine coefficients B_j from

$$[B] = [S] [H] \quad (14)$$

where $[H]$ and $[B]$ are $n \times 1$ matrices and $[S]$ an $n \times n$ matrix with

$$[S] = [S']^{-1} \quad (15)$$

and

$$H_k = \sum_{i=1}^M h(\theta_i) \sin k\theta_i \quad (16)$$

$$S'_{kj} = \sum_{i=1}^M \sin k\theta_i \sin j\theta_i \quad (17)$$

Again, the matrix $[S]$ is totally independent of the value of the function and is solely dependent on the unequally spaced points θ_i . Hence, just as the $[C]$ matrix of the even function, the matrix $[S]$ may also be evaluated separately, independent of $h(\theta_i)$.

3 Application to the CERN Proton Synchrotron Ring

The Fourier analysis above may readily be applied to the CERN PS ring (circumference 628 m), consisting of a vacuum chamber in which the beam is enclosed and circulated by 100 magnet straight sections of nominal length 4.40 m. To locate the beam, monitors have been placed in various magnet sections but these are not equidistant and follow the unequal interval sequence in straight section numbering (in groups of ten) of: 3, 5, 7, 10 and the sequence repeated for every 3rd, 5th, 7th and 10th section up to 100 (i.e. 3, 5, 7, 10 . . . 93, 95, 97, 100). The 40 monitors acquire the beam position with respect to the centre of the vacuum chamber and hence constitute the unequally spaced abscissa points. If we consider the ring as one complete period with straight section 50 at the centre of this period from $-\pi$ to $+\pi$, then each monitor straight section (SS_i) may easily be expressed in terms of radians by:

$$\theta_i = \frac{(SS_i - 50)}{100} 2\pi. \quad (18)$$

The odd and even separation is permitted by considering the straight section position symmetry with respect to the 50th section, i.e. 7 and 93 etc. The Fourier decomposition is then relatively straightforward; for instance, using a 10-term approximation only ($n = 10$), the 11 cosine and the 10 sine amplitude coefficients are readily obtained by using the above equations. Figure 1 illustrates the result of this particular case; curve (a) is the original function obtained from the data acquired remotely from the beam monitors while curve (b) is the function obtained by summations of the Fourier series up to the 10th harmonic using the computed amplitude coefficients. The closeness

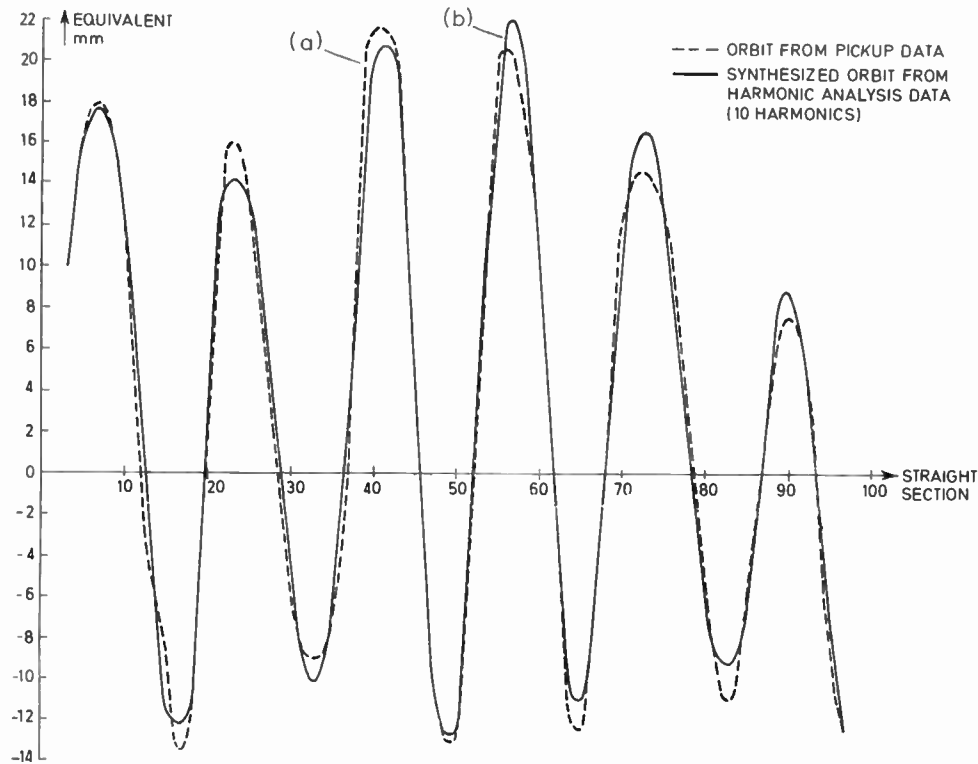


Fig. 1. Measured orbit data and synthesized orbit data based on the algorithm.

of the two curves shows that the approximation is sufficiently accurate even though only the first ten terms of the Fourier series were considered.

An added advantage of this method is that it may readily be applied even if some of the acquisition values are unavailable due to faulty acquisition electronics. In this case, a missing sample point would simply imply a larger interval than before at the same position and at its symmetrical mirror image, but still retaining the mathematical validity in the algorithm above, and without resorting to guesswork for the missing value. Figure 2 illustrates an application for such a case. The acquisition values were not available at monitors 13, 60, 83 and 90 and the symmetry conditions for the odd and even separation implied the effective loss of monitors 87, 40, 17 and 10. Hence the analysis was carried out using only 31 acquisition values of the closed orbit shown in curve (a). The harmonic coefficients produced by using the above algorithm then gave the necessary correction magnet ampere values to yield the minimized closed orbit of curve (b).

4 Method Limitations

The technique of separating the unequally-spaced discrete samples of the periodic signal (or a function) into its odd and even parts has its obvious disadvantages; firstly, some form of symmetry is required for the abscissa sample points over the complete period of interest so as to facilitate this separation; secondly, even if this unequal interval sampling symmetry is partly available, the samples

that do not have their symmetrical counterparts have to be rejected, causing information to be lost. However, taking the worst case of completely random unequal interval sampling, the maximum low-pass filtering (interpolation) required is only over half the period of interest to apply the above technique. While this would imply a not-so-straightforward computer implementation, it should not pose any great problem (for example, in the case illustrated here or in geophysics where such equidistant data occur in time and space).

The method above yields the harmonic coefficients by solving a set of simultaneous equations using matrix inversion. However, the matrix size is directly proportional to the order upto which the spectrum estimation is required. Therefore, for a periodic signal contaminated by non-periodic components, say random noise, the samples would be noisy and a large set of frequencies would need to be considered, requiring cumbersome inversion of large matrices. In such cases, the numerical computer techniques used for inversion may also contribute errors in the coefficient evaluation. Furthermore, for the spectrum estimate of such an ergodic, noisy signal to lie within an acceptable spread of values, it must be averaged over an ensemble of sample batches, say N ; therefore, the large matrix inversions have to be repeated N times.

Finally, although the sampling process may be carried out at unequal interval spacing, the limitation imposed by the sampling theorem still remains and the Nyquist rate must at least be satisfied overall, in terms of the total

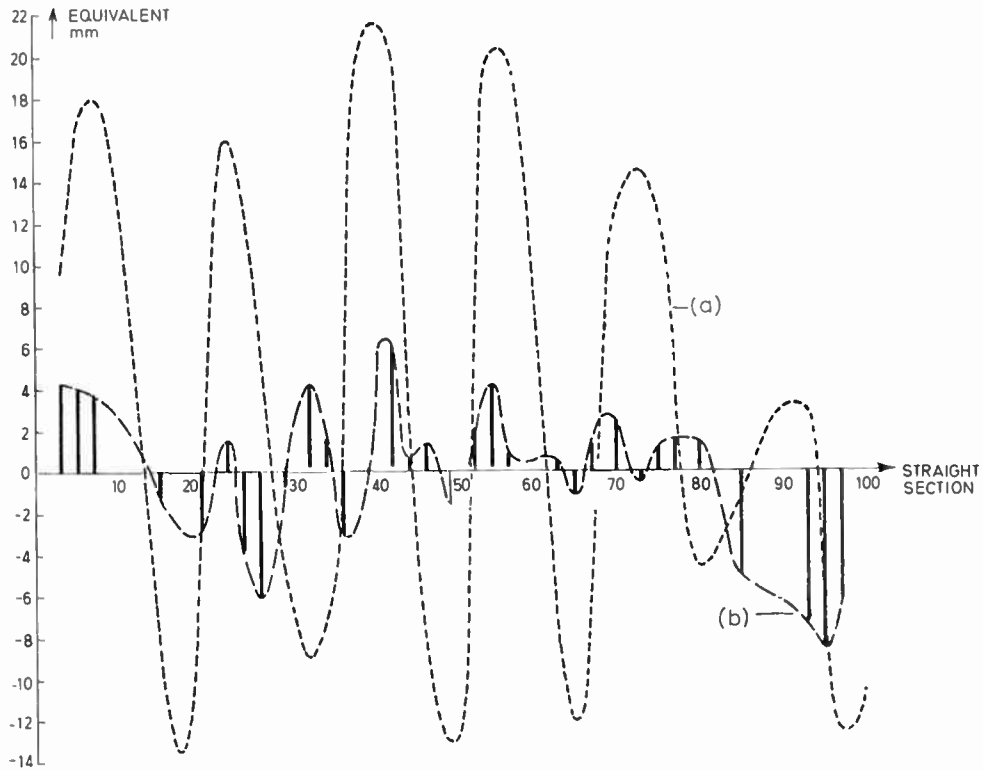


Fig. 2. Discrete data from monitors (36) before correction and closed orbit after correction. Effective number of pick-ups used in approximation is 31.

number of samples per period as related to the highest frequency content of the signal. The signal reconstruction deteriorates with undersampling, consequently affecting the validity of the harmonic components due to aliasing effects. For the low-order (highest frequency < 10 Hz) harmonics as shown in Fig. 1, it was found that the harmonic decomposition was sufficiently accurate even for only 26 samples in that period, which is well within the Nyquist rate requirements.

5 Implementation Details

The above algorithm has been tested and implemented on a PDP 11/45 under the RSX-11D operating system. It consists of three small modules (Fortran Sub-routines) with an overall length totalling about 170 lines. One module carries out the conversion given by eqn. (18), produces the even and odd functions and eliminates the non-symmetrical values. It then evaluates the sine and cosine product matrices (eqns. 8 and 17) ready for inversion. The second module simply inverts these matrices

by the well-known Gauss-Jordan technique. With the odd and even separation as well as the two matrix inversions already carried out, the third module simply computes the sine and cosine coefficients from eqns. (13) and (14).

6 Acknowledgments

The authors would like to thank the referees for pointing out some of the limitations of the analysis technique.

7 References

- 1 Arden, B. W. and Astill, K. H., 'Numerical Algorithms: Origins and Applications', Chap. 8 (Addison-Wesley, New York, 1970).
- 2 Hemming, B. W., 'Introduction to Applied Numerical Analysis', Chap. 12 (McGraw-Hill, New York, 1971).

Manuscript first received by the Institution on 23rd December 1976 and in final form on 10th February 1977.

(Short Contribution No. 1768/Comp. 179.)

© The Institution of Electronic and Radio Engineers, 1977

Tracer decoder—a receiver for radio navigation relay systems

J. D. LAST, B.Sc.(Eng.), Ph.D., C.Eng., M.I.E.E.*
and

E. W. ROBERTS, B.Sc., Ph.D.†

SUMMARY

Tracer is a retransmission system for radio navigation aids which uses a voice frequency communications channel to enable objects to be located and tracked remotely. Its applications are to marine surveying, monitoring of seamarks and automatic vehicle location. The paper describes a unit which accepts the signal transmitted over a link from a Tracer converter and extracts and displays the mobile's position. Its novel features are its ability to indicate position after only one complete sequence of the navigation transmissions, permitting rapid multiplexing of many mobiles, and a frequency correction arrangement which allows the unit to process signals having frequency errors which would normally be unacceptable in high precision systems. Both these techniques are finding applications in radio navigation receiving equipment.

* School of Electronic Engineering Science, University College of North Wales, Bangor, Gwynedd, LL57 1UT.

† Industrial Development (Bangor), University College of North Wales, Bangor, Gwynedd, LL57 1UT.

List of Symbols

a_m, a_s	acceleration with respect to master, slave
c	velocity of propagation
D_{m0}, D_{s0}	distance from master, slave at time t_0
m	noise error of master phase reading
p	correction ratio
s	noise error of slave phase reading
t_0	time of master phase reading
t_d	time interval between master and slave phase readings
v_{m0}, v_{s0}	velocity with respect to master, slave at time t_0
ϕ	phase difference of LOP (line of position)
ϕ_M, ϕ_R etc.	phase count of master, reference etc.
$\Delta\phi$	phase difference
Δt	time difference
$\Delta\phi_{M \rightarrow S_n}$	phase drift between master and slave n readings
ω	angular frequency
ω_m	angular frequency used for phase measurement

1 Introduction

Many radio-navigation systems are of the time-multiplexed type, in which signals at the same frequency are radiated sequentially by a chain of spaced transmitting stations and a receiver measures their phase relationships to provide information to establish its position. In many of these systems, a signal is radiated first from a master station, then from two or more slave stations in turn, the slave stations determining the phase of their transmitted signals by reference to the signals they receive from the master. Examples of such systems are Decca Hi-Fix, Sea-Fix and Hi-Fix/6.^{1,2} The Omega system is similar in principle, but the transmissions of the various stations are derived from high stability frequency standards, with phase corrections being applied from time to time as a result of readings taken at monitor stations.

In all these systems the phase difference between the signals received from any pair of stations is expressed as a lane fraction or position-in-lane. Lane boundaries are usually loci of zero phase difference.

2 Tracer System

A recent addition to the facilities afforded by time-multiplexed radio-navigation chains is the Tracer system,^{3,4} a radio navigation relay system which enables the phase relationships and hence position information received at one point to be relayed or retransmitted over a simple telemetry link to another location. The system can be used to monitor or track the positions of remote objects.

Radio navigation signals are received by the Tracer converter installed on the mobile unit which changes the frequency of the navigation signals from the original radio frequency down to an audio frequency—usually 2 kHz—in a mixer stage. After amplitude limiting, this audio signal is relayed to a receiving centre where the position

information is extracted. The phase differences between the converted audio signal components at 2 kHz are identical to those between the original navigation signals at r.f., so the position information is maintained without significant degradation of accuracy.

The signal telemetered via a Tracer link may be up-converted in frequency and injected into a conventional radio-navigation receiver at radio- or at intermediate-frequency. The receiver then behaves as if it was installed on the mobile. Alternatively, the received audio-frequency signal may with advantage be fed directly to the decoder which is the subject of this paper.

A converter and decoder may, of course, be housed together to form a self-contained navigation receiver.

3 Principles of Operation

3.1 Phase Difference Measurements

In a conventional phase meter, the two signals which are to be compared are present simultaneously. One is used as a phase reference and the phase of the other is measured relative to it. A similar situation exists in frequency multiplexed radio navigation aids such as Decca Navigator⁵ in which, following frequency multiplication, signals of a common frequency are present simultaneously.

In a time-multiplexed system, however, a problem arises because the signals are not present at the same time. One conventional solution to this has been to phase-lock an oscillator in the receiver to the signal burst received from the master station and, as the slave signal bursts are received, to compare their phases with that of this 'master memory' oscillator. This technique is used in Decca Hi-Fix receivers. An alternative solution, used in many other systems, is to phase-lock an individual oscillator to

the signal bursts received from each of the stations. The phase differences between pairs of these memory oscillators may be measured using conventional phase measurement techniques.

When either method is used, the measured position-in-plane on the pattern generated between the master station and slave *n* is given by

$$\phi = \phi_M - \phi_{Sn}$$

The Tracer decoder, in contrast, uses a free-running reference oscillator having a frequency equal to that of the input signal. Here, the procedure is to measure the phase difference between the master and reference and then to subtract from this the phase difference between slave and reference, i.e.

$$\begin{aligned} \phi &= (\phi_M - \phi_R) - (\phi_{Sn} - \phi_R) \\ &= \phi_M - \phi_{Sn} \end{aligned} \quad (1)$$

The frequency and phase of the reference oscillator are arranged to be sufficiently accurate and stable over one complete sequence of the transmissions to allow drift errors to be ignored.

Apart from its simplicity, the main advantage of measuring phase against a free-running reference is that the position information is acquired in one sequence; this leads to fast data updating and makes it possible to time-multiplex several Tracer converters at a relatively rapid rate. In contrast, when a phase-locked oscillator is used, several sequences are required before a satisfactory phase lock is achieved. On the other hand, memory oscillators provide adequate smoothing whilst the above method can be affected by noise and phase perturbations unless the raw data are smoothed using hardware or computer software.

3.2 Phase Discriminator and Digitizer

The basis of the decoder is a phase discriminator linked to an up/down counter. The phase discriminator is a conventional digital arrangement (Fig. 1) in which an edge-triggered bistable is set by a zero-crossing on each cycle of the reference input waveform and reset at the corresponding point on each cycle of the signal input waveform. The output waveform of the bistable is a train of pulses having a mark/space ratio proportional to the phase difference between the two inputs and it is used to gate pulses into a counter (Fig. 2). If the counter is reset once per cycle of the discriminator reference input waveform, and the clock frequency is 1000 times the input frequency, the counter will be clocked to a value in the range of 0-999 which is proportional to the phase difference between the discriminator input signals. This value is called the 'phase count'.

The use of a counter provides a simple and attractive means of combining the phase digitizing and differencing operations. For example, the three-decade up/down counter used as in Fig. 2 enables the phase difference

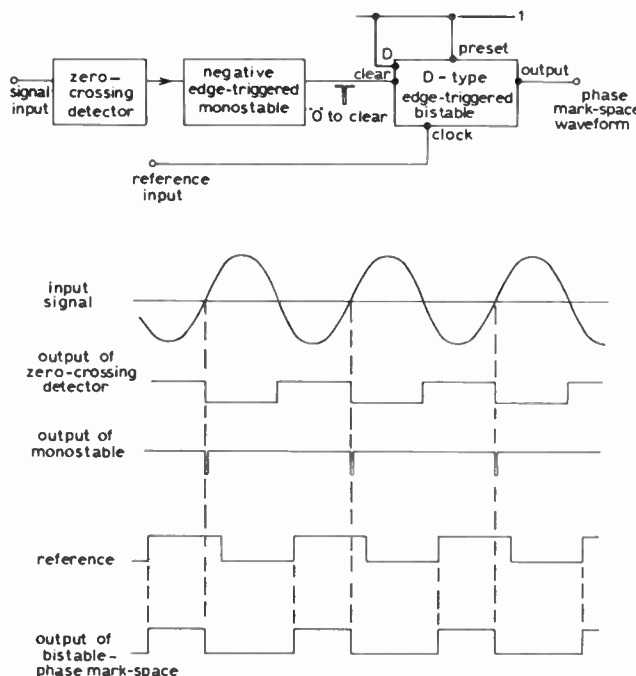


Fig. 1. Phase discriminator.

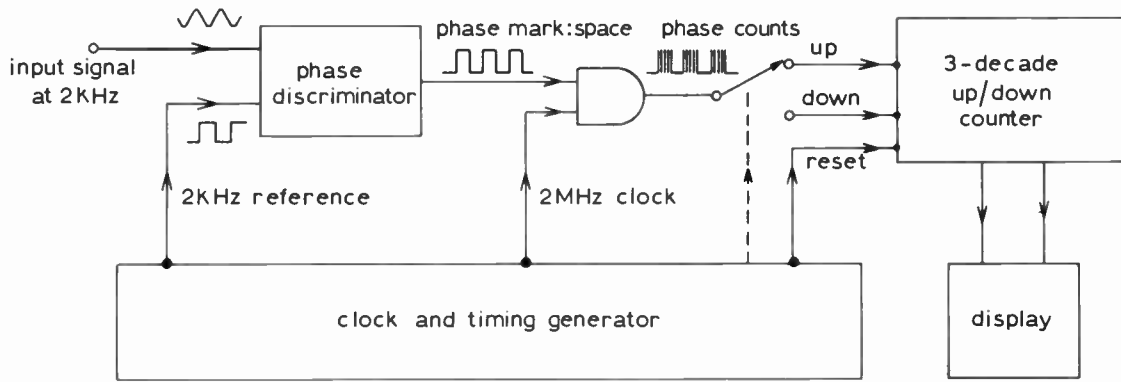


Fig. 2. Measurement of phase difference.

between the time-separated bursts of input signal originating from a master and a slave station to be calculated, using the method described by equation (1) of Section 3. The timing is arranged to implement the following procedure:

- (1) Reset the counter to zero.
- (2) For one cycle of the reference input signal (lasting 500 μ s) occurring whilst the master signal burst is being received, direct the phase count into the counter's up input.
- (3) Repeat (2) during the slave signal burst, but this time clock the counter's down input.
- (4) Latch the contents of the counter out to the display.

At the end of stage (2), the counter holds the value $(\phi_M - \phi_R)$, during stage (3) it counts down by $(\phi_{Sn} - \phi_R)$ leaving the required phase difference, $\phi_M - \phi_{Sn}$ to be latched out to the display.

A simple radio-navigation receiving or decoding system may now be envisaged, based on the above technique

(Fig. 3). The phase discriminator feeds a number of up/down counters (pattern counters), one for each inter-station phase difference which is of interest, via a timing arrangement synchronized to the sequence of signals being received, which also clears the counters and latches the results out to the displays.

4 Phase Difference Measurement with Frequency Correction

The phase difference measurement technique described so far may only be used when the signal frequency is equal to that of the reference. In practice, it is difficult to provide frequency conversion oscillators of sufficient accuracy and stability, especially in mobile Tracer converters, to make the error negligible, and it is necessary to consider the consequences of allowing a frequency difference to exist between the signal and the reference. The result is that the phase count becomes time-dependent and the use of time-separated phase counts to determine the phase difference between burst of signal introduces an error $\Delta\phi(t)$, proportional to the difference Δt between the

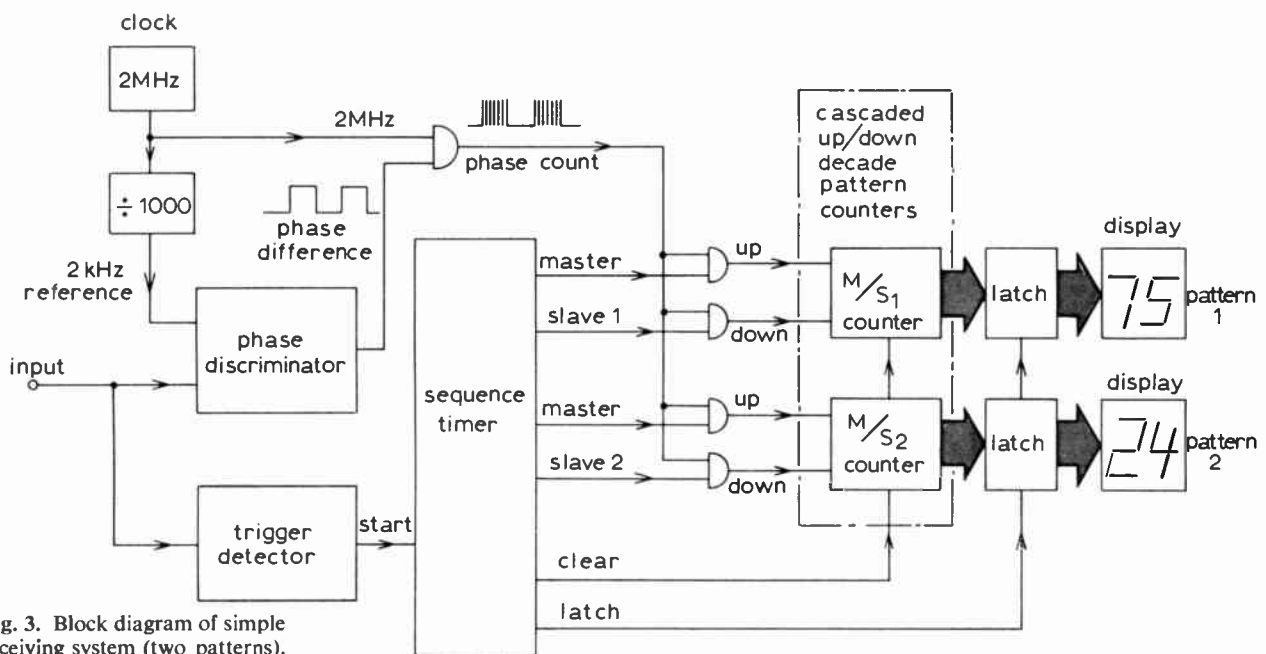


Fig. 3. Block diagram of simple receiving system (two patterns).

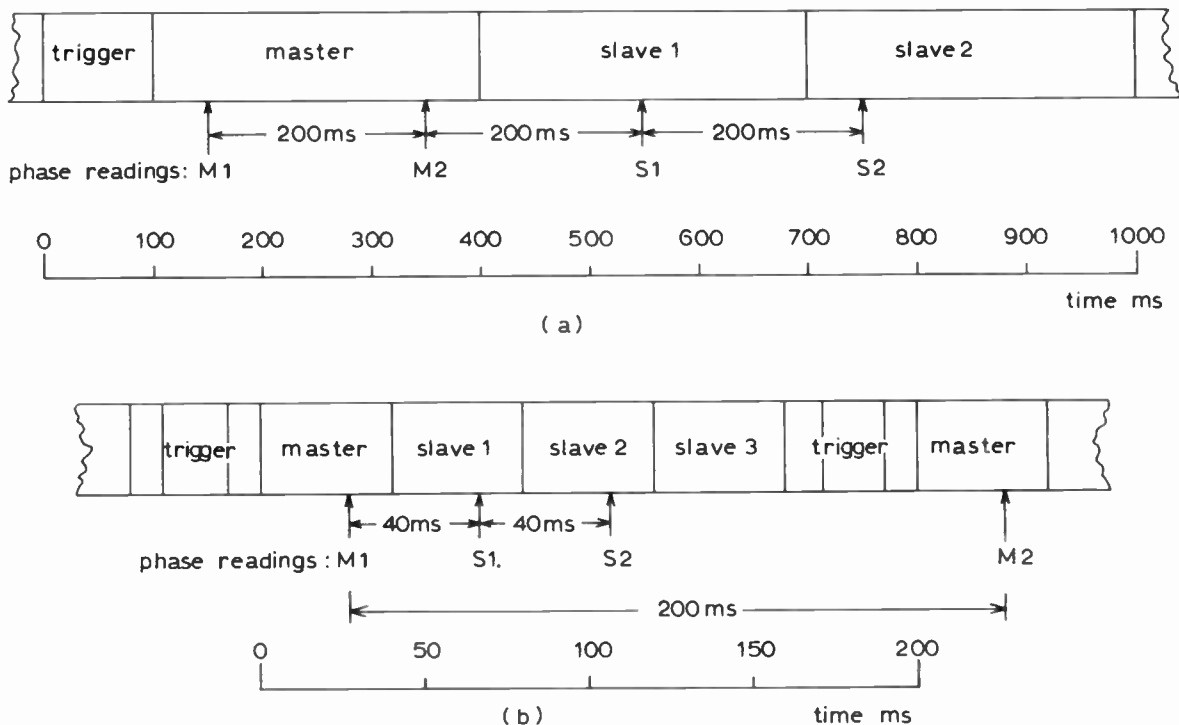


Fig. 4. Timing of sequences (a) Hi-Fix, (b) Sea-Fix.

times at which the two phase samples are taken:

$$\Delta\phi(t) = 2\pi (f_{\text{signal}} - f_{\text{reference}}) \Delta t.$$

When measuring the phase difference between master and slave n , therefore, the cumulative phase error over the period between the master and slave readings ($\Delta\phi_{M \rightarrow S_n}$) must be accounted for. The phase difference measured is now:

$$\phi = \phi_M - \phi_{S_n} - \Delta\phi_{M \rightarrow S_n}.$$

A correction must be applied to eliminate this error; this is evaluated by assessing the cumulative phase error over a fixed length of time. The method assumes that the frequency difference between the input signal and the reference oscillator changes by a negligible amount during a sequence; that is, that the phase difference rate is constant.

An example of the measurement of phase error rate and its correction will now be given, using the timing of the Decca Hi-Fix sequence [Fig. 4(a)]. The transmissions are made by a master and two slave stations and phase differences are measured between each slave and the master. The trigger transmission is for system synchronization (see Sect. 5).

The phase drift is measured by taking two separate phase counts M_1 and M_2 , during the master burst; the change in phase count between the two samples is equal to the cumulative phase error over the time separating them. Once the phase drift over this fixed period is known, the phase error corresponding to any master/slave phase-difference reading may be evaluated by multiplying the measured phase drift by the ratio of the master/slave time

separation to the master/master time separation. This phase error value can then be used to correct the master/slave phase difference reading.

The phase drift is measured using an up/down counter—the ‘correction counter’. The first master phase count (M_1) is counted up and the second master phase count (M_2) is then counted down; this leaves a residual value, $\phi_{M_1} - \phi_{M_2}$ which is the cumulative phase error over the period separating the M_1 and M_2 phase readings.

Because the time separation of M_1 and M_2 is equal to that of M_2 and S_1 (200 ms), the magnitude of the phase error in the master/slave 1 reading is equal to the value held in the M_1/M_2 correction counter. Likewise, since the interval between M_2 and S_2 is 400 ms, the master/slave 2 correction required is twice the correction count.

Assuming positive phase error rate (i.e. $\phi_{M_1} < \phi_{M_2}$), the corrections are now applied by feeding clock pulses to the up input of the correction counter until it reaches zero. Simultaneously the clock pulses are fed to the down input of the M_2/S_1 counter and, for each pulse, two are fed to the down input of the M_2/S_2 counter. These two pattern counters then hold the correct phase difference values, $\phi_M - \phi_{S_1}$ and $\phi_M - \phi_{S_2}$, respectively. This procedure is effected by the circuit shown in Fig. 5. Note that for convenience clock pulses are fed directly to the M_2/S_2 counter and via a divide-by-two stage to the M_2/S_1 and correction counters.

An arrangement of this kind is able to correct for all frequency errors normally encountered in retransmission systems and has been tested with errors of up to ± 500 Hz. The range is limited by certain timing problems although

in practice the trigger arrangement (see Sect. 5) sets the maximum permissible frequency error.

4.1 Interpolation and Extrapolation Correction

In the correction scheme described above, the magnitudes of the corrections applied to the master/slave 1 and master/slave 2 pattern readings are equal to, and twice, the measured phase error respectively. We may say that the correction ratios are 1 and 2. They may be classified as extrapolation corrections; that is, the phase drift of the received signal is measured over a short time interval and used to predict the drift over an equal or longer time interval by extrapolation. The correction ratio is at least unity; furthermore, it is an integer. An alternative arrangement will now be described in which the correction ratio is fractional; the process is one of interpolation rather than extrapolation.

Figure 4(b) shows the timing sequence of the Decca Sea-Fix system. It will be seen that the transmission bursts are much shorter than those of Hi-Fix and timing constraints make it more attractive to take the correction phase samples from two successive master transmission bursts rather than from the same burst. The phase correction to be applied to the master/slave 1 pattern is now $\frac{1}{3}$ th of the measured phase drift, and those of master/slave 2 and master/slave 3 are $\frac{2}{3}$ ths and $\frac{2}{3}$ ths respectively.

An advantage of interpolation correction over extrapolation is that errors in the correction phase measurement due to noise are not multiplied before being incorporated into the corrected pattern values, but rather are reduced. For example, if each reading of master phase contains a noise error of magnitude m and each slave reading, s , the magnitude of the maximum total error of the master/slave 2 pattern in the extrapolation correction example is $3m + s$ and in the interpolation correction example, $1.8m + s$. The error in the pattern reading if correction had not been needed would have been $m + s$, so the increase due to extrapolation correction was $2m$ and interpolation correction $0.8m$ in these particular examples.

A disadvantage of the use of fractional correction ratios is that there is a maximum frequency error beyond which the correction calculated is wrong. If, again referring to Fig. 4(b) for an example, the frequency error is $+6$ Hz, the change of phase between the two master samples is 120 centicycles (1.20 cycles). We now meet an aliasing-type problem: the phase measurements are incomplete sampled data. The only information available is the phase change of 20 centicycles and from this one would calculate the correction to be applied to the master/slave 1 pattern (wrongly) as being $20/5 = 4$ centilanes, instead of $120/5 = 24$ centilanes. Clearly, the phase drift between correction samples must not be allowed to exceed 1 cycle, and, moreover, if one is to be able to determine the sense of the correction (see Sect. 4.2) it must not exceed 50 centicycles. So the maximum frequency error for the Sea-Fix scheme is that which gives ± 50 centicycles of phase change in 200 ms, i.e. ± 2.5 Hz.

The use of interpolation correction always involves a fractional correction ratio and a limitation of permissible frequency error. If extrapolation correction is used, however, the correction ratio may be integral or it may be a rational number (having both an integer part and a fractional part). In the former case, the correction calculated will always be of the right value.

4.2 Sense of Correction

Both the magnitude and the sense of the frequency correction applied to the pattern counters must be right.

Each of the counters used for pattern or correction counting is a cascaded 3-decade unit which holds phase counts in the range 0-999. The counters operate cyclically, that is, counting up from 999 first gives 0, and generates a carry output pulse; counting down from 0 first gives 999 and a borrow pulse.

Referring now to the Hi-Fix timing sequence [Fig. 4(a)], by studying the operation of the correction counter it will be possible to see how the correction system copes with both senses of frequency error. At the start of the sequence the correction and pattern counters are zeroed. The master 1 phase count next increments the correction counter, then the master 2 phase count decrements it, leaving a residue which would be zero if there were no frequency error. One sense of frequency error leaves a positive residue in the counter, i.e. a value less than 500; the other sense causes the counter to pass zero during the master 2 down count, generating a borrow pulse, and leaves a residue greater than 500. When the counter is emptied in order to transfer its correction to the pattern counters it must either be counted down to zero, or up to zero, the direction being determined by whether or not a borrow pulse was generated during the master 2 down count. Further, the correction pulses must be fed to either the up or the down inputs of the two pattern counters as appropriate. A 'routing latch' is used to effect this decision; reset at the start of the sequence, it is set by a borrow pulse and its state controls gates which steer the

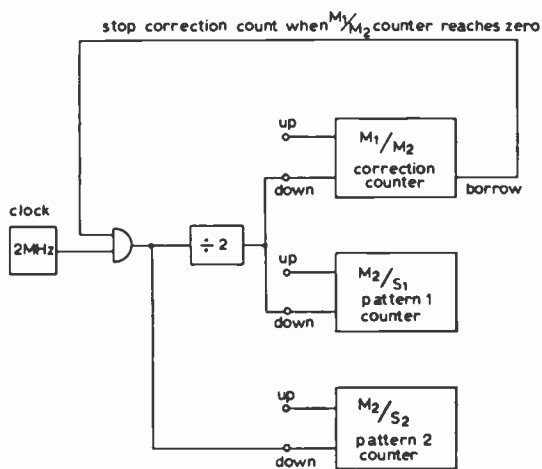


Fig. 5. Correction arrangement of Hi-Fix decoder when phase error rate is positive.

correction clock pulses into the correct inputs of the three counters.

Unfortunately, complications occur because of the cyclic nature of the phase count: a phase drift could result in the master 1 phase count being near to 999 and the master 2 count just above 0, or vice versa. The routing latch would produce faulty corrections in these cases and it is necessary for the system to establish whether the correction counter goes over 500 during the up count and again whether the count crosses 500 on the way down. This information must then be incorporated into the decision as to correction sense.

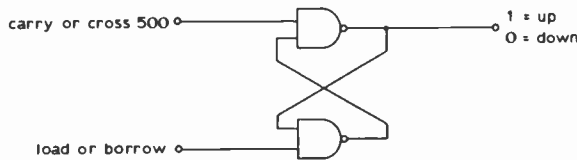


Fig. 6. Circuit which sets sense of corrections.

The procedure may be simplified by starting the correction counter at 500, rather than at 0. The 'crossing of 500' is now indicated by a carry or a borrow pulse from the most significant decade counter and these pulses, together with one from an arrangement which detects the count 500 and an initial reset pulse are fed to the latch shown in Fig. 6, which then determines the proper correction sense in all cases.

In order to prevent a lock-up condition arising due to the 'detect 500' output appearing at the beginning, the correction counter is started at 501, not at 500; the error of 0.1 centicycles is considered acceptable. The pattern counters also have a capacity of 999. However, the percentage phase difference, or position-in-lane expressed centilanes, is given by the two most significant figures and, in order to present the final count rounded to the nearest significant figure, the pattern counters are set initially to 005.

4.3 Complete System

Figure 7 shows a diagram of a receiving system for Hi-Fix using the extrapolation correction scheme described above. The functions of the timing and triggering sections will next be described.

5 Timing and Triggering

The decoder is required to make phase measurements at precisely defined times during the navigation sequence. Other operations such as clearing and loading the counters and latching the position data to the displays also require time allocations. This timing is controlled by a sequence timer which consists of a crystal oscillator driving a chain of binary counters together with a number of multi-input gates. The discriminator generates a phase count once per cycle of the 2 kHz phase reference and these gates direct selected phase counts to the appropriate inputs of the pattern and correction counters. In order to allow one decoder to accept signals received from several different navigation systems, the timing gates are connected to the

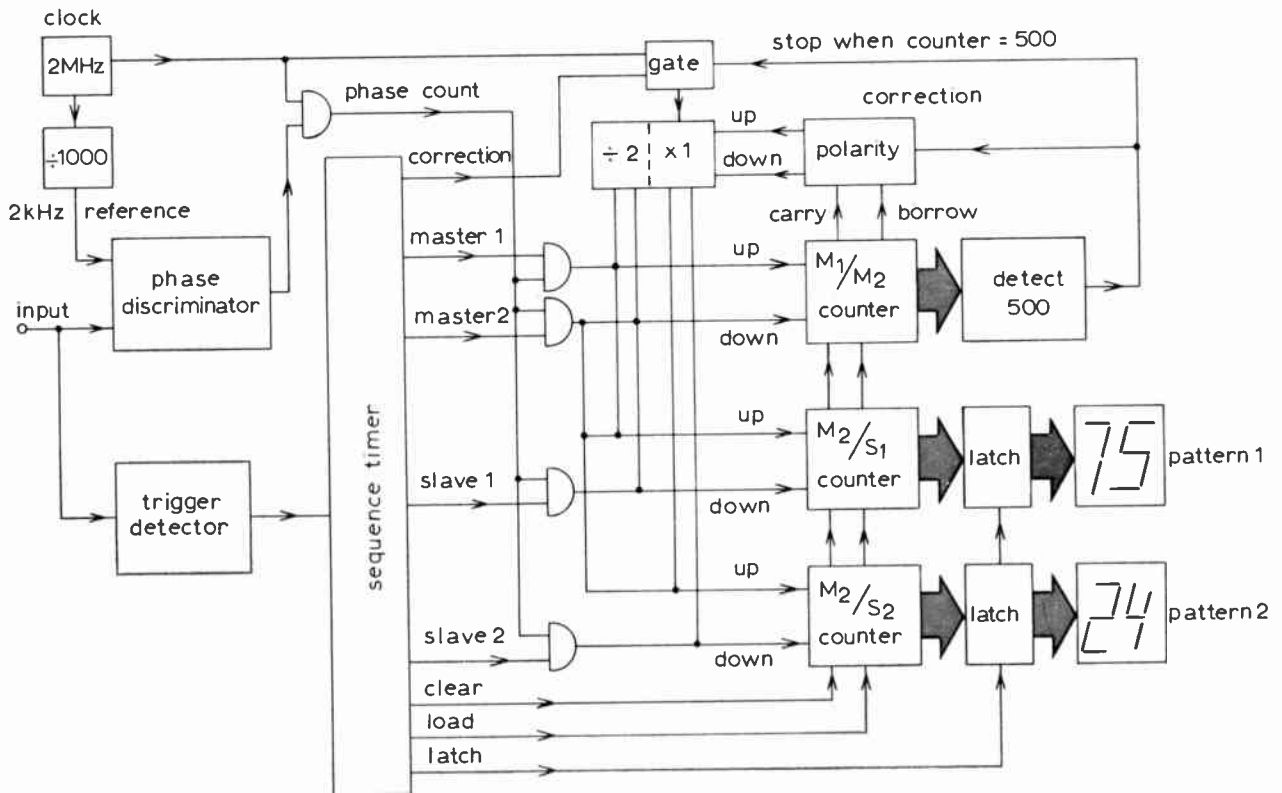


Fig. 7. Block diagram of Hi-Fix decoder.

timing counter outputs via links wired on a patch field, the link pattern being peculiar to the system chosen.

The crystal oscillator and dividers provide all the necessary clock functions including the 2 kHz reference input to the phase discriminator and 2 MHz pulses for transferring the correction from its counter to the pattern counters.

It is essential that the sequence timer in the decoder be synchronized to the received navigation sequence with a maximum tolerance of a few milliseconds. When receiving Decca Hi-Fix and Sea-Fix signals, the decoder achieves this by detecting the trigger burst which the master station transmits at the start of each sequence on a frequency 60 Hz below that used for the patterns. After frequency conversion in a Tracer converter the trigger frequency appears above the pattern frequency, at 2060 Hz. In the decoder the trigger burst is detected by means of a frequency-sensitive switch. The length of the pulse is checked to avoid false triggering under high noise conditions and the back edge is identified and used to start the decoder sequence timer.

The decoder also includes a circuit to examine the incoming 2 kHz sequence to confirm the presence of a burst of signal each time a phase measurement is made which is to be used for position determination. Excessive noise or loss of signal results in the generation of an alarm and the rejection of the data.

6 Significance of the Frequency Correction Process

6.1 Errors which are Cancelled

Frequency differences between the signal and reference inputs to the phase discriminator arise from a variety of sources in a retransmission system or in a receiver. The following contributions are usually substantially constant over the period of a navigation sequence and are cancelled by the frequency correction process.

6.1.1 Small errors

Errors which are usually very small arise in:

- (1) the frequency of the transmitters of the navigation chain;
- (2) Doppler shift in the telemetry link due to the movement of the mobile relative to the base station.

6.1.2 Converter local oscillator frequency error

This is the most serious source of error and the availability of wide frequency range extrapolation correction (and, to a lesser extent, interpolation correction) has permitted a significant relaxation in the accuracy and stability specifications of local oscillators fitted in mobile converters. This is important because the use of a high stability crystal oscillator in an oven is expensive both in cost and in power consumption. Not only is the oven found to be unnecessary but, with careful system design, it becomes possible to switch off the oscillator when it is not required as, for example, in systems of several mobiles

which transmit in turn to a base station. This leads to low duty-cycle operation and makes it possible to design equipment of long endurance. A converter power consumption of, typically, 1.2 W is reduced to an average consumption of 100–200 mW by this technique.

6.2 Errors which arise from the Velocity and Acceleration of the Mobile

The movement of the mobile during the phase readings themselves gives rise to errors which are usually small. The fractional change in the phase reading may be shown to be

$$\frac{\Delta\phi}{\phi} \approx \frac{\omega v}{\omega_m c}$$

where ω_m is the angular frequency at which the phase measurement is carried out.

The practical effect of this error is likely to be negligible. For example, a mobile travelling at 1 km/s (2236 miles/h) in the baseline region of a navigation system at 2 MHz with down-conversion to a measurement frequency of 2 kHz suffers a phase measurement error of only 0.3%.

Much more significant is the movement of the mobile between phase readings. Analysis shows that, if the master transmissions are used for frequency correction, the effective instant of measurement of the line of position is that at which the slave phase reading is taken, irrespective of the velocity and heading of the mobile. In effect, by making a correction phase difference measurement, the master phase rate at the receiver is estimated and the estimate used to calculate a value of master phase which is correct at the instant of the slave phase reading. Both are true simultaneously, as for a stationary receiver.

If the transmissions from the slave station are used for correction measurement rather than those from the master, the effective instant of measurement is placed at the master phase sample. By making two phase measurements on each slave transmission and correcting each LOP independently using the two readings, simultaneously true LOPs may be calculated by projecting back to time t_0 , or forward to some other convenient instant if a second master reading is also taken.

In other cases, however, if it is assumed that the LOPs measured on two (or more) patterns are simultaneously true, the intersection of those LOPs will not in general lie on the path of the mobile; that is, skew errors will result.

Acceleration of the mobile between phase samples introduces further position measurement errors. If the components of this acceleration are constant for the duration of a set of phase readings the corrected LOP measured between two transmitters may be shown to be

$$\frac{\omega}{c} \{D_{mo} - D_{so}\} + \frac{\omega t_d}{c} \{v_{mo} - v_{so}\} + \frac{\omega t_d^2}{c} \{pa_m - a_s\}.$$

In this expression the individual terms may be identified (in order) as the LOP at time t_0 , the velocity term and the acceleration term.

7 Applications of the Decoder

Decoding systems have been designed, built and used in conjunction with the Hi-Fix and Sea-Fix systems, Hi-Fix/6,² with a novel adaptation of Decca Navigator² and with an automatic location system⁶ for motor vehicles. In all cases, both Tracer retransmission and direct receiving systems have been tested.

The system described is the subject of US Patent No. 3,968,493 and of British Patent No. 1,439,850.

8 Acknowledgments

The Tracer system was devised at the School of Electronic Engineering Science, University College of North Wales and the engineering development was carried out by Industrial Development (Bangor), a commercial unit within the School, in collaboration with Decca Survey Ltd., Leatherhead, Surrey.

The authors acknowledge the agreement of the Directors of Decca Survey Ltd. to the publication of this paper and the valuable assistance of Mr. J. M. Thomson, Mr. A. R. Woods, Mr. D. J. Phipps and Mr. D. Boorman of that company.

They also acknowledge the contributions of Mr. A. J. Wade and Mr. S. Jones of the University College of North

Wales to the analysis of the effects of the velocity and acceleration of the mobile, summarized in this paper.

9 References

- 1 Lee, J. K. V., 'A position-fixing aid to oceanography', Conference on Electronic Engineering in Oceanography, p. 48, 1966. (IERE Conference Proceedings No. 8.)
- 2 Phipps, D. J., Last, J. D., Prior-Jones, R. LI, Roberts, E. W. and Simpson, J. H., 'A position determination system and data relay', Conference on Instrumentation in Oceanography, p. 203, 1975. (IERE Conference Proceedings No. 32.)
- 3 Last, J. D. and Roberts, E. W., 'Tracer—a voice-frequency relay system for radio-navigation aids', *The Radio and Electronic Engineer*, 45, No. 6, p. 299, 1975.
- 4 Last, J. D. and Roberts, E. W., 'A buoy tracking system for water movement surveys', *The Dock and Harbour Authority*, 53, No. 628, p. 391, 1973.
- 5 Powell, C., 'The Decca Navigator system for ship and aircraft use', *Proc. Instn Elect. Engrs*, 105, Part B, Suppl. No. 9, p. 225, 1968.
- 6 Deslandes, P. A. and Last, J. D., 'A system for automatic vehicle location', International Conference on Automobile Electronics, p. 112, 1976. (IEE Conference Publication No. 141.)

Manuscript first received by the Institution on 13th December 1976 and in final form on 25th March 1977 (Paper No. 1769/AMMS81.)

© The Institution of Electronic and Radio Engineers, 1977

The Authors



David Last received the degree of B.Sc.(Eng.) in electrical engineering from the University of Bristol in 1961 and from 1961-63 was a Graduate Apprentice with the British Broadcasting Corporation. He received the degree of Ph.D. in electronic and electrical engineering from the University of Sheffield in 1966 and was then appointed to a lectureship in the School of Electronic Engineering Science at

the University College of North Wales, Bangor. Dr. Last has published a number of papers in the fields of non-linear



Emyr Roberts is a Project Engineer at Industrial Development (Bangor), the commercial unit within the School of Electronic Engineering Science of the University College of North Wales, Bangor, where he has stayed since gaining his B.Sc. degree in electronic engineering in 1962; he was awarded his doctorate in 1969. When Dr. Roberts joined IDB in 1968 he was initially responsible for the design and

construction of acoustic and resistivity problems for sea-bed sediments. For the past five years he has been involved with the design and development of Tracer navigation systems.

Simplified solutions of a cross-junction and specially truncated bends in parallel-plate waveguides

M. F. ISKANDER, B.Sc., M.Sc., Ph.D.*

and

M. A. K. HAMID, Ph.D., P.Eng., C.Eng., F.I.E.E.*

SUMMARY

The mode matching technique is employed to solve for the scattering parameters in cross-junction and specially truncated bends in parallel-plate waveguides where the general problem reduces to the solution of three canonical boundary value problems. The basic geometry in all three cases involves a finite wedge region overlapping with a semi-infinite parallel-plate region. The scattering coefficients in each case are determined by matching across the common boundaries the eigenfunctions expansion in each region. Typical results to demonstrate the accuracy of the reflection coefficient and the convergence of the infinite series are presented.

1 Introduction

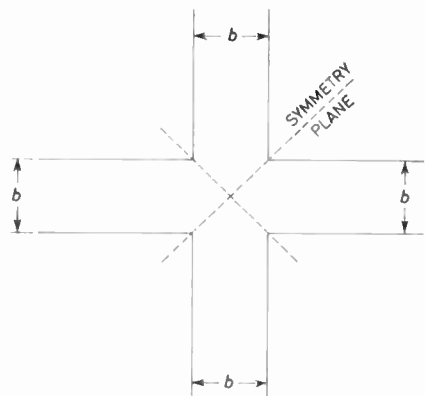
Sharp discontinuities in waveguides and strip transmission lines occur commonly whenever there is a change in the dimension or direction of the line or there is a need for power splitting, impedance matching, filtering, tuning, mode convergence, etc.¹ Although appropriate discontinuities when placed in waveguide sections may perform numerous useful functions, they also introduce several side effects such as distortion, insertion losses due to mode convergence and reflection, and they also limit the operating bandwidth of the waveguide and life time of the generator. As a result the scattering characteristics of a very large number of waveguide discontinuities have been carefully studied in the last few decades and the results have been presented in the literature.²⁻⁴

The classical approach of analysis is to derive an equivalent network for the discontinuity based on the scattering matrix elements of the junction and to use the result for describing the performance of the waveguide component in question.⁵ Several recent extensions in analytical, quasi-analytical, experimental as well as numerical solutions have been reported by various authors.⁶⁻¹⁰

Cross-shaped junctions (symmetrical four-port element) and right-angle corners in rectangular waveguides are among a class of particularly important elements in microwave circuits because of their use in mode conversion and power splitting. Special effort to study these junctions has been reported by Rice who used the conformal mapping technique to solve for the reflection coefficient in a truncated corner,¹¹ while Goyal¹² employed an integral equation approach based on the Lorentz reciprocity theorem. A quasi-optical approach and a wedge diffraction analysis were also employed by Kao¹³ and Lewin,¹⁴ respectively, to generate adequate solutions of overmoded junctions. Bouwkamp,¹⁵ on the other hand, employed an integral formulation to solve for the scattering characteristics of a cross junction. Application of known waveguide solutions to study the characteristics of sharp bends in the centre strip of a balanced strip transmission line was discussed by Oliner.¹⁶ The mode matching solution derived by Campbell¹⁷ and later extended by Wardrop¹⁸ to the overmoded case is found to be particularly attractive for its generality, convenience of manipulation and numerical accuracy. The approach is based on the utilization of electric and magnetic symmetry planes along the diagonals of the junction to identify three canonical boundary value problems. The solution is then obtained by dividing the geometry of each of the canonical problems into a semi-infinite parallel plate region and a triangular region and matching the field expansion of each on a common boundary. However, their lengthy and complicated derivation may be improved by incorporating the method of overlapping regions into the mode matching solution.

It is, therefore, the purpose of this paper to present simple solutions for such junctions by extending the use

* Antenna Laboratory, Department of Electrical Engineering, University of Manitoba, Winnipeg, Manitoba, Canada R3T 2N2.



(a) Schematic diagram of the symmetrical cross-shaped junction.

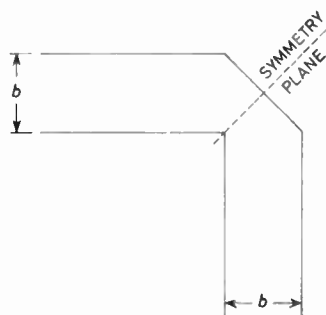


Fig. 1.

(b) Schematic diagram of the right-angled corner.

of the canonical problems recognized earlier by Campbell.¹⁷ The extension involves recognizing each canonical problem as being associated with a finite wedge-shaped region which overlaps with a semi-infinite parallel-plate region. An eigenfunction expansion is employed in each region and the expansion coefficients are determined simply by mode matching across a common boundary with an adjacent region.

2 Formulation

Consider the parallel-plate symmetrical cross-shaped junction shown in Fig. 1(a). From symmetry considerations, it is clear that the scattering characteristics of such junction can be obtained for an incident wave in only one arm, e.g. arm I. Furthermore, it was shown that a complete analysis of such junction as well as the right-angle corner [Fig. 1(b)] can be obtained from the solution of the reflection coefficients in each of the three boundary value problems denoted by A, B and C in Fig.2.¹¹ Hence, if R_A , R_B and R_C are the reflection coefficients of the TEM mode in the cases A, B and C respectively, then the elements α , β and γ of the scattering matrix are given, respectively, by

$$\alpha = (R_A + 2R_B + R_C)/4 \tag{1}$$

$$\beta = (R_A - R_C)/4 \tag{2}$$

$$\gamma = (R_A - 2R_B + R_C)/4 \tag{3}$$

For an incident TEM mode in the semi-infinite parallel-plate region, I, the axial component of the magnetic field

satisfying the two-dimensional wave equation as well as the boundary condition at $y = \pm b/2$ is given by

$$H_{z1}(x, y) = \exp(jkx) + R_0 \exp(-jkx) + \sum_{n=1}^{\infty} R_n \exp(-jk_n x) \cos\left(\frac{n\pi}{b}\right)(y + b/2) \tag{4}$$

where

$$k_n = \left[k^2 - \left(\frac{n\pi}{b}\right)^2 \right]^{1/2}$$

and the $\exp(-j\omega t)$ has been suppressed.

The magnetic field $H_{z11}(r, \phi)$ in region II, which is a finite wedge-shaped region of arm length l and apex at the origin O_2 , has to satisfy the appropriate boundary condition depending on the case in question. Thus for case A we have

$$H_{z11}(r, \phi) = \sum_{m=0}^{\infty} A_m J_{\nu}(kr) \cos \nu(\phi + \alpha/2), \tag{5}$$

$$\nu = m\pi/\alpha$$

while for case B

$$H_{z11}(r, \phi) = \sum_{m=0}^{\infty} B_m J_{\sigma}(kr) \sin \sigma(\phi + \alpha/2), \tag{6}$$

$$\sigma = (m + \frac{1}{2})\pi/\alpha$$

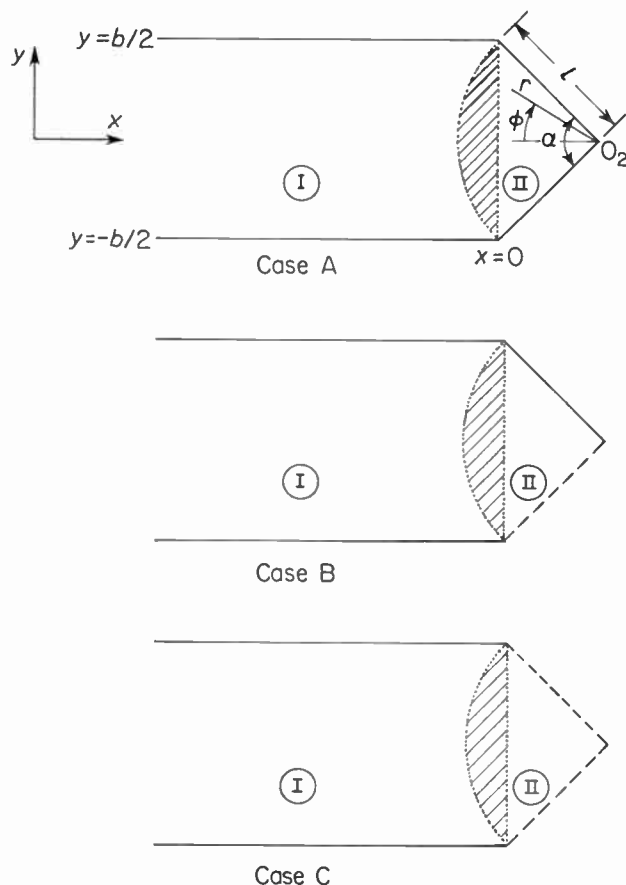


Fig. 2. Schematic diagram associated with the three boundary value problems with different planes of symmetry. — electric plane of symmetry; - - - magnetic plane of symmetry. (The overlapping regions are shaded.)

and for case C

$$H_{z11}(r, \phi) = \sum_{m=1}^{\infty} C_m J_{\mu}(kr) \sin \mu(\phi + \alpha/2),$$

$$\mu = m\pi/\alpha \quad (7)$$

where α is the total flare angle of the wedge-shaped region.

It is clear that the expansions in (5) to (7) are of the same form except for the trigonometric functions which are so chosen as to satisfy the boundary condition in each case. It is to be noted that such expansions are simple and hence offer certain advantages over previous formulations.¹⁷

To determine the expansion coefficients in (4) which are required to calculate the elements of the scattering matrix of the junction, the continuity of the magnetic field and its normal derivative across the boundary $r=l$ (with respect to origin O_2) are imposed. Hence for case A, after using the orthogonality property of the trigonometric function, we obtain

$$A_m = \frac{\epsilon_m}{\alpha J'_\nu(kl)} \int_{-\alpha/2}^{\alpha/2} \cos \nu \left(\phi + \frac{\alpha}{2} \right) \times$$

$$\times \left\{ \exp [jk(d-l \cos \phi)] + \right.$$

$$+ R_0 \exp [-jk(d-l \cos \phi)] +$$

$$+ \sum_{n=1}^{\infty} R_n \exp [-jk_n(d-l \cos \phi)] \times$$

$$\left. \times \cos \left(\frac{n\pi}{b} (l \sin \phi + b/2) \right) \right\} d\phi \quad (8)$$

and

$$A_m = \frac{\epsilon_m}{\alpha k J'_\nu(kl)} \int_{-\alpha/2}^{\alpha/2} \cos \nu \left(\phi + \frac{\alpha}{2} \right) \times$$

$$\times \left\{ -jk \cos \phi \exp [jk(d-l \cos \phi)] + \right.$$

$$+ R_0 jk \cos \phi \cdot \exp [-jk(d-l \cos \phi)] +$$

$$+ \sum_{n=1}^{\infty} R_n \left(jk_n \cos \phi \exp [-jk_n(d-l \cos \phi)] \times \right.$$

$$\times \cos \left[\frac{n\pi}{b} (l \sin \phi + b/2) \right] -$$

$$\left. - \frac{n\pi}{b} \sin \phi \exp [-jk_n(d-l \cos \phi)] \times \right.$$

$$\left. \times \sin \left[\frac{n\pi}{b} (l \sin \phi + b/2) \right] \right\} d\phi \quad (9)$$

where

$$\epsilon_m = \begin{cases} 1, & m = 0 \\ 2, & m > 0 \end{cases} \quad \text{and} \quad d = l \cos \left(\frac{\alpha}{2} \right)$$

Similar expressions can be obtained for cases B and C. It is clear that, by eliminating A_m between (8) and (9) and truncating the two infinite series to N terms, the expansion coefficients R_n can be easily obtained by a simple matrix inversion.

Table 1. Convergence test of the series expansion in (4) for case B and $kb = 40^\circ$

n	R_n					
	N = 3		N = 7		N = 12	
0	-0.8653	-j0.5013	-0.8725	-j0.4896	-0.8731	-j0.4875
1	-0.0730	+j0.2716	-0.0732	+j0.2786	-0.0729	+j0.2801
2	-0.0067	+j0.0247	-0.0113	+j0.0421	-0.0113	+j0.0435
3			-0.0096	+j0.0357	-0.0096	+j0.0368
4			-0.0042	+j0.0147	-0.0040	+j0.0154

Table 2. Convergence test of the series expansion in (4) for case B and $kb = 120^\circ$

n	R_n					
	N = 3		N = 7		N = 12	
0	0.1208	-j0.9927	0.0974	-j0.9952	0.0926	-j0.9957
1	-0.7005	+j0.6204	-0.7129	+j0.6466	-0.7161	+j0.6527
2	-0.0579	+j0.0513	-0.0868	+j0.0787	-0.0898	+j0.0818
3			-0.0615	+j0.0558	-0.0640	+j0.0584
4			-0.0281	+j0.0255	-0.0297	+j0.0271

3 Numerical Results

Numerical values for the reflection coefficient in each of the three cases have been calculated using the IBM 370/158 electronic computer. A sexamples to illustrate the convergence of the series in (4) calculations for different values of N for case B and $kb = 40^\circ$ and $kb = 120^\circ$ are shown in Tables 1 and 2, respectively. From Tables 1 and 2 it is clear that even for large values of kb the truncation of the series to $N = 7$ which requires c.p.u. time equal to 4.8 seconds, gives quite adequate results.¹⁷ The reflection coefficients of the three cases are shown in Fig. 3. It is to be noted that when our results are combined as shown in (1) to (3) to calculate the scattering matrix of the cross-shaped junction they show perfect agreement with previously published data by Campbell†.¹⁷ Therefore the present formulation retains the advantage of simplicity so that the distinction between the three cases is self-explanatory from (5) to (7). It should also be noted that calculations of the equivalent circuit parameters are straightforward¹⁷ and need not be repeated.

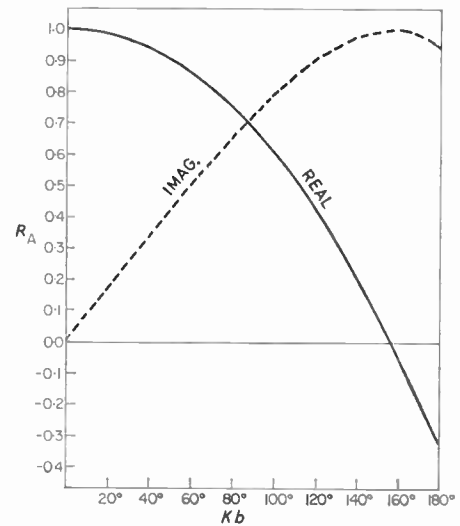
4 Discussion and Conclusions

A simple solution for the cross-shaped and right-angle corner has been presented. The procedure involves dividing the geometry of the canonical boundary value problem derived from symmetry considerations into two overlapping regions for each of which a simple eigenfunction expansion can be easily formulated. The expansion coefficients can then be calculated by applying the mode matching technique. It should be noted that Schwarz's iterative procedure of overlapping regions, which would employ the same overlapping regions used here, can also be used.¹⁹ The present procedure, however, though restricted to simple geometries, is obviously more efficient since no iterations are required. Since the numerical results are in very good agreement with previously published data, the present formulation retains the advantage of being simpler and easier to manipulate. It should be also noted that extension of the solution to deal with overmoded junctions is straightforward since there are no inherent limitations in the analysis on the number of propagating modes.

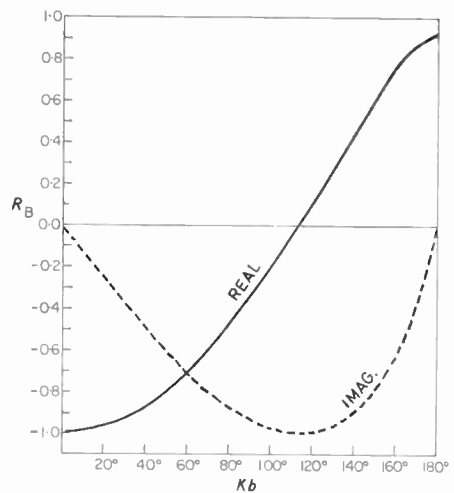
It should also be noted that the results (1) to (3) for the scattering matrix of the parallel-plate waveguide discontinuities can be readily used to describe similar discontinuities in rectangular waveguides as well as in strip lines. For a rectangular waveguide supporting only the TE_{10} mode, the results of Fig. 3 are directly applicable with kb replaced by $k_g b$ where $k_g = 2\pi/\lambda_g$ and λ_g is given by

$$\lambda_g = \frac{1}{\sqrt{\left[\left(\frac{1}{\lambda}\right)^2 - \left(\frac{1}{2h}\right)^2\right]}} \quad (10)$$

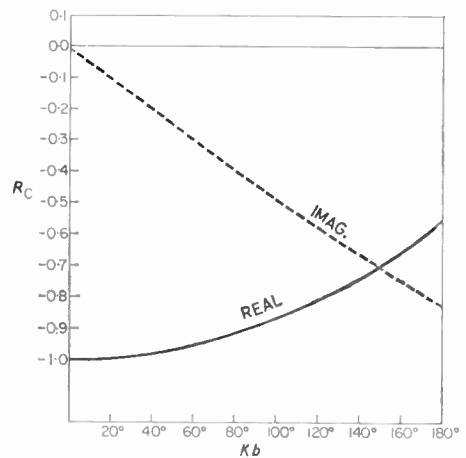
† A phase correction should be introduced to account for the different time harmonic variations encountered in both calculations.



(a) R_A for case A



(b) R_B for case B



(c) R_C for case C.

Fig. 3. Reflection coefficient of the TEM mode for cases A, B and C.

where h is the width of the rectangular waveguide. For stripline discontinuities, on the other hand, the scattering matrix elements are simply given by the negative of those calculated using (1) to (3) together with Fig. 3. This is

simply because the conformal transform of the stripline can be shown to be identical to the Babinet equivalent of the parallel-plate waveguide rotated by 90° .¹⁶ Therefore, the admittance matrix representing the discontinuity in the original structure is equal to the impedance matrix of the Babinet equivalent structure and hence the scattering matrix follows from a similar relation. The width D of the parallel-plate waveguide is related to the actual centre strip width W by¹⁶

$$D = HK(k')/K(k'') \quad (11)$$

where $k' = \tan h(\pi W)/2H$, H is the separation distance between the two ground planes in the balanced stripline, $K(k')$ is the complete elliptic integral of the first kind of modulus k' and $k'' = \sqrt{1-k'^2}$.

5 Acknowledgment

The authors wish to acknowledge the financial assistance of the National Research Council of Canada, Grant A-3326, which made this research possible.

6 References

- 1 Matthaei, G. L., Young, L. and Jones, E. M. T., 'Microwave Filters, Impedance-Matching Networks and Coupling Structures' (McGraw-Hill, New York, 1964).
- 2 Marcuvitz, N., 'Waveguide Handbook' (McGraw-Hill, New York, 1951).
- 3 Ragan, G. L., Editor, 'Microwave Transmission Circuits' (McGraw-Hill, New York, 1948).
- 4 Collin, R. E., 'Foundation for Microwave Engineering' (McGraw-Hill, New York, 1966).
- 5 Southworth, G. C., 'Principles and Applications of Waveguide Transmission', chap. 8 (Van Nostrand, New York, 1959).
- 6 Lewin, L., 'Theory of Waveguides' (Butterworths, London, 1975).
- 7 Yee, H. Y. and Felsen, L. B., 'Ray optics—A novel approach to the scattering by discontinuities in a waveguide', *IEEE Trans. on Microwave Theory and Techniques*, MTT-17, pp. 73–85, February 1969.

- 8 Clarricoats, P. J. B. and Slinn, K. R., 'Numerical solution of waveguide-discontinuity problems', *Proc. Instn Elect. Engrs*, 114, No. 7, pp. 878–86, July 1967.
- 9 Wexler, A., 'Solution of waveguide discontinuities by modal analysis', *IEEE Trans.*, MTT-15, pp. 508–17, September 1967.
- 10 Chow, Y. L. and Wu, S. C., 'A moment method with mixed basis functions for scatterings by waveguide junctions', *IEEE Trans.*, MTT-21, pp. 333–40, May 1973.
- 11 Rice, S. O., 'Reflection from corners in rectangular waveguides—conformal transformation', *Bell Syst. Tech. J.*, 28, pp. 136–56, 1949.
- 12 Goyal, K. G., 'Analysis of the field structure in corners and tees in rectangular waveguides', M.Sc. thesis, University of Toronto, 1966.
- 13 Kao, K. C., 'Approximate solution of the H-plane right-angled corner in overmoded rectangular waveguide operating in the H_{10} mode', *Proc. Instn Elect. Engrs*, 111, No. 4, pp. 624–8, 1964.
- 14 Lewin, L. and Crain, G. E., 'Theoretical performance of the overmoded right-angle corner', *Proc. Instn Elect. Engrs*, 116, No. 5, pp. 667–78, 1969.
- 15 Bouwkamp, C. J., 'Scattering characteristics of a cross-junction of over-sized waveguides', *Philips Tech. Rev.*, 32, pp. 165–77, 1971.
- 16 Oliner, A. A., 'Equivalent circuits for discontinuities in balanced strip transmission line', *IRE Trans.*, MTT-3, pp. 134–43, 1955.
- 17 Campbell, J. J., 'Application of the solutions of certain boundary value problems to the symmetrical four-port junction and specially truncated bends in parallel-plate waveguides and balanced strip-transmission lines', *IEEE Trans.*, MTT-16, pp. 165–76, 1968.
- 18 Wardrop, B., 'Analysis of quasioptical waveguide corners', *Electronics Letters*, 6, No. 6, pp. 154–5, 1970.
- 19 Iskander, M. F. and Hamid, M. A. K., 'Scattering coefficients at a waveguide-horn junction', *Proc. Instn Elect. Engrs*, 123, No. 2, pp. 123–7, 1976.
- 20 Borovikov, V. A. and Eyodus, A. G., 'Diffraction at a right-angled discontinuity in a waveguide with a mirror', *Radio Engng Electronic Phys.*, 21, No. 1, pp. 38–46, 1976.

Manuscript first received by the Institution on 12th November 1976 and in final form on 19th January 1977. (Paper No. 1770(CC272).)

© The Institution of Electronic and Radio Engineers, 1977

The Authors



Magdy F. Iskander received his B.Sc. degree with first class honours in electrical engineering from the University of Alexandria, Egypt, in 1969. He entered the Faculty of Graduate Studies at the University of Manitoba and was awarded the M.Sc. and Ph.D. degrees in microwaves in September 1972 and 1976 respectively. From 1969 to 1971 he was employed as a teaching assistant in the Department of

Electrical Engineering at the University of Alexandria and at that time worked on the design and development of ion sources. In 1976 he was granted a National Research Council of Canada Post-Doctoral Fellowship at the University of Manitoba where he is currently employed as a Research Associate in the Department of Electrical Engineering.



M. A. K. Hamid received the B.Eng., and M.Eng. degrees from McGill University, Montreal in 1960 and 1962 respectively, and the Ph.D. degree from the University of Toronto in 1966, all in electrical engineering. In 1960 he joined the Aluminium Company of Canada, Montreal as a Planning Engineer and in 1963 became Senior Consulting Engineer with the Satellite Antennas Design Group at Sinclair Radio

Laboratory in Toronto. Since 1965 he has been with the Department of Electrical Engineering at the University of Manitoba where he is now a Professor of Electrical Engineering and an Adjunct Professor of Agricultural Engineering. Besides sitting on numerous governmental and academic bodies, he is a member of the Canadian Delegation to the International Union of Radio Science (URSI).

The transfer function synthesis of LC ladder networks

SHYAMA CHARAN PRASAD, M.Sc., Ph.D.*

and

YUGAL KISHORE YADAV, M.Sc.†

SUMMARY

The recurrent-continuant procedure as previously described¹ conforms to a transfer function polynomial that leads to an LCR ladder structure. In this contribution a modification to the procedure is proposed that extends its application to the realization of pure LC structures and thence, by the use of simple frequency transformation, to the realization of pure RC structures, as well. In the proposed scheme, an appropriate function is added to the prescribed LC transfer function to transform the latter into a strictly Hurwitz class of polynomial which is now used to synthesize the desired network. The method is capable of yielding multiple sets of element values.

* Department of Physics, Langat Singh College, University of Bihar, Muzaffarpur 842001, India; now at the Indian Institute of Technology, New Delhi.

† Department of Physics, Ram Dayalu Singh College, University of Bihar, Muzaffarpur 842001, India.

Recurrent-Continuant Transformation

Starting with the voltage transfer function, the recurrent-continuant transformation method is a convenient tool for the synthesis of ladder networks. The transformation techniques described by Holbrook¹ and Jha and Prasad² lead to low-pass lossless ladders singly and doubly terminated in resistances. Although these resistances may be absorbed into a practical generator and/or load and the resulting network may be assumed to be purely reactive, the synthesized network is essentially an LCR network and these techniques cannot be used as such to obtain a purely LC network. In this paper, the authors describe a method for the synthesis of LC ladder networks through this procedure. This method may be used for the synthesis of RC ladders as well, by adopting an appropriate frequency transformation.

The voltage transfer function of a LC network N (Fig. 1, $Z_i = L_i s$, $Y_i = C_i s$) has its poles on the imaginary axis of the complex s -plane and the reciprocal voltage transfer function may be written as

$$T(s) = P_n(s) = \sum_{i=1}^n a_{2i} s^{2i} + 1. \quad (1)$$

If an external resistance R were to be introduced at either of the two ports of N , the network will be modified to an LCR network N' and $T(s)$ will change to $T'(s)$. $T'(s)$ will belong to the strictly Hurwitz class of polynomials with its roots lying anywhere in the left half complex s -plane and one can write

$$T'(s) = P_n'(s) = \sum_{i=1}^n b_i s^i + 1. \quad (2)$$

The polynomial coefficients b_i will obviously be positive, real and will obey the Hurwitz restriction.

The difference in $T(s)$ and $T'(s)$ may be ascribed to R and one must have

$$b_{2i} - a_{2i} = f'(R) \quad (3 a)$$

and

$$b_{2i-1} = f''(R). \quad (3 b)$$

Now, in the case of a low-pass lossless ladder network, which is singly terminated in resistance, it can easily be shown that odd powered terms of the reciprocal voltage transfer functions are linear functions of R while its even-powered terms are independent of R . Consequently relations (3 a) and (3 b) modify to

$$b_{2i} = a_{2i} \quad (4 a)$$

and

$$b_{2i-1} = \alpha_{2i-1} R \quad (4 b)$$

where α_{2i-1} is a positive real constant chosen so that b_i obey the Hurwitz restriction.

Thus starting with a prescribed LC transfer function $P_n(s)$, one can write $P_n'(s)$ by an appropriate selection of α_{2i-1} since R need not be ascribed any numerical value.

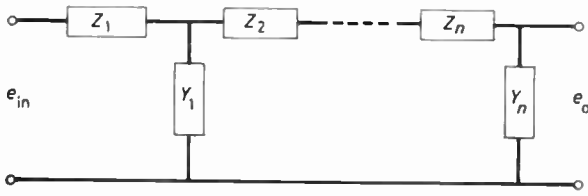


Fig. 1. Ladder network N.

Once $P_n'(s)$ has been formed, the corresponding LCR network N' may be obtained by the Holbrook technique¹ and the removal of the resistive element R will yield the desired LC network N . It is to be pointed out that α_{2i-1} may possess a number of sets of values and correspondingly the network elements will also have a number of alternative sets of values.

Illustration

Let the prescribed transfer function $T(s)$ be

$$T(s) = P_6(s) = s^6 + 5s^4 + 6s^2 + 1. \quad (5)$$

A number of sets of values of α_{2i-1} and the corresponding values of network elements are given in Table 1.

$$M(y) = \begin{vmatrix} x_1 y & 1 & 0 & \dots & 0 & 0 & 0 \\ -1 & x_2 y & 1 & \dots & 0 & 0 & 0 \\ 0 & -1 & x_3 y & \dots & 0 & 0 & 0 \\ \dots & \dots & \dots & \dots & \dots & \dots & \dots \\ 0 & 0 & 0 & \dots & x_{n-2} y & 1 & 0 \\ 0 & 0 & 0 & \dots & -1 & x_{n-1} y & 1 \\ 0 & 0 & 0 & \dots & 0 & -1 & x_n y \end{vmatrix} \quad (6)$$

A simultaneous division of the odd rows and multiplication of the even columns by y and a subsequent reverse transformation $y^2 = s$ yields the following continuant:

$$Q(s) = \begin{vmatrix} x_1 & 1 & 0 & \dots & 0 & 0 & 0 \\ -1 & x_2 s & 1 & \dots & 0 & 0 & 0 \\ 0 & -1 & x_3 & \dots & 0 & 0 & 0 \\ \dots & \dots & \dots & \dots & \dots & \dots & \dots \\ 0 & 0 & 0 & \dots & x_{n-2} s & 1 & 0 \\ 0 & 0 & 0 & \dots & -1 & x_{n-1} & 1 \\ 0 & 0 & 0 & \dots & 0 & -1 & x_n s \end{vmatrix} \quad (7)$$

The elements of the main diagonal of this continuant (relation 7) provide the values of the elements of the desired RC ladder network.

Table 1. Element values for LC transfer functions.

α_5	α_3	α_1	L_1	C_1	L_2	C_2	L_3	C_3
1.00	3.00	2.00	1.00	0.50	2.00	1.00	2.00	0.50
1.50	4.50	2.00	0.67	0.75	2.00	0.46	2.73	0.78
1.50	6.00	4.50	0.66	1.50	0.66	1.50	0.66	1.50
4.00	16.00	10.00	0.25	4.00	0.50	4.00	0.25	2.00
1.00	4.00	3.00	1.00	1.00	1.00	1.00	1.00	1.00

Table 2. Element values for RC transfer functions.

α_5	α_3	α_1	R_1	C_1	R_2	C_2	R_3	C_3
2.00	5.00	2.60	3.00	2.00	1.66	0.50	12.00	0.10
5.00	8.00	2.00	1.20	0.78	3.10	0.54	5.67	0.68
6.00	10.00	3.00	1.00	1.00	2.00	1.00	3.00	1.00
8.00	10.00	2.00	0.75	0.94	4.25	0.50	7.15	0.56

Synthesis of RC Ladders

An important advantage of this synthesis procedure is that it enables the realization of RC ladders (Fig. 1, where $Z_i = R_i$, $Y_i = C_i s$) as well.

If $Q(s)$ be the prescribed RC transfer function then a transformation $S = y^2$ will yield a function $M(y)$, having its roots on the imaginary axis of the complex y -plane. That is, it will be an LC function in y . Consequently one can obtain the following continuant by the method outlined above:

Illustration

Let the prescribed RC transfer function be

$$Q(s) = 6s^3 + 16s^2 + 10s + 1 \quad (8 a)$$

so that

$$M(y) = 6y^6 + 16y^4 + 10y^2 + 1. \quad (8 b)$$

The selection of various sets of values for the constants α_{2i-1} ($i = 1, 2, 3 \dots$) leads to a number of sets of element values which are given in Table 2.

It should be pointed out that the sets of element values presented in Tables 1 and 2 for the prescribed LC and RC transfer functions are only representative. Alternative sets of element values may also be found. The technique may thus facilitate the synthesis of an optimized LC (or RC) ladder network, taking the total sum of any one type of element as one of the optimization criterion.

Conclusion

A method for the synthesis of a purely reactive ladder has been proposed and the method has been extended, by the use of a frequency transformation to the RC network. The method yields a large number of sets of values for the network elements. This feature of the technique may help in identifying a network in which the total sum of the inductive or capacitive elements is minimum. The method does not require the solution of any non-linear equations and can be carried out quite easily.

The Authors



Shyama Charan Prasad studied at L.S. College, University of Bihar, Muzaffarpur, for his B.Sc., M.Sc. and Ph.D. degrees in physics which he received in 1962, 1964 and 1976 respectively. He served temporarily as a Lecturer in Physics at L.S. College from January 1965 to May 1966 and then joined Bhagalpur University as a Lecturer in July 1966. He was appointed a Lecturer in Physics at L.S. College in 1967

and since last autumn he has held a Post Doctoral Fellowship in the Electrical Engineering Department, Indian Institute of

Acknowledgment

Dr. S. C. Dutta Roy, Department of Electrical Engineering, Indian Institute of Technology, Delhi, has been a constant source of encouragement to us. We thank him, with deep sense of gratitude, for many helpful discussions and recommendations.

References

- 1 Holbrook, J. G., 'The recurrent-continuant method of transfer function synthesis', *The Radio and Electronic Engineer*, **38**, No. 2, pp. 73-9, August 1969.
- 2 Jha, D. K. and Prasad, S. C., 'The synthesis of ladder networks with resistances at both ends by the recurrent-continuant method', *The Radio and Electronic Engineer*, **46**, No. 8/9, pp. 421-8, August-September 1976.

*Manuscript first received by the Institution on 20th September 1976 and in final form on 4th February 1977.
(Short Contribution No. 188/CC273.)*

© The Institution of Electronic and Radio Engineers, 1977

Technology, New Delhi. Dr. Prasad's research fields are network theory and filter design.



Yugal Kishore Yadav received his B.Sc. honours and M.Sc. degree in physics from L.S. College, University of Bihar, Muzaffarpur, in 1962 and 1964 respectively. He served temporarily as a lecturer in Physics at T.P. College, Mathipura, in Mithila University from 1965 to 1967, and then joined Bihar University as a lecturer in R.D.S. College, Muzaffarpur, the post he still holds. His earlier interest

was in the field of X-rays and he is now pursuing research work in the field of circuit theory.

The temperature dependence of the low-frequency characteristics of junction field-effect transistors

J. B. GRIMBLEBY, Ph.D., C.Eng., M.I.E.E.*

and

G. S. SANGHERA, Ph.D.*

SUMMARY

A discussion is given of the factors determining the temperature dependence of the low-frequency junction field-effect transistor characteristic. The results of measurements of the temperature coefficient of the pinch-off voltage and the channel mobility for a number of 2N3819 and 2N3820 transistors are presented and these are seen to be in disagreement with values that have been assumed previously. The theoretical and practical considerations in obtaining a zero temperature coefficient are discussed in some detail.

* J. J. Thomson Physical Laboratory, University of Reading, Whiteknights, Reading, Berkshire RG6 2AF.

1 Introduction

In view of the importance of the junction field-effect transistor (j.f.e.t.) as an input device in linear amplifiers and as a variable resistance element in voltage controlled attenuators it is surprising how little is known about the variation of its $I-V$ characteristic with temperature.

Two statements which appear frequently in the technical literature and in text books are the following:

- (i) The temperature variation of the pinch-off voltage V_p is expressed by the relation

$$\frac{d|V_p|}{dT} = 2.2 \text{ mV/K.} \quad (1)$$

- (ii) Under conditions of drain saturation a zero temperature coefficient point, that is a point where

$$\left[\frac{dV_{GS}}{dT} \right]_{I_b} = 0, \quad \left[\frac{dI_D}{dT} \right]_{V_{GS}} = 0,$$

occurs at a gate-source voltage V_{GSZ} given by

$$|V_{GSZ}| = |V_p| - 0.63 \text{ V.} \quad (2)$$

The reference usually given for equation (1) is Evans.¹ Other authors, for example Gosling,² Sevin,³ Hoerni and Weir,⁴ and Todd⁵ give values of $d|V_p|/dT$ ranging from 2.0 to 2.5 mV/K but in none of these cases do the authors state how they were obtained.

Steven⁶ calculated the temperature coefficient of the space-charge voltage of the gate-channel diode on the basis of an estimated doping level of 10^{21} m^{-3} in the device channel and obtained the value of 2.1 mV/K. A certain amount of confusion has arisen because of the statements in the literature^{1,3,8} to the effect that the change in V_p with temperature is the same phenomenon as the change in the base-emitter voltage, V_{BE} , of a bipolar transistor at constant collector current. This is not the case as can be seen by comparing the expression for the bipolar temperature coefficient:

$$\left[\frac{dV_{BE}}{dT} \right]_{I_c} = \frac{(V_{BE} - V_{GO} - 3V_T)}{T} \quad (3)$$

(where V_{GO} is the energy gap at absolute zero and is equal to 1.205 V for silicon) with the expression for the temperature coefficient of V_p obtained by assuming that the doping profile across the gate junction is abrupt:

$$\frac{d|V_p|}{dT} = \frac{d|V_B|}{dT} = \left| \frac{V_B - V_{GO} - 3V_T}{T} \right| \quad (4)$$

where V_B is the built-in voltage, and $V_T = kT/q$ is the thermal voltage. (k is Boltzmann's constant and q is the electronic charge.)

As can be seen expressions (3) and (4) are different; under certain conditions however, the numerical values of dV_{BE}/dT and $d|V_p|/dT$ can be similar.

Equation (2) was apparently first given by Evans.¹ Evans used the value of $d|V_p|/dT$ given in equation (1) together with a temperature coefficient of carrier mobility

of $-7.0 \times 10^{-3}/K$, and assumed a square law form for the drain current I_D against gate-source voltage V_{GS} characteristic:

$$I_D = I_{DSS}(1 - V_{GS}/V_p)^2.$$

Trofimenkoff and Cobbold⁷ also obtained equation (2). They assumed that $d|V_p|/dT$ is given by expression (4) with a value of 0.65 V substituted for V_B and they made a theoretical estimation of the temperature coefficient of mobility, by using the Johnson and Lark-Horovitz⁹ method for combining the theoretical lattice scattering mobility and impurity scattering mobility. By taking an estimated impurity level of 10^{22} m^{-3} they obtain a figure of $-7.02 \times 10^{-3}/K$ for $1/\mu \text{ d}\mu/\text{d}T$.

Sevin³ calculated that the zero-temperature-coefficient point should occur at a gate-source voltage of $|V_p| - 0.60 \text{ V}$. His calculations were based on the incorrect assumption that the temperature coefficient of I_{DSS} is equal to the temperature coefficient of mobility and that the value for V_p can be evaluated from the measurements of I_{DSS} and g_{m0} (the transconductance at I_{DSS}). These authors did not offer any direct experimental verification of their results.

In order to check these statements experimentally a number of n-channel and p-channel transistors of the well-known types 2N3819 and 2N3820 were randomly selected for measurement.

The determination of V_{GSZ} is a lengthy procedure but presents little practical difficulty. On the other hand, in spite of the fact that V_p is one of the two basic parameters characterizing a j.f.e.t., it is by no means obvious how to go about measuring this quantity.

2 The Pinch-off Voltage

It can be shown that, taking the depletion approximation to be valid and by using the 'gradual channel' approximation, the saturated characteristic of a j.f.e.t. is of the general form:

$$I_D = \sigma f(V_{GS} - V_B) \quad (5)$$

where σ is the channel conductivity and $f(V)$ is a function which contains only the variable V and constants. It is recognized, of course, that the 'gradual channel' approximation is not universally applicable. However, in the case of most general purpose j.f.e.t.s, and in particular in the case of 2N3819 and 2N3820 transistors (where the channel length is approximately four times the channel depth), the errors resulting from the use of the approximation are likely to be small. At room temperature the doped impurities can be taken to be fully ionized and therefore the carrier concentration can be assumed to be constant, implying that σ is simply proportional to μ .

The pinch-off voltage V_p can be defined by the relation:

$$f(V_p - V_B) = 0. \quad (6)$$

Thus $(V_p - V_B)$ is a temperature independent quantity and equation (5) may be written:

$$I_D = \mu g(V_{GS} - V_p) \quad (7)$$

where the function $g(V)$ is related to $f(V)$ by:

$$g(V) = n_0 q f(V + V_p - V_B) \quad (8)$$

where n_0 is the carrier concentration. $g(V)$, like $f(V)$, contains only the variable V and constants and has the additional property that:

$$g(0) = 0. \quad (9)$$

Unfortunately equations (7) and (9) do not as they stand provide the basis for a practical method of measuring V_p because they are only intended to describe the behaviour of the device under conditions where the depletion approximation is valid. When the device is biased close to V_p the channel width becomes of the same order as the width of the depletion layer edge and the depletion approximation cannot be used. It is therefore necessary to use some specific form for the characteristic so that V_p can be deduced from measurements made at fairly high currents.

Whilst in principle any function which is of the general form of equations (7) and (9) can be used provided that it fits the measured characteristic over the range of validity of the depletion approximation, this may lead to values of V_p that would not normally be considered acceptable. Consequently it is necessary for the purpose of determining the value of V_p (but not, as will be shown, for determining its temperature coefficient) to introduce an additional restriction on the function $g(V)$: the function should be based on some physically realistic model of the behaviour of the channel width.

Even with the above restriction, however, the quantity V_p for a given device does not have a single correct value but depends on the particular form of the characteristic chosen.

Experimentally it was found that if I_D/g_m (where $g_m = \partial I_D / \partial V_{GS}$) was plotted against V_{GS} for the 2N3819 and 2N3820 devices a straight line was obtained except at currents smaller than about $I_{DSS}/20$. A typical example is shown in Fig. 1. The characteristic can therefore be represented by the equation (which meets all of the requirements stated above):

$$I_D = I_{DSS}(1 - V_{GS}/V_p)^n \quad \text{for } |V_{GS}| < |V_p| \\ = 0 \quad \text{for } |V_{GS}| \geq |V_p|. \quad (10)$$

This is equivalent to the expression:

$$\frac{I_D}{g_m} = (V_{GS} - V_p)/n. \quad (11)$$

A straight line fitted to a graph of I_D/g_m against V_{GS} will have a slope of $1/n$ and an intercept on the V_{GS} axis of V_p . A typical example of such a graph for a 2N3819 j.f.e.t. is shown in Fig. 1. A straight line is obtained over a wide range of I_D with the graph, as expected, deviating from this line close to pinch-off, where the depletion approximation is no longer valid.

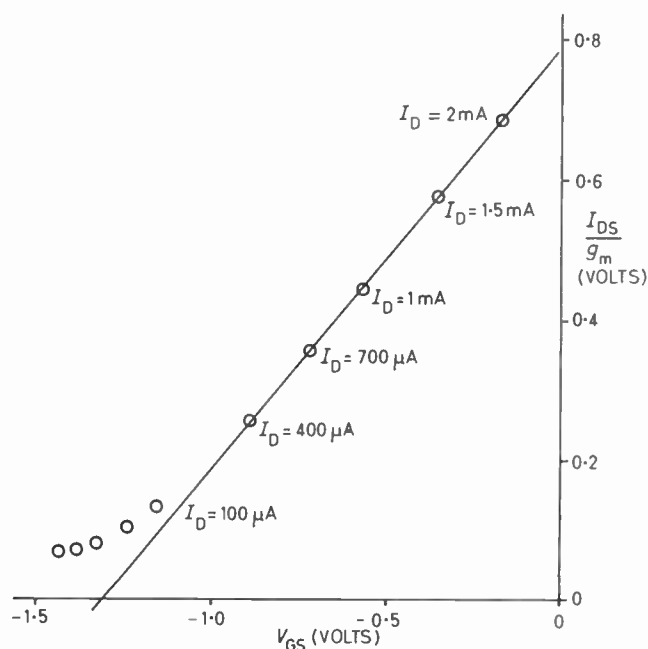


Fig. 1. A graph showing the variation of I_D/g_m with V_{GS} for a 2N3819 transistor.

3 Results of Measurement of V_{GSZ} and V_p

The results of measurements on five n-channel transistors (2N3819) and five p-channel transistors (2N3820) are shown in Table 1.

Taking first the case of the p-channel devices, it can be seen that V_{GSZ} varies widely with respect to V_p , contrary to the previous statements in the literature. An interesting result of the measurements on the p-channel devices is that if V_p is obtained from the extrapolation of the straight line section of the I_D/g_m against V_{GS} graph, we find that for some devices V_{GSZ} occurs at a higher voltage than the pinch-off voltage. However, at these gate-source biases the j.f.e.t. device cannot be employed effectively in a circuit because of the low drain current.

In the case of the n-channel transistors the quantity $|V_p| - |V_{GSZ}|$ is fairly constant with values over the range 0.42 V to 0.52 V and none of these results are close to

Table 1

No.	V_{GSZ}	V_p	$ V_p - V_{GSZ} $	n
2N3819:				
	(V)	(V)	(V)	
1	-1.43	-1.86	0.43	1.65
2	-1.09	-1.58	0.49	1.69
3	-0.81	-1.33	0.52	1.68
4	-1.92	-2.34	0.42	1.65
5	-1.75	-2.17	0.42	1.62
2N3820:				
	(V)	(V)	(V)	
1	2.22	2.49	0.27	1.82
2	5.03	4.78	-0.25	2.14
3	5.46	5.31	-0.15	1.86
4	3.12	3.45	0.33	1.67
5	4.28	4.19	-0.09	2.04

those quoted by Evans.¹ These measurements suggest, therefore, that in previous publications incorrect values for dV_p/dT and $1/\mu d\mu/dT$ have been used in the determination of V_{GSZ} .

4 The Temperature Coefficients of V_p and μ

It can readily be shown that the temperature coefficient of V_p is equal to the temperature coefficient of V_B , the built-in voltage of the gate-channel p-n junction. However the temperature coefficient of V_B is dependent on the type of impurity profile across the junction. dV_B/dT for an abrupt junction is given by:

$$\frac{dV_B}{dT} = \frac{(V_B - V_{GO} - 3V_T)}{T} \tag{12}$$

and for a linearly-graded junction is given by:

$$\left\{ 1 - \frac{2V_T}{3V_B} \right\} \frac{dV_B}{dT} = \frac{(V_B - V_{GO} - 3V_T)}{T} \tag{13}$$

However, these expressions cannot be evaluated because V_B cannot be measured directly, nor can it be deduced from the doping profiles because these are not known to sufficient accuracy.

In principle it should be possible to determine dV_p/dT by measuring V_p (using the method described above) at a number of temperatures. However, it is found that the errors in determining V_p are such that when two values of V_p of the order of a few volts are subtracted to obtain the temperature coefficient, the error in the result is unacceptably large. An additional problem is that n must be constrained to be temperature independent. An alternative method was therefore developed.

Differentiating I_D with respect to temperature gives:

$$\frac{dI_D}{dT} = \frac{\partial I_D}{\partial \mu} \frac{d\mu}{dT} + \frac{\partial I_D}{\partial V_{GS}} \frac{dV_{GS}}{dT} + \frac{\partial I_D}{\partial V_p} \frac{dV_p}{dT} \tag{14}$$

From equation (7) we see that:

$$\frac{\partial I_D}{\partial \mu} = \frac{I_D}{\mu} \tag{15}$$

and

$$\frac{\partial I_D}{\partial V_{GS}} = \frac{-\partial I_D}{\partial V_p} = g_m \tag{16}$$

Thus

$$\frac{dI_D}{dT} = \frac{I_D}{\mu} \frac{d\mu}{dT} + g_m \frac{dV_{GS}}{dT} - g_m \frac{dV_p}{dT} \tag{17}$$

This equation can be written in the form:

$$\frac{dV_{GS}}{dT} = \frac{I_D}{g_m} \frac{dI_D}{dT} - \frac{I_D}{g_m \mu} \frac{d\mu}{dT} + \frac{dV_p}{dT} \tag{18}$$

If one considers the case where I_D is constant, then equation (18) becomes:

$$\left[\frac{dV_{GS}}{dT} \right]_{I_D} = \frac{I_D}{g_m \mu} \frac{d\mu}{dT} + \frac{dV_p}{dT} \tag{19}$$

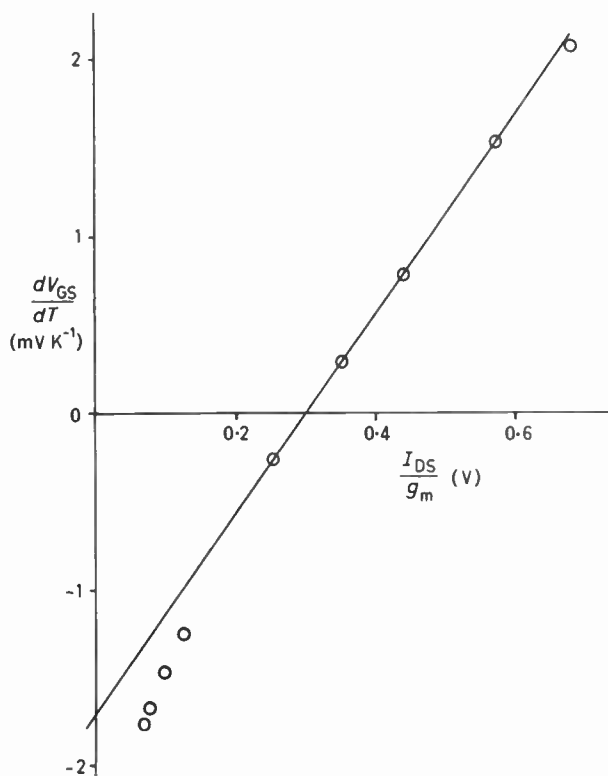


Fig. 2. A graph of dV_{GS}/dT against I_D/g_m with I_D as the varying parameter.

A straight line fitted to a graph of $[dV_{GS}/dT]_{I_D}$ plotted against I_D/g_m with I_D as a parameter will then have a slope of $1/\mu \cdot d\mu/dT$ and an intercept on the $[dV_{GS}/dT]_{I_D}$ axis of dV_p/dT .

In Fig. 2 is shown the graph obtained from a typical 2N3819 transistor. It can be seen that there is a well-defined straight line part of the graph at the higher currents, the graph deviating from this straight line at low currents.

The results of measuring the coefficients of mobility and V_p for the 2N3819 and 2N3820 transistors are given in Table 2.

Values of $d|V_p|/dT$ for both n-channel and p-channel devices are considerably different from the values of

Table 2

	$d V_p /dT$	$1/\mu \cdot d\mu/dT$
	(mV/K)	($\times 10^{-3}/K$)
2N3819:		
1	1.60	-5.57
2	1.68	-5.47
3	1.78	-5.74
4	1.66	-5.56
5	1.68	-5.59
2N3820:		
1	1.59	-6.85
2	1.30	-6.45
3	1.21	-6.32
4	1.42	-6.54
5	1.28	-6.50

2.0 to 2.5 mV/K which have appeared in the literature and which were used to derive equation (2).

The values for the temperature coefficient of mobility lie in the range of -5.47 to $-5.74 \times 10^{-3}/K$ for n-channel and -6.32 to $-6.85 \times 10^{-3}/K$ for p-channel devices. These results can be compared with those obtained by Bullis *et al.*¹⁰ who evaluated the temperature coefficients of mobility in bulk material over a wide range of doping levels. At the doping levels present in our devices (approximately 10^{21} m^{-3} to 10^{22} m^{-3}) they obtained values of $-6.9 \pm 0.3 \times 10^{-3}/K$ and $-7.3 \pm 0.3 \times 10^{-3}/K$ for n-type and p-type material respectively.

5 The Zero Temperature Coefficient Point V_{GSZ}

From equation (19) the following condition for the zero temperature coefficient point can be derived:

$$I_D/g_m = V_0 \tag{20}$$

where

$$V_0 = \frac{dV_p/dT}{1/\mu \cdot d\mu/dT} \tag{21}$$

Substituting the empirical form of the characteristic [eqn. (10)] into this equation gives an expression for V_{GSZ} :

$$V_{GSZ} = V_p + nV_0 \tag{22}$$

If the average measured temperature coefficients of μ and V_p are inserted into this expression the results shown in Table 3 are obtained.

Table 3

	$\frac{1}{\mu} \frac{d\mu}{dT}$	$\frac{d V_p }{dT}$	n	$ V_0 $	$ V_p - V_{GSZ} $
	($\times 10^{-3}/K$)	(mV/K)		(V)	(V)
n-channel	-5.58	1.68	1.66	0.30	0.50
p-channel	-6.53	1.38	1.87	0.21	0.40

In the case of the n-channel devices the calculated value of $|V_p| - |V_{GSZ}|$ shown in Table 3 agrees with the measured values given in Table 1 to within 0.1 V. The value of 0.40 V predicted for the p-channel devices was not realized in practice because it places the zero temperature coefficient point at a part of the characteristic not represented by equation (8).

In any practical application we are interested, not directly in how close the operating point is to V_{GSZ} , but in the magnitude of the resulting temperature coefficient. Combining the general expression for dV_{GS}/dT [eqn. (19)] with equation (10) gives the result:

$$\left[\frac{dV_{GS}}{dT} \right]_{I_D} = \frac{1}{n} \cdot (V_{GS} - V_p) \cdot \frac{1}{\mu} \cdot \frac{d\mu}{dT} + \frac{dV_p}{dT} \tag{24}$$

$$= \frac{1}{n} \cdot (V_{GS} - V_{GSZ}) \cdot \frac{1}{\mu} \cdot \frac{d\mu}{dT} \tag{25}$$

An error of 0.1 V in the value of V_{GS} will therefore lead to a temperature coefficient of about 300 $\mu\text{V}/\text{K}$ in V_{GS} .

For comparison we use equation (24) to calculate dV_{GS}/dT at other bias points. For example, if we consider the case of a transistor with a V_p of 2 V operating at approximately zero gate-source bias, then at constant I_D the gate-source voltage will have a temperature coefficient of about 5 mV/K.

6 Zero Temperature Coefficient of g_m

The discussion has so far concentrated on the zero temperature coefficient point when operated under drain-saturated conditions. In many applications, however, it is the stability of g_m with temperature, at constant I_D , that is of primary importance and shifts in, for example, V_{GS} are of little consequence.

By differentiating equation (7) we see that g_m is of the form:

$$g_m = \mu h(V_{GS} - V_p) \tag{26}$$

where $h(V)$ is the derivative of $g(V)$. This expression is of the same form as equation (7) and following the derivation of equation (22) for V_{GSZ} we find that the zero temperature coefficient point for g_m occurs at a bias voltage V'_{GSZ} given by:

$$V'_{GSZ} = V_p + (n-1)V_0. \tag{27}$$

Substituting the averaged values of V_0 and n into this expression gives a value of $|V'_{GSZ}|$ 0.2 V below $|V_p|$ for n-channel and 0.18 V below for p-channel devices. These values give an operating point that is very close to pinch-off and in both cases V'_{GSZ} is at a point on the characteristic where the depletion approximation is not valid. Unfortunately, therefore, in the case of the devices we measured, equation (27) is of no practical use in obtaining a zero temperature coefficient for g_m .

7 Practical Considerations in Determining the Zero Temperature Coefficient Point

It is difficult to utilize the V_{GSZ} point for a j.f.e.t. because of the problem of finding the value of V_p to be used, as it is clearly not practicable to make a series of measurements and plot a graph for each device. The method used by most commercial j.f.e.t. testers is to measure the value of V_{GS} at a current that is known to be very much less than I_{DSS} ; typically a value of 10 μA is used. This technique, while being very convenient and quick to apply, results in values of V_p which differ from those obtained by fitting an empirical characteristic by as much as 0.5 V. In most cases this is insufficiently accurate to give a satisfactory temperature coefficient.

Fortunately another biasing method is available which relies less on the precision of measurement of V_p . If instead of setting the bias point by adjusting V_{GS} relative to V_p the drain current I_D is controlled at I_{DZ} , which is the value of I_D calculated by substituting eqn. (22) into the empirical form for the characteristic [eqn. (10)] the

sensitivity to errors in V_p is very much reduced. I_{DZ} is given by:

$$I_{DZ} = I_{DSS} \left(\frac{nV_0}{V_p} \right)^n. \tag{28}$$

Now, defining the sensitivity coefficient S_x^y as:

$$S_x^y = \frac{x}{y} \frac{dy}{dx} \tag{29}$$

then if the bias point is set by making V_{GS} equal to $V_p + nV_0$, and assuming that $V_0 \ll V_p$, we can write:

$$S_{V_p}^{V_{GS}} \approx 1. \tag{30}$$

If, however, I_D is set to the value given by eqn. (28) then the sensitivities of I_D to the values of V_p and I_{DSS} are:

$$S_{V_p}^{I_D} \approx -n \tag{31}$$

$$S_{I_{DSS}}^{I_D} \approx 1. \tag{32}$$

In order to compare these sensitivities with eqn. (30) they must be put in terms of V_{GS} using eqn. (10):

$$S_{V_p}^{V_{GS}} = S_{I_D}^{V_{GS}} S_{V_p}^{I_D} \approx \frac{V_0}{V_p} (-n) \approx \frac{-nV_0}{V_p} \tag{33}$$

$$S_{I_{DSS}}^{V_{GS}} = S_{I_D}^{V_{GS}} S_{I_{DSS}}^{I_D} \approx \frac{V_0}{V_p} \approx \frac{V_0}{V_p}. \tag{34}$$

From these results it can be seen that, while it is necessary to measure an additional parameter of the device, the bias point is less sensitive to the accuracy of measurement of V_p by a factor of V_0/V_p which is of the order 0.1, say. The additional parameter, I_{DSS} , is easy to determine and is only required to a moderate degree of accuracy.

Consider a typical j.f.e.t. with a V_p of approximately 2 volts and suppose that V_p and I_{DSS} are both measured to 10%. If the device is biased at a current given by eqn. (28) the error in V_{GS} (relative to V_{GSZ}) is 0.05 V corresponding to a temperature coefficient of less than 150 $\mu\text{V}/\text{K}$.

8 Conclusion

It has been shown that it is practicable to make use of the zero temperature coefficient point in certain types of j.f.e.t.s, namely those types in which the saturated characteristic can be fitted to an empirical form to within, say, 0.2 V of pinch-off. In such devices fairly rough measurements of I_{DSS} and V_p can be used to calculate a bias point at which the magnitude of the temperature coefficient of V_{GS} will be smaller by a factor of 20 than might otherwise be obtained.

Uncertainties in the values of the temperature coefficient of μ and V_p and the difficulties of determining V_p limit the use of zero temperature coefficient biasing to fairly undemanding applications where a temperature coefficient of V_{GS} of, say 300 $\mu\text{V}/\text{K}$ can be tolerated.

In devices where the zero temperature coefficient point cannot be predicted the measurements of dV_p/dT and $1/\mu \cdot d\mu/dT$ are nevertheless of considerable interest because

they permit accurate calculation of dV_{GS}/dT at operating points remote from pinch-off.

The temperature coefficients of V_p that we measured were significantly different from previously published results, particularly in the case of the p-channel transistors.

9 References

- 1 Evans, L. L., 'Biasing f.e.t.s for zero d.c. drift', *Electro-technology*, 7, pp. 93-6, August 1964.
- 2 Gosling, W., Townsend, W. G. and Watson, J., 'Field-Effect Electronics' (Butterworths, London, 1971).
- 3 Sevin, L. J., 'Field-Effect Transistors' (McGraw-Hill, New York, 1965).
- 4 Hoerni, J. A. and Weir, G., 'Conditions for a temperature compensated silicon f.e.t.', *Proc. IEEE*, 5, pp. 1058-9, July 1963. (Correspondence.)
- 5 Todd, C. D., 'Silicon epitaxial f.e.t.s', *Electronic Components* 7, pp. 571-6, June 1966.
- 6 Stevens, M., 'The FET in Direct Coupled Amplifier Circuits', Texas Instruments Application Report B.51.
- 7 Cobbold, R. S. C. and Trofimenkoff, F. N., 'Theory and application of the field-effect transistor', *Proc. Instn Elect. Engrs*, 111, pp. 1981-92, 1964.
- 8 'Temperature Dependence of JFET Parameters', Mullard Application Note TP1112, 1969.
- 9 Johnson, V. A. and Lark-Horovitz, K., 'The combination of resistivities in semiconductors', *Phys. Rev.*, 82, pp. 977-8, 1951.
- 10 Bullis, W. M., Brewer, F. H., Kolstad, C. D. and Swartzendruber, L. J., 'Temperature coefficients of resistivity of silicon and germanium near room temperature', *Solid State Electronics*, 11, pp. 639-46, 1968.

Manuscript first received by the Institution on 7th January 1977 and in final form on 21st March 1977 (Paper No. 1771/CC274).

© The Institution of Electronic and Radio Engineers, 1977

The Authors



Gurbakhash Sanghera received the B.Sc. degree in physics from the University of Reading in 1972 and he then carried out research in the field of semiconductor devices and was awarded a Ph.D. in 1976. In 1975 he moved to the Department of Computer Science at Reading as a research fellow and since then has been working on the Monte-Carlo simulation of GaAs devices, in particular m.e.s.f.e.t.s.



Jim Grimbleby graduated in physics at Reading University in 1965 and for the next four years he carried out research into electron spin resonance and spin-lattice relaxation in S-state rare-earth ions; he was awarded a Ph.D. in 1970. Since completing his doctoral research his interests have been more in the field of electronics and in 1971 he received his present appointment as a lecturer at the J. J. Thomson

Physical Laboratory, University of Reading.



Future Telecommunications Services

A. A. L. REID, M.A., M.Sc., Ph.D.

Based on a presentation at the Institution's Golden Jubilee Convention held in Cambridge from 28th June to 1st July 1976.

A classification of existing and future electronic communication services is made and the interactions among the electronic communications services, and between them and the non-electronic media are discussed. Some policy issues arising from these trends are identified.

During the last 50 years the telephone has become far and away the most important telecommunications service, and it has come to dominate the activities of telecommunications administrations such as the British Post Office. However, in the last ten years there has begun to emerge a bewilderingly wide range of new telecommunications services, whose growth is likely to be the most spectacular feature of the next fifty years of telecommunications.^{1, 2, 3}

These new services include computer communications, facsimile, mobile telephone and paging, videotelephone, tele-conference services, Viewdata, cable television, remote metering, surveillance, and alarm services. It is easy enough to recite a shopping list of such services; the purpose of this paper is to classify them in ways which clarify the general trends of which these particular services are specific manifestations.

This classification is applied to the main existing electronic communication services in Fig. 1 (dealing with the wireless services) and Fig. 2 (dealing with the wired services). These two groups of existing services are at present very separate. Because they occupy opposite corners of the classification matrix they are not in competition with each other. They are clearly different in the degree to which they are interactive; whereas telephony services involve rapid exchange of messages between participants, broadcast services involve the one-way flow of information. In the case of the UK (and most other countries) they are clearly different in their institutional embodiment—the telecommunications and broadcasting authorities being separate. Finally, they are clearly different in the nature of the content which they carry. The telephone and telegraph services have amateur content (in the sense that

	NUMBER OF RECIPIENTS							
	1	1-10	10-10 ²	10 ² -10 ³	10 ³ -10 ⁴	10 ⁴ -10 ⁵	10 ⁵ -10 ⁶	10 ⁶ -10 ⁷
ALPHA-NUMERIC & GRAPHIC								
AUDIO					BROADCAST SOUND			
AUDIO-VISUAL					BROADCAST TELEVISION			

Fig. 1. Main existing wireless services.

	NUMBER OF RECIPIENTS							
	1	1-10	10-10 ²	10 ² -10 ³	10 ³ -10 ⁴	10 ⁴ -10 ⁵	10 ⁵ -10 ⁶	10 ⁶ -10 ⁷
ALPHA-NUMERIC & GRAPHIC		TELEX DATA FAX						
AUDIO		TELE- PHONE						
AUDIO-VISUAL								

Fig. 2. Main existing wired services.

The classification scheme is based on three dimensions:

- (a) The *form* of the message, i.e. whether it is audio-visual (e.g. television), audio (e.g. telephone), or alpha-numeric or graphic (as is the case with any written or printed material).
- (b) The *number* of people who receive the same message (not necessarily simultaneously). This may vary from one to many millions.
- (c) The principal means of *transmission*, i.e. wired (as in the case of the telephone) or wireless (as in broadcast radio or television).

telephone conversations are not professionally scripted and produced), whereas the broadcasting services require for their success not only the technical means of distribution, but also an elaborate and highly professional apparatus of programme production.

In order to grasp the likely future trends in electronic communication services it is useful to look at the rich variety of non-electronic media, i.e. those which involve face-to-face gatherings or the transport of stored information in the form of letters, memos, reports, books, magazines, newspapers, gramophone records, films and cassettes. These non-electronic media are set out, using the same classification scheme, in Fig. 3.

It is in principle possible to devise electronic alternatives for each of these existing media. Why has this not happened already? In some cases (e.g. audio-visual services as substitutes for face-to-face meetings) it is because the electronic means require wide transmission bandwidths which cannot be shared simultaneously among many users, and which are therefore very expensive.⁴ In other cases (e.g. books,

Dr. A. A. L. Reid is a Deputy Director of Post Office Telecommunications Headquarters with responsibility for Long Range Studies. He was formerly Director of the Communications Studies Group at University College London, where he was granted his Ph.D. for research on the substitution of telecommunications for face-to-face contact. He is Chairman of the Long Term Studies Working Group of the European Conference of Posts and Telecommunications Administrations.

	NUMBER OF RECIPIENTS							
	1	1-10	10-10 ²	10 ² -10 ³	10 ³ -10 ⁴	10 ⁴ -10 ⁵	10 ⁵ -10 ⁶	10 ⁶ -10 ⁷
ALPHA-NUMERIC & GRAPHIC	LETTER	MEMORANDA + REPORTS		CIRCULARS + SPECIALIST PUBLICATIONS		BOOKS MAGAZINES AND NEWSPAPERS		
AUDIO					GRAMOPHONE RECORDS AND AUDIO CASSETTES			
AUDIO-VISUAL	FACE-TO-FACE CONVERSATION AND MEETINGS			CINEMA, THEATRE, PUBLIC GATHERINGS AND VIDEO CASSETTES				

Fig. 3. Communication media not involving electronic transmission.

magazines, and newspapers) the conventional technology is highly developed, highly satisfactory and still extremely cheap. In yet other cases (such as cable television) market forces may be heavily constrained by Government regulation.⁵ But as the cost of electronic technology falls, we can expect to see electronic media expanding to offer the full range of services in terms of form of message and size of audience, and we can expect the electronic alternative to be increasingly competitive, in economic terms, with the previous means.⁶

This development of electronic communication services will in many cases be closely tied up with the development of information storage and processing techniques. A good example is that of the communicating word-processing typewriter, which can be used to store text, to edit it, and to transmit it over a distance. However, the main concern of this paper is with the transmission of information, rather than its storage or processing.

Consider first the future trends in the wireless services (Fig. 4). The wireless services seem likely to expand into the third form of message (alpha-numeric and graphic) with Ceefax and Oracle.⁷ When there are only a few hundred or a few thousand teletext decoders, the services will lie in the 10² to 10³ part of the matrix, shifting to the right as the number of decoders increases. As the service grows, the limited space available on conventional television channels will have to be used for the most popular messages; more specialized markets will have to be met by 'high capacity Ceefax' (where a substantial part or all of a television channel is devoted to Ceefax) or by wired systems discussed below. There are also signs of wireless sound services extending down the audience size range in the form of Citizens' Band radio of the type which has recently grown so rapidly in the USA.⁸

Turning to the wired services, their future trends are illustrated in Fig. 5. Starting with the existing telex, data, and

	NUMBER OF RECIPIENTS							
	1	1-10	10-10 ²	10 ² -10 ³	10 ³ -10 ⁴	10 ⁴ -10 ⁵	10 ⁵ -10 ⁶	10 ⁶ -10 ⁷
ALPHA-NUMERIC & GRAPHIC				BROADCAST CEEFAX AND ORACLE				
AUDIO			CITIZENS BAND RADIO	BROADCAST SOUND				
AUDIO-VISUAL					BROADCAST TELEVISION			

Fig. 4. Existing and new wireless services.

Table 1

Examples of advanced telephone facilities

Abbreviated dialling
Automatic advise duration and charge
Automatic alarm calls
Automatic booked calls
Automatic transfer charge
Call waiting indication
Called line free indication
Centralized answering and recording service
Conference calls
Follow-me transfer
Personal code calling
Simple computing service
Transfer of call meeting busy
Transfer of call on ring tone—no reply

telephone services, there is scope for both expansion and improvement. Almost half the homes in the UK are still without telephone (residential penetration at March 1976 being 52%) and despite the present economic recession, telex and data continue to grow steadily. There is further scope for the use of micro-electronics and digital technology in the telecommunications systems, and for computers in design, manufacture, procurement, and operation.⁹ For existing services, this will decrease real costs, increase reliability, and create the opportunity to offer advanced telephone facilities of the kind listed in Table 1.

Both the established groups of wired services will extend to serve groups of up to around 100 participants. In the case of alpha-numeric and graphic communication this will be through the increased use of word processing and computer conferencing;¹⁰ in the case of audio communication through the increased use of multi-party telephone calls and studio-based audio teleconference systems. In the case of audio-visual communication, there are embryonic services for both one-to-one (videotelephone) and group purposes (e.g. Confravision).¹¹ There is an important role here too for micro-electronics in reducing the high transmission costs of switched visual services through bandwidth compression.¹²

The provision of the same message to many hundreds or thousands of different people is normally thought of as the province of wireless broadcasting rather than wired telecommunications. However, wired telecommunications are moving into this area with services providing all three forms of message.

The experimental Post Office Viewdata service provides such communication in alpha-numeric and graphic form.¹³ Using the customer's telephone line as the transmission path, with a decoder built into or associated with the customer's

	NUMBER OF RECIPIENTS							
	1	1-10	10-10 ²	10 ² -10 ³	10 ³ -10 ⁴	10 ⁴ -10 ⁵	10 ⁵ -10 ⁶	10 ⁶ -10 ⁷
ALPHA-NUMERIC & GRAPHIC	TELEX DATA FAX	WORD PROCESSING + COMPUTER CONFERENCING		VIEWDATA				
AUDIO	TELEPHONE	AUDIO CONFERENCE	AUDIO INFORMATION SERVICES					
AUDIO-VISUAL	VIEWPHONE	VIDEO CONFERENCE	CABLE TELEVISION					

Fig. 5. Existing and new wired services.

television set and a keypad to select pages, Viewdata provides a display similar to that of Ceefax or Oracle. The disadvantage of Viewdata compared to Ceefax and Oracle is that the customer incurs a call charge. The advantages of Viewdata are that there is no technical limit on the number of pages offered, there is (for up to say 100,000 pages) virtually no waiting time to retrieve a page, and appropriate prices may be set for viewing individual pages. This latter point is important, since it allows a full range of bargains to be struck between information provider and information user. In the case of advertising or public service information, the information provider may bear the whole cost including the call charge; in other cases (such as expensive financial information) the information provider may charge the user a price for the information in addition to the call charge. Viewdata is currently undergoing pilot trial within the premises of the Post Office and potential information providers, with a view to a wider market trial if there is evidence of viable demand.

The telephone information services (time, weather, dial-a-disc etc.) are an analogy to Viewdata in the audio medium, and have similar potential. A list of such services already provided by the Post Office is shown in Table 2(a). Between them, these receive over 500 million calls per annum. Yet the scope for further expansion is considerable. Without any technical improvement it is possible to conceive a much wider range of services; some additional services offered by overseas administrations are listed in Table 2(b). Technical improvements are certainly possible. These include the offering of loudspeaking wall-mounted receivers for telephone information services, with pushbutton and repertory dialling (giving Viewdata-like convenience of access), and equipment at the exchange in which the interest profiles of individual customers are stored for the automatic assembly of personalized information and entertainment bulletins.

Both Viewdata and the telephone information services are like broadcasting in that they make information widely available; they are like telecommunications in their use of switching technology to allow the user to select from many information sources. The ultimate service of this kind is advanced 'remote channel selection' cable television providing

instant access and response to dozens or hundreds of television programmes.¹⁴ Although the high hardware costs of such a system, coupled with the high costs of producing the necessary quantity of television programmes, rule it out as a practical economic possibility for the foreseeable future, there is a real prospect of expansion in more modest cable television systems providing access to 12 or 24 channels.

There are three good reasons why wired telecommunications should be moving into broadcast services in this way. Firstly, there is the argument of choice. Wireless broadcasting tends to be a mass medium, offering much less variety of material than is available through books, magazines, and newspapers. And the viewer or listener is heavily constrained in choosing from even this limited variety by the dictates of time. This can be overcome to some extent by the use of audio or video recording equipment (triggered either by time switch or by index codes embedded in the wireless broadcast signals). But there is no doubt that if the customer is to be offered an electronic variety of choice equivalent to that of the printed word, wired telecommunications are necessary.^{15, 16}

The second argument concerns the creation of an efficient market in information. The technology of wireless broadcasting imposes a very peculiar market mechanism.¹⁷ Because money cannot be collected from individuals for individual programmes, payment is either *en bloc* by compulsory licence fee, or indirectly through a willingness to tolerate intermittent advertising. This form of market is open to strong criticism on economic and social grounds. In economic terms, it allows the individual to control neither the amount of money he spends on programmes, nor the mix of programmes which he buys. In social terms, it imposes a central control on the provision of programmes which is at odds with the hard-won principles of a free press.¹⁸ This may not be a serious matter while there is a press to provide pluralism and diversity; but as electronic technology invades the world of paper the risks of central control loom larger. By contrast services with unlimited capacity and individual charging arrangements, such as Viewdata, could bring together information providers and information users in a free and open market.

The third argument for using wired distribution as a means of expanding broadcast programming is that this would conserve scarce radio frequency spectrum which is badly needed for other purposes such as mobile services.¹⁹

In summary, the future prospect for telecommunications services is one of improvement, of growth, of convergence between telecommunications and broadcasting, and of impact, through economic change, on the traditional media of print and face-to-face contact. Each of these trends calls for new approaches and new insights.

There is a danger that the improvement of existing services will encounter diminishing returns. To avoid this, it will be necessary to develop accurate and comprehensive measures of quality, and to assess the elasticity of demand with respect to these measures. Given information on the cost of quality improvements, this would make it possible to adopt a relatively rational traffic-maximizing approach to the setting of service quality. It would, however, leave the thorny problem of setting acceptable levels for the extremes, for due to differences in calling patterns individual customers will meet widely differing levels of service quality.²⁰

For a telecommunications administration, growth presents the great problem of expanding a massive and live system with high inertia;⁹ there is the prospect of an equally serious problem (particularly for manufacturers) of adapting to a reduced rate of growth when the telephone service begins to saturate the market.

The convergence between telecommunications and broadcasting will give rise to technological, legal, and institutional

Table 2
Recorded information services

(a) <i>Principal services provided by the UK Post Office</i>
Speaking clock
Dial-a-disc
Cricket scores
Weather forecasts
Financial news
Recipe
Tourist information
Motoring information
Bedtime story
Gardening
Record release
Food price news
Children's London
Wimbledon results
(b) <i>Other services offered by overseas administrations</i>
Horoscope
Dial-a-joke
News
Election results
Dial-a-prayer
Lottery results
Television/Film/Theatre programmes
Racing results

questions.^{21,22} Just as wireless broadcasting, libraries, and the press have established their own principles of access and control over content (varying from the scrupulous neutrality of the library to the paternalism of the wireless broadcasting organizations) so the new electronic information services must develop principles for access and control over content. I would advocate that, in the interests of democracy and freedom of speech, these principles should be closer to those of the library than those of the broadcasting monopoly.

The impact of telecommunications on the traditional media of print and face-to-face contact need not cause alarm. If the traditional media are ready to adapt and specialize, there is every reason why they should find ways of living alongside, and indeed drawing sustenance from, the electronic media. But this kind of adaptation will require systematic and timely analysis of the interactions between the traditional and the electronic media.²⁴ Much work has already been done on the potential interaction between teleconference systems and business travel.^{25,26} There is room for similar work on the interaction between electronic message systems and the traditional paper-based media; an interaction in which computer technology plays a central role.

References

- 1 Martin, J., 'Future Developments in Telecommunications' (Prentice Hall, Englewood Cliffs, N.J., 1971).
- 2 'Telecom 2000: an exploration of the long term development of telecommunications in Australia'. Australian Telecommunications Commission, 199 William Street, Melbourne 3000, December 1975.
- 3 Short, J., 'Residential telecommunications applications: a general review'. UK Post Office, Telecommunications Headquarters, Long Range Studies Division, Intelligence Bulletin No. 6, October 1975.
- 4 Communications Studies Group, 'The effectiveness of person-to-person telecommunications systems'. Communications Studies Group, University College, London, May 1975.
- 5 Short, J., 'Residential telecommunications applications: the growth of cable TV'. UK Post Office, Telecommunications Headquarters, Long Range Studies Division, Intelligence Bulletin No. 4, September 1974.
- 6 Seyler, A. J., 'Communicating by recorded information: a need for changing technology', *IEEE Trans. on Communications*, COM-23, No. 10, pp. 1065-70, October 1975.
- 7 Redmond, J., 'The future of broadcasting', *BBC Engineering*, No. 103, pp. 3-10, June 1976.
- 8 'Citizens band: a new way of life', *The Economist*, p. 65, 22nd November 1975.
- 9 Merriman, J. H. H., 'Engineering innovation in a service industry: Post Office Telecommunications', *Electronics and Power*, 20, No. 18, pp. 809-13, October 1974.
- 10 Vallee, J., Lipinski, H. M. and Miller, R. H., 'Group Communication through Computers: Vol. 1—Design and Use of the FORUM System' (Institute for the Future, Menlo Park, California 94025, 1972).
- 11 Collins, H., 'Business telecommunications applications: field trials of television conference systems'. UK Post Office, Telecommunications Headquarters, Long Range Studies Division, Intelligence Bulletin No. 5, December 1974.
- 12 Haskell, B. G., Mounts, F. W. and Condy, J. C., 'Interframe coding of videotelephone pictures', *Proc. IEEE*, 60, No. 7, pp. 792-800, July 1972.
- 13 Fedida, S., 'Screened information at the touch of a button', *Post Office Telecommunications J.*, 27, No. 4, pp. 4-6, Winter 1975-6.
- 14 Baer, W. S., 'Interactive television: prospects for two-way services on cable, Rand Corporation, Santa Monica, California. Report No. R-888-MF, November 1971.
- 15 'On the cable—the TV of abundance', Sloan Commission on Cable Communications (McGraw-Hill, New York, 1971).
- 16 'Urban cable systems: summary', Mitre Corporation, McLean, Virginia 22101, May 1972.
- 17 Jay, P., Evidence to Lord Annan's Committee on the Future of Broadcasting, Mimeographed, 1975.
- 18 Smith, A., 'The Shadow in the Cave' (George Allen & Unwin, London, 1973).
- 19 'Is mobile at a standstill?' *Telecommunications*, 9, No. 2, pp. 30-33, February 1975.
- 20 Short, J., 'Long range social forecasts: congestion and quality of service', UK Post Office, Telecommunications Headquarters. Long Range Studies Division, Intelligence Bulletin No. 10, 1975.
- 21 Oettinger, A. G., 'A prospective on information resources', Program on Information Technologies and Public Policy, Harvard University, Annual Report 1974-5, Vol. 1, 1975.
- 22 Reid, A. A. L., 'The grey area between telecommunications and broadcasting', *Television*, 16, No. 3, pp. 26-28, May/June 1976.
- 23 Short, J., Williams, E. and Christie, B., 'The Social Psychology of Telecommunications' (Wiley, New York, 1976).
- 24 Tyler, M., Cartwright, B. and Bush, G., 'Interaction between telecommunications and face-to-face contact: the energy factor', UK Post Office, Telecommunications Headquarters, Long Range Studies Division, Intelligence Bulletin No. 3, September 1974.
- 25 Nilles, J. M., Carlson, F. R. and Hanneman, G., 'The Telecommunications—Transportation Tradeoff' (Wiley, New York, 1976).

*Manuscript first received by the Institution on 10th December 1976 and in final form on 23rd March 1977.
(Paper No. 1772/Com. 151)*

© The Institution of Electronic and Radio Engineers, 1977

Letters

From: Franz Demmelmeier, Dipl.Ing.;
T. Jayashri, B.E. and D. Basu, M.Tech.

J. D. Ralphs, B.Sc., C.Eng., M.I.E.R.E.;
R. Benjamin, D.Sc., Ph.D., C.Eng., F.I.E.E., F.I.E.R.E.

Professor D. A. Bell, M.A., Ph.D., C.Eng., F.I.E.E.,
F.I.E.R.E.

Cellular Arrays for Multiplication of Signed Binary Numbers

In their paper,† T. Jayashri and D. Basu make a very interesting proposal for the multiplication of two signed binary numbers with an array of $n \times n$ cells. The array makes use of Robertson's first algorithm.‡

Based on this idea we have realized a multiplier which is the heart of a very fast arithmetic unit for a microprogrammable signal processor being developed. Detailed investigation of the multiplier demonstrated that the array according to Jayashri and Basu does not work correctly upon Robertson's first algorithm in that it can produce a wrong result for some combinations of the input operands (multiplier, multiplicand). This can be derived as follows.

A formulation of Robertson's first algorithm is given by:

$$\begin{aligned} P_i &= \frac{1}{2}(P_{i-1} + Xy_{n-i}) \\ P_0 &= 0, \quad i = 1 \dots n-1 \\ P_n &= Z = XY = P_{n-1} - Xy_0 \end{aligned} \quad (1)$$

where $P_i = i$ th partial product

$Z = XY$ product

$$X = x_0 + \sum_{i=1}^{n-1} x_i 2^{-i} \quad \text{multiplicand} \quad (2)$$

$$Y = -y_0 + \sum_{i=1}^{n-1} y_i 2^{-i} \quad \text{multiplier} \quad (3)$$

It is evident that the partial products are '0' as long as no multiplier bit, starting from the least significant bit, is '1'.

An example may illustrate this fact:

$$\begin{aligned} X &= 1.011 \quad (-\frac{5}{8}) & Y &= 0.1000 \quad (\frac{1}{2}) \\ P_1 &= \frac{1}{2}(0 + 1.011 \times 0) = 0 \\ P_2 &= \frac{1}{2}(0 + 1.011 \times 0) = 0 \\ P_3 &= \frac{1}{2}(0 + 1.011 \times 1) = 1.1011 \\ P_4 &= Z = 1.1011 - 1.011 \times 0 = 1.1011 \quad (-\frac{5}{16}) \end{aligned}$$

This numeric example yields to an incorrect result $Z' = 1.011100 (= -\frac{9}{16})$ (Fig. A when using the array of Jayashri and Basu, due to the sign extension of the partial products. For the correct implementation of Robertson's first algorithm we suggest the following solution.

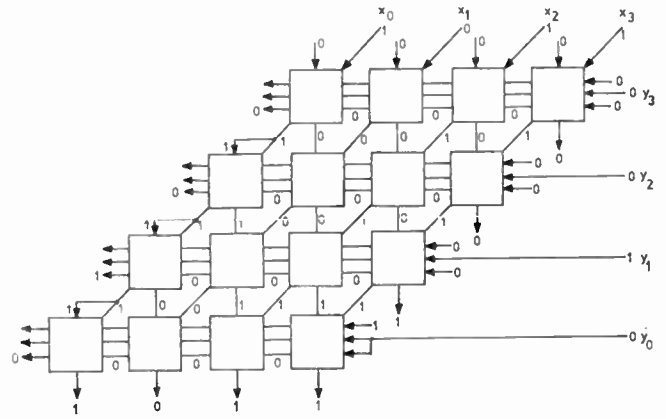


Fig. A. Array for multiplication based upon Jayashri and Basu† giving incorrect results.

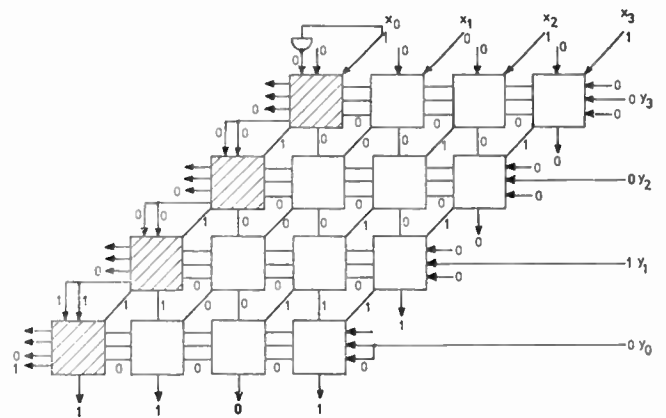
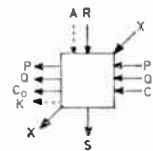


Fig. B. Proposed array for multiplication with sign correction of the partial products (shaded cells are sign cells).



$$\begin{aligned} S &= R \oplus C_1 \oplus Q(P \oplus R) \\ C_0 &= RC_1 + Q(P \oplus X)(R + C_1) \\ K &= (Q + A)X \end{aligned}$$

Fig. C. The cell (dashed input and output for sign cells only).

The sign extension must be stopped until a multiplier bit is one, starting from the least significant bit. The sign generation for the partial products must not be derived directly from the sign of the multiplicand but has to be computed according to the iteration

$$K_i = (K_{i-1} + Y_{n-i})X_0 \quad (4)$$

$$K_0 = x_0, \quad i = 1 \dots n-1$$

Figure B shows an array producing correct partial products and consequently a correct product for all combinations of input operands.

FRANZ DEMMELMEIER

Institut für Informationstechnik,
Lehrstuhl für Prozessrechner,
Franz Joseph Strasse 38,
8 München 40,
West Germany.

18th November 1976

† Jayashri, T. and Basu, D., *The Radio and Electronic Engineer*, 44, No. 1, pp. 18-20, January 1974.

‡ Chu, Y., 'Digital Computer Design Fundamentals', pp. 29-30 (McGraw-Hill, New York, 1962).

The comment made by Franz Demmelmeier is fully appreciated and we thank him for his interest in our paper. The array proposed in Fig. 2 of our paper implements Robertson's first algorithm correctly except for the case where the sign of the multiplicand is negative and the multiplier starting from the least significant bit has one or more zeroes. The problem is due to the improper sign extension of the partial products when the initial multiplication cycles are non-addition cycles. This can be corrected quite easily and Dr. Demmelmeier has suggested one such correction. In Fig. A, he has used some special cells with six inputs and six outputs for generating the sign bit of the partial products correctly. On close scrutiny, it can be observed that the additional input A is not needed for the sign cell. The output K is then given by:

$$K = (Q + R)X.$$

The R input to the sign cell in the first row is made equal to zero and the R inputs to the sign cells of successive rows can be derived directly from the K outputs of the sign cells of the previous rows. Thus, the inverter shown in Fig. B can also be eliminated.

It is to be further noted that the scheme given in Fig. 3 of our paper not only results in faster operation but also generates the correct result for all combinations of operands, without the necessity of having to use special sign cells.

T. JAYASHRI
D. BASU

Digital Processors Section,
Vikram Sarabhar Space Centre,
Trivandrum 695022,
India.

7th January 1977

The Physical Basis of Signal Theory

I would like to make two minor comments on the very interesting paper by Dr. R. Benjamin.[†] The first is on the difficulty mentioned in Section 2 of the paper, of the ambiguity in the use of the term 'bit'. Ask any practising communication engineer the number of bits in an ASCII character and he will almost certainly answer either 10 or 11. The minority (probably suspecting a trap) may say 8 but very few indeed would give the correct answer of 7 (there being 128 combinations in the ASCII code). In the days when the designer of a communication modem, the designer of the information source and code, and the communication theorist, were three separate men working under three separate disciplines, with remarkably little contact between them, the practical consequences of this ambiguity were acceptable. To the first, the bit was a single binary element arriving on an input wire and to be delivered to an output wire. The code and source engineer usually dealt within the relatively rigid framework of a specific application, while the theorist operated largely with generalized mathematical concepts which hardly left room for practical ambiguities of this nature.

With the ever-increasing direct application of theoretical concepts to practical equipments, and the increasing (and very laudable) practice of attempting to integrate a complete system design, the opportunities for confusion are increasing.

Although it may be considered too late to amend the education of a couple of generations of engineers and designers,

[†] Benjamin, R., 'The physical basis of signal theory', *The Radio and Electronic Engineer*, 47, No. 1/2, pp. 49-58, January/February 1977.

should not careful thought be given to the advisability of inventing another term which distinguishes between the two cases? May I (rather tongue in cheek) suggest that the word 'bit' be reserved for the binary unit of information, as defined by Dr. Benjamin, and the term 'bat' applied to the 'apparent bit' or signal element. The information contained in a signal stream of n 'bats' would then be identical to n 'bits' if, and only if, each separate element were independent and uncorrelated.

The second point, arising from Section 2.4 of the paper, is the influence of system bandwidth on communication efficiency. If one takes the Shannon Law in its more general form:

$$C = B \log_2 \left(1 + \frac{S}{N} \right)$$

where

N = information rate in bits per second

S = mean signal power

N = mean noise power in the channel

B = bandwidth of the channel (which is assumed to be the same as the bandwidth of the signalling system).

This equation can be recast in terms of two fundamental normalized parameters:

B_0 = normalized bandwidth

$$= B/C$$

R = normalized signal-to-noise power ratio

$$= \frac{\text{signal energy per bit}}{\text{noise power per unit bandwidth}}$$

$$= \frac{S}{C} \cdot \frac{B}{N} = \frac{S}{N} \cdot B_0$$

so that

$$R = B_0(2^{1/B_0} - 1).$$

The minimum value of the normalized bandwidth B_0 is 0.5 (the Nyquist limit) for which $R = 1.5$, while for an infinite bandwidth $R = 0.69$. Thus for the very common practical case of fixed data rate and signal power, the choice of system bandwidth cannot affect the limiting performance of the system by more than 3½ dB.

It follows that although bandwidth is obviously an extremely important practical parameter, there would seem to be no justification for attempting to argue the case for wide bandwidth versus narrow bandwidth systems on the basis of the theory of linear channels with additive noise. Provided that the basic principles derived in the paper of spreading the information evenly over the frequency band and the time available are adhered to, the detailed structure of the signal (such as element duration and frequency shift) can be chosen relatively freely from considerations of the time-varying and frequency-varying characteristics of the communication channel itself or practical limitations on equipment.

J. D. RALPHS

Development Section,
Communications Engineering Department,
Foreign and Commonwealth Office,
Hanslope Park, Milton Keynes MK19 7BH.

8th March 1977

I am grateful to Mr. Ralphs for his thoughtful comments on my paper. I do of course agree on the importance of distinguishing the 'bit' as a unit of information from the binary digit as an element of message structure, and I agree that a distinct name for the latter could help. The 'baud', which I believe is strictly speaking the rate of one (structural) bit *per second*, seems to be increasingly used, colloquially, to denote a binary digit of message structure. Mr. Ralph's proposal of the new name of 'bat' would be a cleaner solution. (However, I dread the misprint 'the bat bit the bawd'.)

Regarding Mr. Ralph's second point, I fully agree on the significance and utility of the two normalized parameters:

(i) the bandwidth *per unit bit rate*, i.e. the spreading factor: time-bandwidth product per bit of information;

(ii) the signal energy *per bit*/noise power *per unit bandwidth* (assuming this noise-power density is indeed uniform).

I also agree with his conclusion that, for a given normalized signal/noise ratio, the spreading factor and detailed modulation structure have little effect on the actual signal/noise ratio, provided the noise is statistically well behaved and the signal and noise are spread appropriately over the frequency-time domain. Indeed a given signal sample over multiple frequency-time cells can be converted coherently into any other optimal spreading over one or more cells, without changing the resultant signal/noise ratio.

However, an enlarged frequency-time coverage does increase the amount of information which can be conveyed by a given energy. (Spreading each given information digit and its energy more widely in frequency-time is not the same thing.) At the other extreme, a given amount of information can be crowded into an indefinitely narrow time—bandwidth domain, provided the signal/noise ratio is high enough for the amplitude of each signal sample to convey sufficient independent 'bits' of information. This subject, and its relation to the other design criteria touched upon by Mr. Ralphs, have been discussed more fully elsewhere.†

R. BENJAMIN

Chief Scientist,
Government Communications Headquarters,
Oakley,
Cheltenham, Glos. GL52 5AJ.

21st March 1977

The Statistical Nature of Communication Theory

Although one of the major books in the field is entitled 'Introduction to Statistical Communication Theory'¹ one usually finds Shannon's channel-capacity theorem stated as a deterministic equation.

$$C \leq W \log(1 + P/N) \quad (1)$$

where C = communication rate (in bits per second if the logarithm is to base 2), W = bandwidth in hertz of a uniform channel, and P, N , are the signal and noise powers. Probing will usually elicit the further information that powers are to be measured in terms of mean-square amplitudes and the noise is assumed to have a Gaussian distribution. (But the distribution of signal amplitudes is less usually specified.)

† Benjamin, R., 'The scope for signal processing in systems design', introductory paper in International Specialist Seminar on The Impact of New Technologies in Signal Processing', IEE Conference Publication 144, September 1976.

Let us accept these limitations in order to get a tractable result, though the limitation of signal by mean power may not apply to practical situations where one is limited in peak power. But reference to Shannon's derivation of the channel-capacity theorem by the use of a model in multi-dimensional geometry² shows that the basic theorem is as follows: Given the use of a channel of bandwidth W for time T with signal-to-noise mean-square amplitude ratio P/N , the ensemble of possible messages which can be mutually distinguished with probability of error ϵ is of number M given by

$$M \leq (1 + P/N)^{TW} + \epsilon \quad (2)$$

Discrete information is defined as the logarithm of the number of distinct messages in the ensemble so that,

$$I = \log M \leq TW \log(1 + P/N) + \log \epsilon \quad (3)$$

The step from equation (3) to equation (1) is performed by dividing by T and assuming T large so that $(\log \epsilon)/T \rightarrow 0$. It is not always realized however, that TW is the time-bandwidth domain of the signals, each signal being a recognizable entity.

Since $\epsilon \leq 1$, $\log \epsilon$ is negative, so that there is subtracted from the TW term a term which becomes numerically greater as the acceptable risk of error is reduced. To get some idea of the numbers involved, suppose we wish to communicate at the 0dB signal/noise level with an error rate of 10^{-3} . Then $\log_2(1 + P/N) = 1$ and $\log_2 \epsilon \approx -10$ and use of $TW = 100$ will allow us 90% of the rate indicated by the excessively idealized formula.¹ (It is also about the least value of TW which gives the derivation of formula (2) some approximate validity.³ The formula is not valid unless TW is large, but it may be used as an analogy for small TW . We must reduce the information rate of the channel in order to allow for the negative term $\log \epsilon$. The reduction in rate is effected by adding redundancy: in the language of coding theory one does this to increase the Hamming distance between signals, a form of presentation which relates back to the geometric symbolism.

With present-day signal processing it may be possible to identify individual signals with 100 parameters ($TW = 100$) but it is not practicable to provide means to identify separately each one of the M possible signals. In the absence of such means the relation of the channel-capacity theorem to a practical communication system resembles at best that of Carnot's theorem to a petrol engine. Telegraphic signals (e.g. binary) are usually detected digit-by-digit so that $TW = 1$ and the channel-capacity theorem is inapplicable. But some approach to increasing TW is obtained by the use of some form of soft-decision decoding which is therefore a most important line of development.

D. A. BELL

University of Hull,
Department of Electronic Engineering,
Hull HU6 7RX

24th May 1977

References

1. Middleton, D., 'Introduction to Statistical Communication Theory' (McGraw-Hill, New York, 1960).
2. Shannon, C. E., 'Communication in the presence of noise,' *Proc. I.R.E.*, 37, pp. 10-21, 1949.
3. Bell, D. A., 'Information Theory and its Engineering Applications,' 4th Edition (Pitman, London, 1968).

IERE News and Commentary

Silver Jubilee Exhibition in Hyde Park

British industry and commerce will demonstrate their achievements and potential at a Silver Jubilee Exhibition which is to be held in Hyde Park, London, from 1st July to 4th September 1977. The exhibition will be near Marble Arch and is the first exhibition in this Royal Park since the famous Exhibition of 1851. This time, instead of a 'Crystal Palace', the great tent which was specially designed and built for the Royal Inauguration in Aberdeen of the Forties Oil Field is to house the exhibition which, with its co-ordinated design and specially invited exhibits, may well have some of the atmosphere of the pavilions of the Festival of Britain in 1951.

The organizers anticipate large crowds because of the theme and convenient access to the exhibition and are arranging specific evenings set aside for different professions to view it under leisurely conditions. The evening reserved for the engineering profession is Tuesday 9th August, from 7 p.m. to 8.30 p.m. A limited, *pro rata*, allocation of tickets has been made to the IERE and members are invited to apply as soon as possible for these on the understanding that requests will be dealt with in order of their receipt. Please write to Mr. R. W. Stobbart, Personal Assistant of the Secretary, IERE, 9 Bedford Square, London WC1B 3RG. Applications received after all the available tickets have been allotted will *not* be acknowledged.

'Applications of Charge-Coupled Devices'

A 2-day residential Summer School is to be held at the University of Wales Institute of Science and Technology, Cardiff, on 15th to 16th September 1977. Its aim is to introduce applications engineers to charge-coupled devices and to the ways in which these devices may be used in a wide range of applications. For those not familiar with c.c.d.s. there will be an introductory session outlining the general operating principles which will lead to an appreciation of their capabilities, as well as their limitations. This will be followed by three applications sessions which will deal in turn with the three areas—imaging, analogue signal processing and digital applications—in which c.c.d.s. are expected to make a major impact in the next few years. Lectures will be given by staff from industrial, government and University establishments.

Further details may be obtained from Professor J. D. E. Beynon, Ph.D., C.Eng., F.I.E.R.E., Department of Physics, Electronics and Electrical Engineering, UWIST, Cardiff CF1 3NU. (Telephone: 0222-42522).

The HND and its Future

At the time of the Colloquium on 'The Future of Higher Diplomas' reported in the May issue, the future of the HND in Electrical/Electronic Engineering was uncertain. It was known that there was strong feeling that the HND should not be axed merely in the interests of 'tidiness' or 'uniformity', but there was some doubt as to whether its special value was sufficiently appreciated in the 'corridors of power' for it to escape this fate. All those concerned about the future of the HND will therefore doubtless be pleased to learn that the Institution of Electrical Engineers, our senior partner in the Joint Committee concerned, has just announced that there are no plans to withdraw the HND, since the development of Technician Education Council courses is not yet sufficiently advanced for the Joint Committee to feel confident that they are an adequate alternative to the present system.

A summary history of the Joint Committee and its awards, which was not covered in the Colloquium report, may be of interest. National Certificate and National Diploma courses in Electrical Engineering were set up in 1922. The division of responsibility was clearly defined between the Institution of Electrical Engineers (IEE) and the Board of Education—now Department of Education and Science (DES), whereby IEE was responsible for the standards of coursework and examinations including the appointment of assessors to monitor those standards, while the Board of Education was responsible for advising on the staffing and equipment for colleges. At the same time a Joint Committee was formed consisting of representatives from both the IEE and the Board.

In 1958 the IEE and the Board of Education agreed to modify the procedure for the appointment of the Joint Committee. This modification was to provide for the representation of technical colleges on the nomination of the Association of Technical Institutions, the Association of Teachers in Technical Institutions and the Association of Principals of Technical Institutions. In 1961 it was decided that an additional member should be invited to serve, nominated by the British Institution of Radio Engineers (now the IERE), since when the IEE and the IERE have jointly shared the responsibility for the appointment of assessors for monitoring the standards of the Electrical and Electronic Engineering courses.

A final change in the Constitution of the Joint Committee followed the merger with the Northern Ireland Joint Committee and the advent of Technician Education Council (TEC) in 1973 to include nominees from both these bodies. The make-up of the Electrical and Electronic Engineering Joint Committee is now as follows:

Three representatives from the IEE

Three representatives from the IERE

Two representatives from the DES

plus

One representative from each of the following:

The Ministry of Education, Northern Ireland

The Association of Principals in Colleges

The Association of Colleges of Further and Higher Education

The Association of Teachers for Further and Higher Education

The Technician Education Council

There are twenty-five other Joint Committees concerned with the further education and development of technicians in various fields. Similar statements regarding the continuation of HND courses have been issued in relation to Mechanical, Production and Aeronautical Engineering.

Institution Premiums for 1976

The Council of the Institution announces that authors of the following papers are to receive Premiums for outstanding papers published in the Journal during 1976.

MAIN PREMIUMS

CLERK MAXWELL PREMIUM *Value £75*

'An automatic clarifier for s.s.b. speech communication'
Professor W. Gosling (University of Bath), Dr. S. R. Al-Araji (University of Iraq) and J. R. Olivera (University of Havana) (Published in the February 1976 issue of the Journal)

HEINRICH HERTZ PREMIUM *Value £50*

(Physical or mathematical aspects of electronics or radio)
'Performance of an M-ary p.s.k. system in the presence of additive noise, inter-symbol interference and fading'
Dr. J. Nowakowski and I. Protasewicz (Technical University of Gdansk) (February)

MARCONI PREMIUM *Value £50*

(Engineering)
'An h.f. channel simulator using a new Rayleigh fading method'
J. D. Ralphs (Communication Engineering Department, Foreign and Commonwealth Office) and F. M. E. Sladen (University of Southampton) (December)

SPECIALIZED TECHNICAL PREMIUMS

CHARLES BABBAGE PREMIUM *Value £25*

(Computers)
'Integrator design for a differential analyser'
W. Forsythe (University of Technology, Loughborough) and S. L. Houseman (IBM Laboratories, Hursley Park) (December)

LORD BRABAZON PREMIUM *Value £25*

(Aerospace, maritime and military systems)
'Identification of complex geometrical shapes by means of low-frequency radar returns'
Dr. Y.-T. Lin (Sandia Laboratories, Albuquerque, New Mexico) and Professor A. A. Ksienski (Ohio State University Electro Science Laboratory) (October)

REDIFFUSION TELEVISION PREMIUM *Value £50*

(Communications)
'Surface acoustic wave matched filters for communications systems'
K. V. Lever, Dr. E. Patterson, P. C. Stevens and I. M. Wilson (GEC Hirst Research Centre) (May)

A. F. BULGIN PREMIUM *Value £25*

(Components and circuits)
'Signal processing applications of charge-coupled devices'
Dr. J. Mavor (University of Edinburgh) (August/September)

SIR CHARLES WHEATSTONE PREMIUM *Value £25*

(Instrumentation and measurement)
'The use of a high sensitivity ultrasonic current meter in an oceanographic data acquisition system'
T. Gytre (Institution of Marine Research, Bergen) (December)

GENERAL PREMIUMS

ARTHUR GAY PREMIUM *Value £25*

(Production techniques)
'The development of an automatic wiring analyser system for testing telephone switching rack backplanes'
B. Coolbear (Standard Telephones & Cables Ltd.) and G. Lovitt (Standard Telecommunication Laboratories Ltd.) (July)

LESLIE MCMICHAEL PREMIUM *Value £25*

(Management techniques)
'The comparative cost of associative memory'
R. M. Lea (Brunel University) (October)

SIR J. C. BOSE PREMIUM *Value £25*

(Outstanding paper by an Indian scientist or engineer)
'The synthesis of ladder networks with resistances at both ends by the recurrent-continuant method'
Dr. D. K. Jha and Dr. S. C. Prasad (University of Bihar, India) (August/September)

LOCAL SECTIONS PREMIUM *Value £25*

(Outstanding paper first read before any of the Local Sections in Great Britain outside London)
'Submillimetre waves—a survey of the 'state of the art' and some recent developments in research'
Dr. R. J. Batt (Portsmouth Polytechnic) and Professor D. J. Harris (UWIST) (August/September)

Papers of sufficiently high standard were not published within the terms of the following Premiums and they are withheld:

J. Langham Thompson Premium (Control engineering)
Dr. V. K. Zworykin Premium (Medical and biological electronics)

P. Perring Thoms Premium (Radio or television reception)
Dr. Norman Partridge Premium (Audio frequency engineering)

Lord Rutherford Premium (Atomic or nuclear physics)
Eric Zepler Premium (Education of electronic and radio engineers)

Admiral Sir Henry Jackson Premium (History of radio or electronics)

Hugh Brennan Premium (North Eastern Section paper)

The presentation of Premiums will be made by the President at the Institution's Annual General Meeting in London on Thursday, 13th October, 1977.

The Electronics Industry in Review

In his speech at the Annual Dinner of the Electronic Engineering Association in London on May 11th, the President, Mr P. E. G. Bates (Plessey) admitted to being at first sight depressed to see that things had not apparently progressed very much since his predecessor's speech in 1976. Inflation continued to rage at an intolerable rate, taxation continued at a level discouraging investment, the erosion of differentials and the inability to provide worthwhile incentives continued to put a straitjacket on the skills and initiative of those engaged in British industry. Although the Government had acknowledged that the creation of wealth was basic to economic wellbeing and that the private sector had a vital role to play, it seemed to be relying on a limited armoury of weapons: the so-called industrial strategy and phase 3 of the prices and incomes policy floating on an ocean of North Sea oil may not be enough. Mr Bates continued:

'But I would instead like to be positive—because I think that progress is being made in a number of ways—and the EEA is making a valuable contribution, as befits the trade association of such an important industry as ours.

'In 1976 the Industry's total output was estimated at £1400 M, an increase of 28% over 1975 and of this total some 42% was directly exported; the radio, radar and other electronics capital goods sector itself had a positive balance of trade of £206 M, an improvement of nearly 40% over 1975. These are figures we can be proud of. This is a high technology industry which is well suited to the inventiveness, brains and talents of our countrymen.

'The Industry has a fine record of achievement and a tremendous potential for the future. We naturally expect it to be regarded as one of the high flyers which should have every assistance from Government. Other nations, who are our competitors, have appreciated the importance of electronics and continue to channel investment and other help into their electronics industries. If we wish to compete on equal terms then we as a country must do the same. The Government has a major influence on us—as a customer—and on the national environment within which we have to operate. We trade in the public sector, and we are much affected by public policy in our efforts at home and overseas. If public policy is right, we and the nation will thrive.

'We have therefore as an Association continued to support the development of policy through the evolution of the industrial strategy—since this seems to us to offer an opportunity for a continuing constructive dialogue with the Government on the issues which we believe to be important. We have supplied representatives to the Sector Working Parties, and we have made carefully-considered submissions and set in hand much study work to clarify particular issues.

'Many of the issues we regard as important affect the electronics industry as a whole—I have tried to revive the role of the Conference of the Electronics Industry in co-ordinating the efforts of the EEA with our sister associations—but we need to promote this far more positively still. It is important to avoid fragmentation—combined strength is much more likely to produce results and I was delighted to see this approach leading to the consolidation of the interests of our components colleagues during the year into the ECIF.

'Possibly the most important problem facing the entire electronics industry in this country today arises from the far-reaching changes in technology which are in the process

of being developed. The future development of integrated circuits makes it increasingly difficult to draw the line between components and sub-systems and while the equipment industry will adapt to this situation, it will involve radical and expensive changes both to our laboratories and factories and to the skills and training of our work force over the next five or ten years. EEA is joining with other authorities concerned in thinking out some of the problems and threats contained in these developments and the policies which need to be pursued both on a national and on a European basis.

'Overseas, our business prospers but at home the picture is darker. Once more our major customer, the Ministry of Defence, has been through yet a further defence review, with further cuts in expenditure. As a citizen of this country, one can only be appalled at the further decline in our ability to defend ourselves—as a member of this industry, one can only grit one's teeth and at the same time marvel at the ingenuity of the MoD in their efforts to minimize the effect on the efficiency of the Armed Services.

'EEA is a Trade Association, not an employers' organization—and we do not normally get directly involved in pay negotiations—or indeed in the wider field of industrial relations. There are however certain current issues on which we cannot and must not remain silent.

'In common with the rest of industry, our members have been greatly exercised by the majority report of the Bullock Committee on Industrial Democracy. Starting from a diverse base in this country, the majority report proposals have taken an enormous leap to an extreme solution which could strike at the heart of managerial and industrial efficiency. EEA has expressed its full support for the CBI approach in this matter, and believes that we should strive to develop Industrial Democracy in companies from a practical base, rather than attempt a political solution which is likely to over-complicate and weaken the decision-making processes in the Boards of our major companies'.

Mr Bates concluded by introducing the principal guest, Sir Derek Ezra, Chairman of the National Coal Board. He believed that Sir Derek was in a unique position to take a broad view of the country's industrial scene and its future prospects. EEA was convinced that electronics has a major part to play in the modernization of nationalized industries and would welcome Sir Derek's thoughts on the way in which it may be able to work more effectively with the nationalized industries in the export field in the future.

In reply Sir Derek Ezra said that the long-term plans of the coal industry depended on new technological developments and the electronics industry had a vital role to play in these.

In looking to the future, the Coal Board was seeking to minimize the impact of short-term variations. 'We have to move our thinking away from short-term policy horizons, with the accompanying risk of sudden changes of direction, towards evolving longer-term planning with consistency as the key. Both our industries demonstrate this need—electronics, where the development of new technology is a long-term matter, and coal, where mine construction times are also long.'

Sir Derek said there were 'exciting possibilities' in the developing relationship of the electronics and nationalized industries. The state industries were large, with resources

(cont. on facing page)

Nominations for Election to the 1977-78 Council of the Institution

In accordance with Bye-law 49, the Council has nominated the following members for election at the Annual General Meeting to be held in London on Thursday, 13th October 1977.

The President

For Election: Professor W. A. Gambling, PH.D., D.SC.

The Vice-Presidents

Under Bye-law 46, all Vice-Presidents retire each year but may be re-elected provided they do not thereby serve for more than three years in succession.

For Re-election:

Professor D. E. N. Davies, PH.D., D.SC.; D. W. Heightman; Professor J. R. James, PH.D.; R. C. Hills, B.SC.;
Professor W. Gosling, A.R.C.S., B.SC.; J. Powell, B.SC., M.SC.

Honorary Treasurer

For Re-election: S. R. Wilkins

Ordinary Members of Council

Under Bye-law 48, Ordinary Members of Council are elected for three years and may not hold that office for more than three years in succession.

MEMBERS

The following must retire: M. S. Birkin

For Election: A. F. Dyson, DIP.EL.

The remaining members of Council will continue to serve in accordance with the period of office laid down in Bye-law 48. Bye-law 50 provides that:

Within twenty-eight days after the publication of the names of the persons nominated by the Council for the vacancies about to occur any ten or more Corporate Members may nominate any one other duly qualified person to fill any of these vacancies by causing to be delivered to the Secretary a nomination in writing signed by them together with the written consent of the person nominated undertaking to accept office if elected, but each nominator shall be debarred from nominating any other person for the same vacancy.

By Order of the Council

S. M. DAVIDSON

9th June 1976

Secretary

The Electronics Industry in Review

(cont. from facing page)

of finance, techniques and personnel to match; they occupied a fundamental position in the country's economy; they were seeking rapid technological development and they were responsible for a large amount of electronics business.

'In the coal industry our own need is to make another leap forward in our productivity performance. Having reached total mechanization we are now working hard to extend remote control and automatic operations. Even small advances, however, can be difficult. It took a considerable effort to establish effective radio communication underground—a basic, essential step to free men for a greater range of duties while yet retaining effective control of the mining operation.

'The next steps in mining processes promise almost limitless opportunities for technology of many kinds. Among them electronics must be foremost. We have now a long-range research group whose job is to consider new mining methods for the period around the end of the century. They have

visited the leading research establishments of every kind in the country in their quest for new, useful technology. They have been given every co-operation and it requires, I am sure, only the right outlook for the chances of co-operation in any projects that these researches call for, to be continuing and effective, and to contribute substantially to the development of technological expertise in Britain. The spin-off in overseas business would be substantial,' he concluded.

At the EEA Annual General Meeting earlier in the day, Mr R. H. Newham was elected President of the Association and Chairman of Council for the year 1977/78, in succession to Mr Bates. Mr Newham who is a Director of EMI Electronics Limited, joined the Gramophone Company (now EMI Records Limited) in 1937 and during the past 20 years with EMI he has held a number of senior positions, principally concerned with research and development. He is currently the Director of EMI Electronics Limited responsible for engineering and marketing. Since 1974 he has been the leader of the United Kingdom delegation in the NATO Industrial Advisory Group.

Synthesizer Circuits for H.F. and V.H.F. Radios

A range of purpose-designed synthesizer circuit kits primarily intended for citizens' band radio applications has been developed by Plessey Semiconductors of Swindon, Wiltshire. These circuits form two-chip solutions to the requirements of synthesized local oscillators for 40-channel citizens' band transceivers, needing only a minimum number of external components. ('Citizens' band' operation is not permissible in the United Kingdom.)

The circuits are designed so that the programmable frequency division is carried out by one chip, and the clock, phase detection and recognition on the other. The two-device set, comprising the SP8921 and SP8922, forms a 40-channel binary programmed synthesizer, while the two-device set, comprising the SP8921 and SP8923, forms a 40-channel b.c.d. (binary coded decimal) programmed synthesizer.

In the b.c.d. version, the channel to frequency translation is carried out on the chip by means of a read-only memory. The set contains a preamplifier followed by a fixed divide-by-four prescaler. The amplifier input will accept a signal at a frequency up to 30MHz. The prescaler is followed by seven bits of programmable division, which are programmed directly by seven inputs on the SP8922. On the SP8923 the inputs are programmed via a decoding matrix which converts the 40 input codes to the appropriate Citizens Band frequencies. A receive/transmit input is provided to give an offset of minus 91 counts (corresponding to -455kHz) when in the receive mode. An emergency channel select output is provided (EC O/P) on the SP8923 giving a low output when a non-permitted channel or channel 9 is selected. When this occurs the synthesizer programs automatically to the emergency channel 9.

The SP8922/3 provide a clock and preset output to the SP8921 and receive a recognition signal.

The SP8921 contains the six most significant bits of the programmable counter, a 10.24MHz crystal oscillator maintaining circuit, a 2^{13} fixed divider, and a digital phase/frequency comparator. The six bits of the programmable counter have a fixed preset code which, combined with the variable code of the SP8922/3, give the total count required to select the 40 citizens' band frequencies.

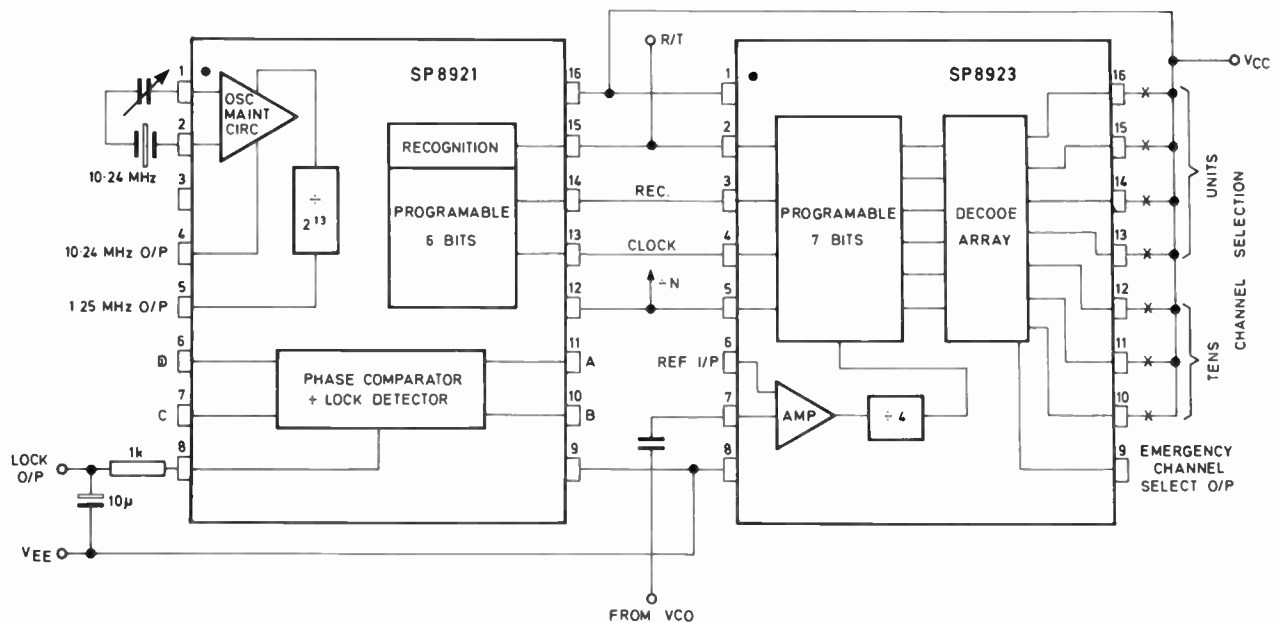
The receive/transmit input to the SP8921 gives an offset in the programmable count of -2048 (corresponding to -10.24MHz) when in the receive mode.

The crystal oscillator has a direct emitter follower output which may be used as an input to the second mixer in a double conversion transceiver. The oscillator is connected internally to the fixed divider which gives an output reference frequency of 1.25kHz with a 10.24 MHz crystal.

The phase/frequency comparator has two outputs which may be used to drive a variety of charge pump filter circuits. There is also an in-lock detector output.

All these circuits are fabricated by the Plessey bipolar Process III, which is a high-frequency low-power process, giving the facility of direct counting without the need for a mixer, an important advantage over competitive products made by the c.m.o.s. process. The high packing density of Process III makes it possible to produce the synthesizers in large quantities at a very competitive selling price.

The SP8922 and SP8921 may be used in other applications such as marine band radio. 128 channels with spacing can be selected and with mixing techniques the synthesizer frequency band can be selected at will. The input may be prescaled by 5 to give 25kHz spaced channels. The fixed program divider in the SP8921 may also be replaced by a c.m.o.s. presettable divider to produce a 5 MHz to 200MHz non-mixing synthesizer on 25kHz channels.



Block diagram of 40-channel b.c.d. programmed synthesizer for 'citizens' band' radio, using the Plessey SP8921 and SP8923 integrated circuits.

Signal Generators of the Future

Two important new synthesized sources

Few engineers would deny that one of the instruments without which development work would be impossible is a reliable signal generator and the electronic instrument industries of the world have laboured assiduously to provide this requirement. For many years refinements of the straightforward variable frequency oscillator reigned supreme until, during the sixties, developments of synthesized signal sources, based on quartz crystal oscillators, for communications led to synthesized signal generators for the measurement laboratory. But these had limitations in certain respects which two leading British companies have now claimed to remove, or at least minimize. Within a few weeks of each other Racal Instruments of Windsor and Marconi Instruments of St. Albans announced new synthesized signal generators with impressive specifications. We believe that the following information which they have respectively issued will prove of interest to every electronic engineer, who has any involvement with the precision measurements that are called for today in production and servicing just as much as in R & D and calibration.

The Marconi Instruments TF 2020

This 50 kHz to 520 MHz high performance synthesized signal generator is claimed to combine the accuracy and stability of a synthesizer with the spectral purity associated with the very best signal generator. It has full a.m. and f.m. facilities, a stability better than 1 in 10^8 per day, very low noise, spurious signals better than 90 dB down, 20 Hz to 99 kHz internal modulation oscillator, frequency deviation from 0 to 300 kHz and amplitude modulation to 99%. All of these functions are controlled initially by m.o.s. logic manually operated by front panel switches or remotely through parallel b.c.d. control. These features, together with very low leakage

and a wide output range (from 0.2 μ V to 4 V e.m.f. from 50 Ω) make the instrument ideal for research and design work in the laboratory as well as for quality control and testing in production and service organizations.

In the production or servicing applications the synthesized signal generator can be controlled by a test jig—a single box of diodes and wires connected to a series of switches so that each switch will set up all of the functions of the generator for one particular measurement. Alternatively the instrument can be programmed by a desk top calculator or built into a fully automatic computer-controlled test system.

The carrier frequency can be set in 1 Hz increments over the 50 kHz to 520 MHz range with a rapid settling time after switching of less than 10 ms. A feature of the instrument is its very low sideband noise making it suitable for use as the off-channel signal generator in two-generator selectivity measurements. At 20 kHz offset frequency the sideband noise is at least 130 dB/Hz below the carrier level.

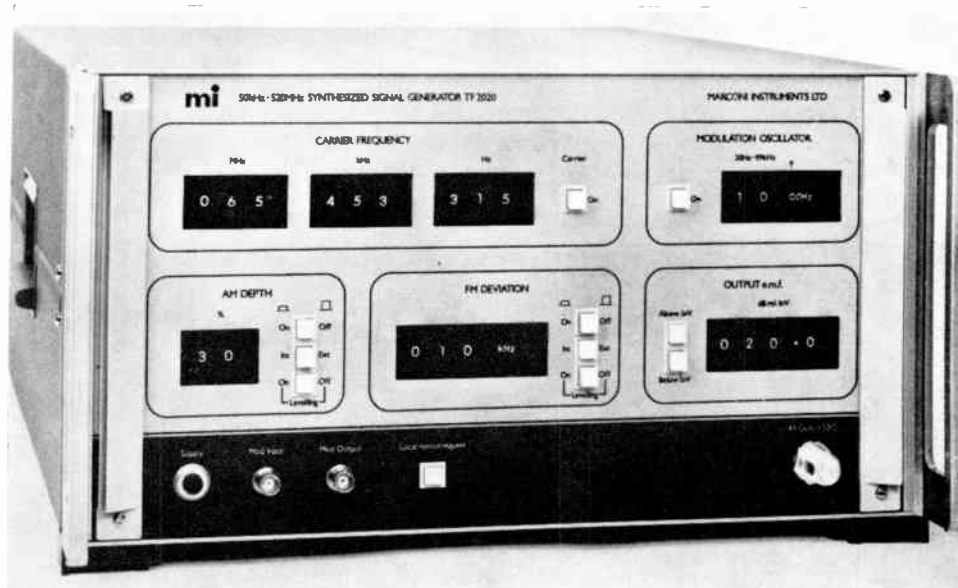
The output level is selected by an attenuator settable in 0.1 dB steps. Automatic level control is incorporated and the total level accuracy is ± 1 dB for outputs above 1 μ V. Despite the high output level of the instrument, leakage has been kept to a minimum, allowing the generator to be used at the minimum output of 0.2 μ V (-13.9 dB μ V e.m.f.) without ambiguity. Non-harmonically related signals are very low, at least 90 dB below the carrier level and this combined with the excellent sideband noise specification makes the TF 2020 an ideal signal source.

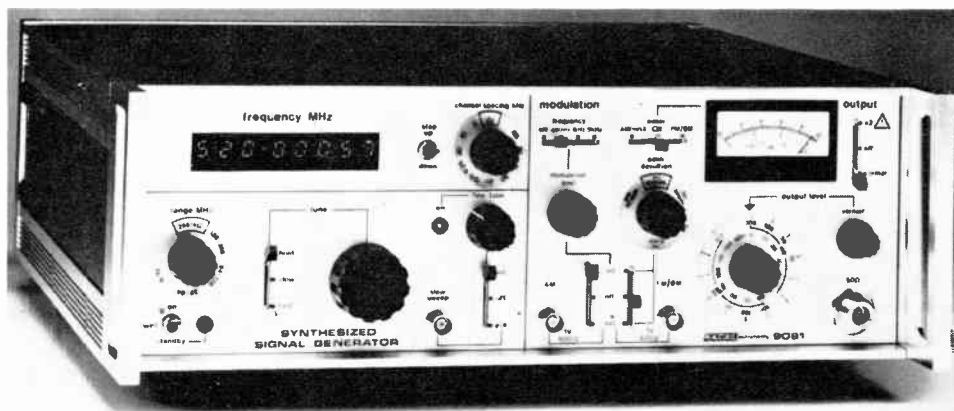
The TF 2020 conforms to IEC 348 and BS 4743 safety specifications. Its dimensions are: width 450 mm, height 234 mm, depth 610 mm; weight is 36 kg. The UK price is £9,500.

The Racal Instruments 9081

The 9081 provides a facility which is stated to be totally new in signal generators—synthesized performance with analogue tuning. Covering 5–520 MHz in 5 bands, tuning is achieved by a spin wheel which is electronically geared to provide fast or slow tuning rates. A hold facility electrically disconnects the spin wheel to prevent accidental frequency changes. The output changes in discrete synthesized steps and is locked automatically to a fast warm-up, ovened crystal frequency standard with an ageing rate better than ± 3

The Marconi Instruments TF 2020





The Racal Instruments 9081

parts in 10^9 per day (± 5 parts in 10^{10} optional). Interpolation between steps is achieved with a fine tune vernier which gives the 9081 a continuous tuning capability. When the vernier is in use the output is still locked to a crystal frequency standard and has a stability at least one order better than any analogue generator.

For channelized radio communications 9081 includes special facilities for the rapid and precise selection of radio channel frequencies. Operators can choose a step size to correspond to any international channel spacing from 5 to 60 kHz. As the spin wheel is rotated the output then alters in channel-related steps. Secondly, a step switch shifts the frequency up or down by one channel space. The operator can thus step quickly and accurately through as many channels as required. The output is monitored by an l.s.i. counter that continuously updates the 8-digit l.e.d. display. High accuracy and stability is derived from the frequency standard used for the synthesizer.

The 9081 has full a.m., f.m. and phase modulation facilities and the carrier can be modulated either from an internal oscillator or external sources. A front panel meter gives a clear indication of modulation amplitude, frequency deviation or phase deviation. Simultaneous modulation using a.m. combined with either f.m. or phase modulation is achieved using internal or external sources.

Automatic voltage levelling maintains the output typically within ± 0.5 dB over the entire frequency range. Output level is shown on the meter and is controlled by a switched attenuator and vernier control providing a range from -130 dBm to $+3$ dBm ($0.07 \mu\text{V}$ to 316 mV into 50Ω). Discrete spurious frequencies are at least -70 dB at 20 kHz or more off the carrier.

The signal generator is designed to meet European safety standards and is built to military environmental specifications. It measures $133 \times 408 \times 445$ mm and weighs approximately 13 kg. The current UK price is £2250.

Microwave Radio System for Bolivia

Within two months of successfully completing one of the world's highest microwave-radio telecommunications systems, in the Bolivian High Andes, GEC Telecommunications Limited of England has been awarded a £6.7 million turnkey contract to expand its capacity and provide an extension to the principal cities in the southern part of the country. The extension will also allow interconnexion between the Bolivian national telecommunications network and the telecommunications system in Argentina.

The contract, awarded by Empresa Nacional de Telecomunicaciones (ENTEL) of Bolivia, will more than double the capacity of the existing 2 GHz microwave-radio network and extend it to the important legislative, trading and mining centres of Sucre, Tarija and Potosi. These cities will be connected with the rest of the world via the international telecommunications link which GEC has just installed between Bolivia and Peru.

For the extension system completely semiconductored 2 GHz microwave-radio and associated multiplex equipment will provide a bothway radio channel with a capacity of 960 telephone circuits and a standby or protection radio channel which can carry television programmes. A 6.8 GHz radio route will connect Potosi into the network via a passive reflector and the whole system will be equipped with a comprehensive supervisory indication and remote control system.

High reliability of microwave-radio equipment is particularly important when stations are sited at altitudes up

to 4880 m (16 000 ft), in ambient temperatures varying between -25°C (-13°F) and $+40^\circ\text{C}$ ($+104^\circ\text{F}$) and with external humidity varying from as low as 2% up to almost 100%.

The current Bolivian contract entails: initial survey of the proposed sites; construction of site access roads, many of them to mountain top areas; clearing and developing the sites and providing the massive foundations for the antenna towers; construction of the buildings and erection of the towers; programming and controlling all work to ensure that the sites are ready to receive the power-generating and microwave-radio equipment; installing and commissioning the whole system and then maintaining it for a period so that it can be handed over to ENTEL as a fully working system.

This contract will make an important addition to the development of telecommunications in South and Central America where GEC Telecommunications Limited has installed microwave-radio-equipped national trunk networks for Bolivia, Chile, Costa Rica, El Salvador, Guyana and Peru. These systems together with international connexions represent a total radio-channel length in Latin America alone of 30 000 km (18 650 miles).

The contract is in two distinct phases: the first phase—amplifying the facilities of the existing system—is scheduled for service early in 1978, and the extension to the South is due to be in service in 1980 which will integrate with other developments in that region. The equipment will be made in the company's Coventry factories.

Recent Developments in Data Modems

Organized by the Communications Group and held in London on 27th January 1977

The Colloquium opened with a survey in which Mr R. K. P. Galpin (Plessey Telecommunications), under the title of 'Transmission Aspects of Modem Design', discussed the development of data modems over the past ten years. The more conventional f.m. and p.m. modems can be used over telephone circuits at transmission rates of up to 4800 bit/s, and at higher transmission rates a linear modulation method is normally used together with an adaptive equalizer at the receiver. The linear modulation method generates either a vestigial sideband suppressed carrier a.m. (v.s.b.-a.m.) signal, or else a quadrature a.m. signal which comprises two double sideband suppressed carrier a.m. signals in phase quadrature and element synchronism. A pilot carrier must be transmitted with a v.s.b.-a.m. signal but is only required at the start of transmission with a q.a.m. signal. Special techniques must be used to extract symbol timing from the received data signal when this is severely distorted.

'The Modem as a System Component' by Mr J. Blackwell of Racal-Milgo (presented by Mr A. F. Payne) described how flexible modems can be used for multiplexing data streams with different transmission rates. A network of interconnected modems can be controlled by a central processor unit which may also be used to identify and correct faults and more sophisticated systems can be developed to test both modems and lines. Techniques such as network reconfiguration and dialled back-up can be used to maintain the operation of a communication network in the presence of faults.

In his paper on 'The Evaluation of Data Modem Performance by Computer Simulation', Mr J. T. L. Sharpe of Standard Telecommunications Laboratories described how computer simulation can evaluate the characteristics of telephone circuits and their effects on data modems. The distortion introduced by a telephone circuit is mainly a combination of amplitude, frequency and group delay frequency distortions. The effects of these on the data signal can be studied in terms of the equivalent baseband channel, the channel impulse response and the eye pattern of the demodulated data signal at the receiver. Both v.s.b. and q.a.m. signals have been studied. From the impulse response of the equivalent baseband channel, the eye pattern of the demodulated data signal can be determined and hence the distortion introduced into this signal. By comparing the distortions obtained with different modulation methods, carrier frequencies, transmission rates, and so on, it is possible to optimize the modem design.

Mr M. N. Y. Shum of Case Computer and Systems Engineering, described a 4800 bit/s modem in his paper on 'A Microprocessor Based Modem'. In this modem, a micro-computer based on a microprocessor is used to perform the operations of signal shaping, modulation, demodulation, carrier and timing recovery, adjustment of the equalizer,

scrambling and descrambling of the data signal, and automatic testing of the line. The microprocessor has a remarkable capacity for processing signals and is both flexible and cheap. There is clearly a great future for the use of microprocessors in data modems.

Finally, Dr A. P. Clark of the Department of Electronic and Electrical Engineering of Loughborough University of Technology, in his paper on 'The Way Ahead', compared linear and non-linear transversal equalizers with more sophisticated detection processes. In the presence of amplitude distortion, a non-linear equalizer can achieve a better tolerance to additive white Gaussian noise than can a linear equalizer, and the non-linear equalizer itself involves a sub-optimum detection process. Ideally a maximum-likelihood detector is the most suitable arrangement for transmission rates above 4800 bit/s. Two practical approximations to such a detector are the Viterbi-algorithm detector and the arrangement of 'detection and signal cancellation'. Each of these can be made to give a near optimum performance but they are implemented quite differently. It seems likely that the technique that is finally adopted will be some modification or development of one or other of these two systems, operating with a 4- or 16-point q.a.m. signal.

A. P. CLARK

Standard Frequency Transmissions—April 1977

(Communication from the National Physical Laboratory)

April 1977	Relative Phase Readings in Microseconds NPL—Station (Readings at 1500 UT)		
	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	6.1	9.1	-14.3
2	6.1	9.0	-14.0
3	6.1	9.2	-13.7
4	5.9	8.8	-13.4
5	5.7	8.8	-13.3
6	5.9	8.6	-12.9
7	6.1	8.8	-12.5
8	5.8	8.6	-12.3
9	5.8	8.4	-12.1
10	5.6	8.5	-12.1
11	5.6	8.1	-11.6
12	5.8	8.1	-11.2
13	5.8	8.1	-10.6
14	5.8	8.4	-9.6
15	5.6	8.1	-8.9
16	5.6	8.2	-8.1
17	5.6	7.8	-7.7
18	5.6	7.8	-6.9
19	5.6	7.2	-6.5
20	5.4	7.3	-6.1
21	5.4	7.1	-5.4
22	5.6	6.8	-4.4
23	5.8	7.0	-3.4
24	5.6	7.0	-2.6
25	5.4	7.0	-2.2
26	5.4	7.3	-1.3
27	5.3	7.1	-0.0
28	5.4	6.8	+1.0
29	5.2	6.3	+2.2
30	5.4	6.5	+3.6

Notes: (a) Relative to UTC scale $(UTC_{NPL}-5_{station}) = +10$ at 1500 UT, 1st January 1977.

(b) The convention followed is that a decrease in phase reading represents an increase in frequency.

(c) Phase differences may be converted to frequency differences by using fact that $1 \mu s$ represents a frequency change of 1 part in 10^{11} per day.

Members' Appointments

CORPORATE MEMBERS

Professor D. W. Lewin (Fellow 1974, Member 1960, Graduate 1957) who has been Head of the Department of Electrical Engineering and Electronics at Brunel University since 1974, and Professor of Digital Processes since January 1972, has been awarded the degree of Doctor of Science by the University. Professor Lewin has served on the Institution's Council, and has been a member of the Papers Committee since 1971; he is currently Chairman of the Computer Group Committee and has contributed a number of papers to the Journal.

Professor D. R. Towill (Fellow 1970) has been awarded the degree of Doctor of Science by the University of Birmingham. He has been at the University of Wales Institute of Science and Technology since 1966, where he founded the Dynamic Analysis Group in the Department of Mechanical Engineering and Engineering Production, in which he now holds the Chair of Engineering Production. Professor Towill has contributed numerous papers to the Institution's Journal, five of which have gained him Premiums, and he has served as Chairman of the Automation and Control Systems Group Committee.

Major M. S. Body, REME (Member 1967, Graduate 1958) has been appointed Electronic Project Officer with 34 Central Workshop REME at Donnington, Shropshire. He was previously Officer in Charge, Radar Production, with 35 Central Workshop REME at Old Dalby, Leicestershire.

Mr. R. F. Burns (Member 1973, Graduate 1964) who joined Rank Pullin Controls in 1976 as Group Leader, has been promoted to Chief Engineer, Civil Telecommunications Division, with responsibility for the design and production of v.h.f. and u.h.f. radio telecommunication equipment.

Wing Cdr. J. Granger, RAF (Ret.) (Member 1952, Associate 1947) has retired from the Royal Air Force, where he was serving in the Directorate of Signals (Air), Ministry of Defence, and is now a Technical Officer with the British Standards Institution in London.

Mr. J. B. Gray (Member 1967, Graduate 1963) who has been with the British Broadcasting Corporation since 1962, has been seconded to the Kranji Transmitting Station, Singapore, as Project Manager.

Mr. J. P. Hawkins (Member 1966, Graduate 1961) has taken up the post of General Manager of Forsyth Brothers Ltd., Manchester. He was formerly Computer Sales Office Manager with Honeywell Information Systems Ltd.

Mr. T. Jacobs, B.Sc. (Member 1970) is now Marketing Manager in the Data Processing Electronics Division of Mullard Ltd., London. Mr. Jacobs joined Mullard Research Laboratories in 1953 as an engineer working on colour television research and development, and since then has held a number of progressively senior positions with the company.

Mr. P. J. Lewis (Member 1976, Graduate 1970) who first joined the British Steel Corporation in 1964 as an Electrical Apprentice, is now a Shift Electrical Engineer at the Orb Works, Newport, Gwent. He was previously a Process Control Engineer with the Corporation concerned with development projects and systems design.

Mr. T. Lomas, E.R.D. (Member 1963) has been appointed to the position of Deputy Controller of Contracts, Post Officer Telecommunications Headquarters. He was formerly Head of Management Sciences Division at the same headquarters.

Mr. A. H. Lord (Member 1973, Graduate 1969) who has been with EMI Ltd. since 1967, has been appointed Manager of the Product Support Group of EMI Medical Ltd., concerned with technical information support for service engineers on the EMI-Scanner.

Sqdn Ldr V. J. Ludlow, M.Sc., RAF (Ret.) (Member 1976) has recently retired from the Education Branch of the Royal Air Force to take up an appointment as Trials Engineer with EMI Ltd., Weston-super-Mare. From 1972 to 1976 he was Officer Commanding Radar Applied Principles Flight at No. 1 Radio School, RAF Locking.

Mr. A. Miller (Member 1973, Graduate 1968) has taken up an appointment with the Ministry of Overseas Development at Kandilli Observatory, Cengelköy, Istanbul, where his work involves assisting the Turkish Government in the setting up of a telemetered seismic data network around the Sea of Marmara. Mr. Miller has been with the Institute of Geological Sciences, Edinburgh, since 1967, and from 1969 to 1972 he was Engineer in Charge of the development of a seismic array network based in Brasilia, South America.

Mr. M. R. Perkins (Member 1972, Graduate 1970) has joined Texas Instruments, Bedford, as Account Manager for the South West Area. From 1971 to 1977 he was with F.R. Electronics Ltd., Wimborne, latterly as Principal Engineer.

Mr. B. G. Smee (Member 1971, Graduate 1966) is now Marketing Manager with the Valiant Electrical Wholesale Company, London. He was previously Product Manager with the company, which he joined in 1973.

Mr. R. V. Whittle (Member 1972, Graduate 1966) has taken up an appointment as Depot Quality Manager of the Royal Naval Armament Depot, Plymouth. He was previously a Senior Weapons Systems Engineer (Polaris) at the Royal Naval Armament Depot, Coulport, Dunbartonshire.

Lt. Col. W. L. Wood, R.Sigs (Member 1971) who since 1974 has been Trials Director at the School of Signals, Blandford Camp, Dorset, has been appointed GSO1 Scientific Advisory Group (Army) with the Ministry of Defence in London.

Lt. R. J. Wright, RN (Member 1973, Graduate 1970) who has been an Assistant Electrical Engineer with the Ministry of Defence since 1974, has joined HMS *Eskimo* as Weapons Electrical Engineer Officer.

NON-CORPORATE MEMBERS

Mr. J. E. Gillard (Graduate 1967) who is with J. & S. Sieger Ltd., Poole, Dorset, has been promoted from Senior Engineer (Section Leader) to Principal Engineer (Project Leader).

Correction: It is regretted that the name of **Major R. M. A. Joy, B.Sc.(Eng.), REME** (Graduate 1971) under 'Members' Appointments' in the April 1977 issue (p. 190) was given incorrectly.

Mr. W. J. Lindley (Graduate 1970) who was previously a Technical Officer in the Engineering Production Department of the University of Birmingham, is now a Technician Engineer in the Telemetry Section of the South Staffordshire Waterworks Company Ltd., Birmingham.

Mr. I. M. March (Graduate 1968) is now North and Midlands Area Supervisor for EMI Medical Ltd., Slough, with responsibility for the service and installation of computerized X-ray equipment. He was previously a Senior Service Engineer with the company.

Mr. A. K. Sarin (Graduate 1965) has been appointed Manager, Calibration Services, at the Canadian Forces base in Lahr, West Germany. He was previously Engineer in Charge of the Calibration Centre at the Canadian Forces base, Chatham, New Brunswick.

Staff Sgt. P. J. Selby, R.Sigs (Graduate 1970) has returned to the UK from 4th Division Headquarters Signal Regiment BAOR, where he had served since 1974, to become Foreman of Signals in the UK Static Communication System with 241 Signal Squadron at Donnington.

Mr. P. W. Justin, (Associate Member 1976) who was previously a Flight Simulator Engineer with BAC, has been appointed Chief Engineer Avionics with Airwork Services Ltd. at Ruwi, Sultanate of Oman.

Mr. A. R. Abbott (Associate 1975) is now a Senior Avionic Systems Design Engineer with BAC Military Aircraft Division, Warton Aerodrome, near Preston. He has been with the company since shortly after leaving the RAF in 1970

Recent Accessions to the Library

This list of additions to the Library covers the period November 1976 to May 1977. These books may be borrowed by members in the British Isles by personal call or by post; information on loan conditions can be obtained from the Librarian, Miss E. J. Keely.

Communication systems

- 'Modern Communication Systems'. Coates, R. F. W., *Macmillan*, London, 1975.
- 'The Origin of Maritime Radio. The story of the introduction of wireless telegraphy in the Royal Navy between 1896 and 1900.' Pocock, R. F. and Garratt, G. R. *H.M.S.O.*, 1972.
- 'Telecommunication Economics'. Morgan, T. J., 2nd ed., *Technicopy*, Gloucester, 1976.
- 'Telegraphy.' Renton, R. N., *Pitman*, London, 1976.
- 'Digital Transmission Systems.' Bylanski, P. and Ingram, D. G. *Peter Peregrinus*, Stevenage, 1976.
- 'The Radio Amateur's Handbook.' American Radio Relay League, 1976, 1977.
- 'Radio and Line Transmission (B)'. Green, D. S., *Pitman*, London, 1971.

Computers

- 'Economic Evaluation of Computer Based Systems. Book 2, Computers and the manager. Blaxter, C. W. and Morris, W., Workshop Report. N.C.C.L., 1971.
- 'Scientific Analysis on the Pocket Calculator.' Smith, Jon M., *Wiley-Interscience*, London, 1975.
- 'Electronic Computers'. Hollingdale, S. H. and Tootill, G. C., 3rd rev. ed. *Pengium*, Harmondsworth, 1975.
- 'Introduction to Programming and Computer Science.' Ralston, Anthony, *McGraw-Hill*, Tokyo, 1971.

Electronic Engineering

- 'Digital Electronic Circuits and Systems.' Morris, Noel M., *Macmillan*, London 1974.
- 'Designers Guide on Electromagnetic Compatibility'. Electronic Industries Association, New York.
- 'Electronic Measurements and Instrumentation.' Oliver, Bernard M. and Cage, John M., *McGraw-Hill*, Tokyo, 1971.
- 'The Principles and Techniques of Engineering Estimating.' Calder, Granville, *Pergamon*, Oxford, 1976.
- 'Large Scale Integration.' Altman, L. (ed.), *McGraw-Hill*, New York, 1976.
- 'Noise Reduction Techniques in Electronic Systems.' Ott, H., *Wiley*, New York, 1976.
- 'Television Theory and Servicing: Black/White and Colour.' Herrick, Clyde N., *Reston*, Virginia, 1976.
- 'Thermocouple Temperature Measurement.' Kinzie, P. A. and Moore, David R., *Wiley Interscience*, London, 1973.

- 'Field Effect Transistors in Integrated Circuits.' Wallmark, J. T. and Carlstedt, L. G., *Macmillan*, London, 1974.
- 'Introduction to Switching Theory and Logical Design.' Hill, Frederick J. and Peterson, Gerald R., 2nd ed. *Wiley International*, New York, 1968.
- 'Electronics Reliability—Calculation and Design.' Dummer, Geoffrey W. A. and Griffin, Norman B., *Pergamon*, Oxford, 1976.

Miscellaneous

- 'Operations Research: An Introduction.' Taha, Hamdy A., 2nd ed. *Collier-Macmillan*, London, 1976.
- 'A Treatise on Electricity and Magnetism.' Maxwell, James Clerk, Vol. 2. Unabridged, 3rd ed. *Dover*, New York, 1954.
- 'Eurosace: Europe and Space. An assessment and prospects.' Eurosace, Paris, 1973.
- 'Acoustic Techniques for Home and Studio.' Everest, Alton F., *Foulsham-Tab*, Slough, 1973.
- 'A Conspectus of Electrical Timekeeping.' Aked, Charles K., Antiquarian Horological Society, London, 1977.

Conference Proceedings

- Technology of Scientific Space Experiments. International Conference, Paris, 26th-30th May 1975. Centre National d'Etudes Spaciales.
- 'Perspectives in Biomedical Engineering.' Kenedi, R. M. (ed.), Proceedings of a symposium held in the University of Strathclyde, Glasgow, June 1972. *Macmillan*, 1973.
- 'A Discussion on the Scientific Results from the *Prospero* and *Ariel 4* Satellites.' *Proceedings of the Royal Society of London. A. Mathematical and Physical Sciences*, Vol. 343, pp. 159-287, No. 1633. The Royal Society, London, 1975.

Reference Works

- The Computer User's Yearbook 1976.
- Guide to Official Statistics. H.M. Stationery Office, 1976.
- Report of the Committee on the Future of Broadcasting. (Chairman: Lord Annan), 1977.
- 'A Guide to International Recommendations on Names and Symbols for Quantities and on Units of Measurement.' Lowe, Armstrong D., Geneva, W.H.O., 1975.
- The Library receives the D.A.T.A. series of publications produced by Derivation and Tabulation Inc., which are for reference in the library only.

The following issues are available:

- Discontinued Diodes (1975-76) 1st ed.
- Discontinued Integrated Circuits (1976-77) 6th ed.
- Discontinued Semiconductor Diodes (1976) 2nd ed.
- Microcomputer Data Book (1976) 1st ed.
- Microwave Tube Data Book (Autumn 1976) 37th ed.
- MSI-LSI Memory Data Book (Spring 1975) 8th ed.
- MSI-LSI Memory Data Book (Spring 1976) 10th ed.
- MSI-LSI Memory Data Book, Vol. 2 (1976) 11th ed.
- Optoelectronics Data Book (Spring 1976) 3rd ed.
- Power Semiconductor Data Book, Vol. 2 (1976) 5th ed.
- Relay Data Book (Miniature & Sub-Miniature) (1975-76) 4th ed.
- Semiconductor Application Notes Data Book (Spring 1976) 9th ed.
- Semiconductor Application Notes (Autumn 1976) 10th ed.
- Semiconductor Diodes Data Book (Spring 1976) 37th ed.
- Semiconductor Diodes Data Book (Autumn 1976) 38th ed.
- Thyristor Data Book (Autumn 1976) 7th ed.
- Transistor Data Book (Autumn 1976) 41st ed.

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meeting on 12th May 1977 recommended to the Council the election and transfer of the following candidates. In accordance with Bye-law 23, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

Meeting: 12th May 1977 (Membership Approval List No. 234)

GREAT BRITAIN AND IRELAND CORPORATE MEMBERS

Transfer from Member to Fellow

BURROWS, Keith. *Maidenhead, Berkshire.*

Direct Election to Fellow

DOLLIMORE, Reginald Paul. *Watford, Hertfordshire.*

Transfer from Student to Member

FURNESS, Roger Keith. *Twyford, Berkshire.*

Transfer from Graduate to Member

ARMSTRONG, Alan. *Morpeth, Northumberland.*
WILSON, Alexander Mackay. *Ewell, Surrey.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

GILLMAN, David Kimberley. *Billericay, Essex.*
OGBU, Christopher Omoche. *London.*
PRESTON, Roger Graham. *Stockport, Cheshire.*

Transfer from Student to Associate Member

McBRIDE, John. *Croydon, Surrey.*

Direct Election to Associate Member

BROWN, Alan Peter. *Wool, Dorset.*

STUDENTS REGISTERED

BADCOCK, Alan Raydon Cooper. *Tavistock, Devon.*

BROOKS, Stephen Edward. *Reading, Berkshire.*

CRESWELL, Hugh Thomas. *Littlehampton, Sussex.*

LIN, Wai Kwong, Edmond. *Brighton, Sussex.*

SMITH, Robert Arthur. *Southampton, Hampshire.*

UNO, Nduonyi Efong Offisong. *London.*

Direct Election to Member

BUTCHART, Robert. *Salisbury, Rhodesia.*

CARDNELL, David Robin. *North Rockhampton, Queensland, Australia.*

OLUMBA, Amaraegbu Asic. *Owerri, Nigeria.*

NON-CORPORATE MEMBERS

Direct Election to Associate Member

BAKHSHI, Manohar Singh. *Ndola, Zambia.*

SO, Patrick Kin-Kun. *Hong Kong.*

TWOMEY, John David. *Willowdale, Ontario, Canada.*

STUDENTS REGISTERED

CHEUNG, Ping Chow. *Hong Kong.*

HAM, Seng Peng. *Singapore.*

HO, Pak Kin. *Singapore.*

LI, Kin Man. *Hong Kong.*

RAY, Ajit Kumar. *Calcutta, India.*

Financial Surveys of UK Electronics Industry

Studies of major parts of the British electronics industry have been published by Jordan Dataquest Ltd. covering, generally, the last two years (1975 and 1974) for which results have been filed at Companies House. The areas covered in the six surveys are: Component Manufacturers; Component Distributors; Instruments and Communications; Computer-related Companies; Consumer Product Manufacturers; Consumer Product Distributors.

Each survey reveals the salient financial statistics of some 200-300 companies (332 in the case of the more fragmented instruments industry). The data provided on each company include sales turnover, profits, fixed and current assets, liabilities and bank loans and an indication of ownership. The companies are ranked in descending sales order, and cross-referenced alphabetically in an index which also gives the name of the chief executive, the company's address and a description of its activities.

Some 270 companies were included in the Component Manufacturers survey and outstanding companies from a performance viewpoint were: Vero Electronics Ltd. (sales £3,941,000, pre-tax profits £291,000), Tectonic Electronics Ltd. (sales £1,974,000, pre-tax profits £262,000), Whiteley Electrical Radio Co. Ltd. (sales £1,774,000, pre-tax profits £214,000).

The survey of the Instruments and Communications industry points out that the sector recently passed through a problematic phase with heavy write-offs, low profitability and the need for external financial support, for example in the case of George Kent Ltd. However, the authors conclude that, thanks to high export performance and a depreciating pound, the sector has performed surprisingly well of late.

The outstanding company selected is Coulter Electronics

Ltd. (turnover £7,859,000, exports 70%) whose sales rose 31.5% and whose profit margins are an astounding 21.0%.

Some 180 firms are covered under the heading of Electronics Consumer Product Manufacturers, although a further 250 company files were searched in vain for their figures. As well as manufacturers of electrical consumer products the analysts have included major domestic electrical appliance makers and companies operating in smaller areas.

The performance of the privately owned firms in this sector is not considered to be particularly encouraging: 5 out of the 12 largest manufacturers were losing money, but the survey singles out Roberts Radio Co. Ltd. as competing particularly profitably against large UK firms as well as those overseas.

According to the survey of the Computers and Data Processing Manufacturing and Service industry foreign owned companies predominate—95 out of 237 and furthermore, 18 of the largest firms are foreign owned. Over 25% of the firms made losses in 1975. The survey comments on good performance by ICL among British owned companies while of those from abroad, Hewlett-Packard is regarded as outstanding.

Individual surveys cost £28; but if two surveys are purchased, the cost is £23 each; for 3 or more surveys the unit price is £20 only. They are available direct from the publishers: Jordan Dataquest, Jordan House, 47 Brunswick Place, London N1 6EE. Tel. 01-253 3030. The company published 'Britain's Top 1000 Electronics Companies 1975', predecessor of the present Surveys.

Jordan Dataquest also market studies by DAFSA Analyse SA of France, including 'Data Processing in the World' (price £350) and 'L'Electronique Medicale' (price £200).