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

Volume 44 No. 7 July 1974

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

Is a Degree Essential?

ALL of the fifteen Chartered Institutions which, collectively, comprise the Council of Engineering Institutions, have ruled that a degree from a University is a desirable prerequisite to securing the cachet of 'Chartered Engineer'.

In the IERE Conference on 'The Electronics Industry and Higher Education', *Professor G. D. Sims made the point that 'Advanced technological industries make greater demands on the supply of highly qualified manpower than average. Not surprisingly the electronics industry employs a higher proportion of graduates than other branches of manufacturing industry'. This statement poses the question as to whether the output of University students who have graduated as a result of studies of radio and electronic engineering or as an adjunct to courses in physics, electrical or mechanical engineering etc., is sufficient to meet the demand. This is not borne out by total graduate output as described, or the vacancies currently advertised.

If such thinking and conferences are dominated by the University approach, is it because Polytechnics and other Technical Institutes are unsure of their future role in training beyond the technician level? There has been some amelioration by CEI recognizing CNAA degrees, but will this be enough?

Another school of thought, however, raises doubt as to whether Great Britain can afford to maintain—let alone extend—existing Universities. In such events can Great Britain afford to relegate other institutions of advanced teaching to a lower role than encouraging every engineering student to secure recognition as a chartered engineer?

It will still take some years for the CEI examination to be accepted as an alternative to a University degree as a means of securing the basic academic qualifications. If, however, there is a limitation on University entrants, then it behoves CEI to extend recognition to other than University degrees—and on an extensive scale if the net is to catch such pioneers of technical innovation as have hitherto just missed or were ultimately recognized by membership of the individual Institutions.

An expanding electronics industry urgently needs men capable of innovation, based upon an understanding of existing knowledge. Economy in University expansion need not imply economy of opportunity; all that is needed is recognition of the alternative sources of acquiring knowledge so as to recognize ability to advance the science or technique of his chosen profession. Not least, to understand that no man is the complete engineer but tends in study and in work to do that which attracts him most in enthusiasm and interest—now described as job interest.

It is this 'interest' which motivates a man to use his own spare time or to gain employer interest in giving him part-time day release to acquire the necessary knowledge to meet academic requirements of his professional body. Does this not therefore provoke argument on the need—and a very urgent need—to support recruitment from other than Universities of the embryo engineer (a University graduate is not an engineer) in order to advance the practice of radio and electronic engineering and indeed other branches of applied science?

G.D.C.

* A report of the Conference is given on page 352.

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Mr. R. A. Shaw graduated B.Sc. in physics (1968) and B.E. (Hons). in electrical engineering (1970) from the University of Auckland, New Zealand. He subsequently worked on 11-kV systems design for an Auckland consulting firm, Kerslake and Partners. In 1971 he became a graduate teaching assistant at the University of Alberta, Canada, completing an M.Eng. in 1972. Following this he worked on the Communications Technology Satellite programme at RCA Aerospace, Montréal. He is currently a systems design engineer in digital microwave communications with Canadian Marconi, Montréal.

Dr. David P. Morgan received the B.A. degree in physics from the University of Cambridge in 1962. In 1966 he obtained the M.Sc. degree from Chelsea College, London, where he studied helicon waves in solids. He later did research on dispersive delay lines using surface acoustic waves at University College London, and obtained the Ph.D. degree in 1969. In 1970–71 he spent one year at the Central Research Laboratories of the Nippon Electric Company, in Kawasaki, Japan, where he worked on surface acoustic wave filters. He is now a Research Fellow at the University of Edinburgh, Scotland, where his main interest is in surface acoustic wave convolvers using non-linear interactions.

Dr. D. Routledge graduated in engineering physics from the University of Alberta, Canada in 1964. After a few months' work in microwave devices with Bell Northern, Ottawa, he enrolled at Queen's University in Kingston, where he earned an M.Sc. and Ph.D. in radio astronomy. In 1969 he returned to the University of Alberta with a Killam post-doctoral fellowship, and was appointed assistant professor in electrical engineering in 1971. His current research interests are decametric radio astronomy and wildlife radio tracking.

Professor Jeffrey H. Collins (Fellow 1972) received the B.Sc. degree in physics and the M.Sc. degree in mathematics from the University of London, in 1951 and 1954. From 1951 to 1956, his experience in microwave tubes and ferrite parametric amplifiers was obtained during employment at the GEC Hirst Research Centre, Wembley, and with Ferranti Ltd, in Edinburgh. From 1957 to 1967, he was with the Electrical Engineering Department of the University of Glasgow, where he taught in the fields of network theory and materials science, and researched microwave ferrites and microwave acoustics. During the scholastic years 1966–1968, he was a research engineer in the W. W. Hansen Laboratories of Physics, Stanford University, California, engaged in research on surface acoustic waves, magnetic garnet delay lines, opto-acoustic interactions, and pulse compression applications. From 1968 to 1970 he was Director of Physical Sciences at the Electronics Division of Rockwell International, California. In 1970 he returned to Scotland, where he is currently Professor of Industrial Electronics at the University of Edinburgh.

Mr. R. Dennis Lambert received his O.N.C. and H.N.C. in applied physics in 1965 and 1969 by part-time studies at Napier Technical College, Edinburgh. From 1960 to 1971 he gained considerable experience in semiconductor and thin film fabrication with Emihus Microcomponents, Glenrothes, Scotland. Following early experience on point contact and gold bonded diode fabrication, he became senior production engineer in 1970 with full responsibility for planar diode wafer assembly. In 1971 he joined the University of Edinburgh as a Technical Officer (Research) responsible for the processing assembly and packaging of SAW devices and associated microelectronic circuitry. Mr. Lambert is now a production engineer with General Instruments Microelectronics, Glenrothes.

*See also page 382.

UDC 534.21: 621.372.54: 621.3.049.776

Programmable surface acoustic wave devices utilizing hybrid microelectronic techniques

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SUMMARY

Programmable matched filters can be realized using surface acoustic wave propagation in conjunction with external microelectronic circuitry. Here two devices, a programmable analogue matched filter and a diode convolver, are discussed, highlighting the similarities in constructional technique. Specific device requirements include an r.f. circuit which for 10 MHz signal bandwidth must possess a 310 μm pitch. Factors influencing the choice of the thin film hybrid microelectronic technique are outlined and the design, fabrication and performance details are described.

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1 Introduction

Many radar and communication systems[1,2] now utilize signal formats consisting of sequences of coded waveforms, or *signatures*, to overcome propagation effects, improve detection probabilities and permit multiple access capability. Signature generation (expansion) in the transmitter and detection (compression) at the receiver can be performed with a suitable filter designed to match to the transmitted signal. Such a matched filter[2] has an impulse response which is the time-reverse of the signature and thus performs correlation. The correlation process selectivity enhances the required signal as compared to noise and interference, giving a characteristic narrow correlation peak preceded and followed by a number of time sidelobes when the desired signal is received. The particular waveforms considered here are phase-shift keyed (p.s.k.) modulated r.f. carriers where each of the contiguous time segments of modulating code is known as a chip. For a binary coded signal the r.f. phase is therefore 0 or π rad during each chip period depending on the particular code used.

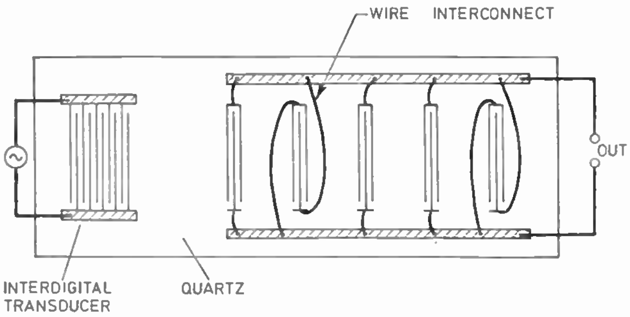
This paper examines the design and construction of matched filters, fabricated by combining thin film hybrid and surface acoustic wave (SAW) technology. Specifically, programmable devices are considered where the response can be electronically controlled to correlate a large variety of coded waveforms. Programmability has wide ranging attractions including: high level security, adaptability to overcome variable interference, and multiple address capability—facets which cannot be conveniently achieved with a bank of fixed coded filters. The devices under consideration are the programmable analogue matched filter a.m.f. and diode convolver, which pose similar technological and fabrication problems and are conveniently fabricated with hybrid microelectronic technology.

Section 2 details the background to realizing matched filters with SAW techniques and describes the implementation and performance of the two devices with conventional discrete component hardware. Sections 3 and 4 describe technological considerations and detail the design and construction of a hybrid programmable a.m.f., then outline the performance achieved and compare SAW a.m.f.s against competing technologies. In conclusion, Section 5 outlines the future programme required to evaluate applications potential of these devices.

2 Surface Acoustic Wave Programmable Matched Filters

SAW devices are generally constructed using polished piezoelectric substrates, where the acoustic wave is launched by the application of a suitable electric field.[3] This is achieved efficiently by the interdigital transducer, shown schematically in Fig. 1, which consists of interleaved metal electrodes photoetched from a deposited metal film. The transducer, which can also detect these waves, applies a field whose spatial periodicity corresponds to the SAW wavelength, typically 30 μm for a frequency of 100 MHz.

Figure 1 shows the structure of an a.m.f., whose



(a) Fixed-coded analogue matched filter.

(b) Impulse response.

Fig. 1

impulse response is required to be the p.s.k. waveform shown. When an impulse is applied at the left transducer, a surface wave packet with a corresponding number of periods is produced. The packet is subsequently sampled in turn by a series of taps, which are simply transducers with a relatively small number of electrodes. When the taps are spaced by a distance corresponding to the duration of the wave packet, the device expands the impulse into a continuous p.s.k. waveform where the phase of the signal from an individual tap is determined simply by the polarity of the connexions as shown in Fig. 1.

The substrate material is usually chosen to be quartz, since the ST-cut of this material gives a very low temperature coefficient of delay (less than 3 parts/10^6/deg C). The weak electromechanical coupling in this material helps to reduce second-order effects, an important factor since the number of electrodes is large. On the other hand, it leads to relatively large insertion loss, typically 50 dB in expansion. The SAW velocity, 3158 m/s, implies attractively compact size (<5 cm) for a device with 13 μs delay.

2.1 Programmable Analogue Matched Filters

The a.m.f. can be made programmable by using an electronic switching network between the taps and summing bus to control the phase contribution of individual taps, in accordance with a binary code stored in an associated shift register (Fig. 2). Reprogramming to a new code is accomplished simply by reading the desired code from a read-only-memory (r.o.m.) which stores a library of possible codes. Switching and summing of the r.f. signals requires a high performance switching network, in contrast to the remaining logic circuitry which simply provides d.c. levels except when reprogramming.

In certain applications there are however requirements for a fast reprogramming capability, necessitating consideration of both switching network and code store transient speeds.

Programmable a.m.f.s have been constructed using the 4 diode ring circuit of Fig. 3 to switch each tap. One device[4] with 90 MHz centre frequency included 7 taps at 1·24 mm pitch (400 ns delay) using discrete electronic components (glass encapsulated diodes) in printed circuitboard construction. When correlating a 7-chip Barker coded p.s.k. waveform the ratio of correlation peak to side-lobes was within 1·5 dB of the theoretical value. The insertion loss in expansion (60 dB) was however 10 dB greater than that exhibited by the fixed-coded version of the same a.m.f. An alternative switching circuit proposed by Hunsinger and Franck[5] which uses fewer components, gave a similar performance with only 5 dB additional insertion loss over a fixed coded device. Despite these encouraging results, this hard-wired circuitry is not suitable for high-performance wide-band devices (10 MHz chip rate). The small tap-to-tap spacing (310 μm) implies that discrete r.f. circuitry needs to be fanned out to overcome layout difficulties, leading to inferior performance. An alternative technology is therefore required in which the tap switches can be fabricated in a size compatible with the tap spacing.

2.2 Diode Convolvers

An alternative type of SAW device, the diode convolver, was recently introduced by Reeder.[6] This device has a pair of interdigital transducers and an array of regularly spaced taps with each tap connected to a diode (Fig. 4). In operation *two* input waveforms, signal and reference, are applied, with frequencies f_1 and f_2. The contra-directed surface waves generated are sampled by the taps. At each tap the non-linearity of the diode causes an output signal, at frequency $f_3 = f_1 + f_2$, proportional to the product of the amplitudes of the two surface waves. Suitable choice of f_1 and f_2 enables the harmonics to be filtered from the required signal at frequency f_3.

The input waveforms are assumed to have finite duration, such that the SAW wave packets produced have a spatial extent less than that of the tap array. If the signal waveform is a short r.f. pulse, the device action is

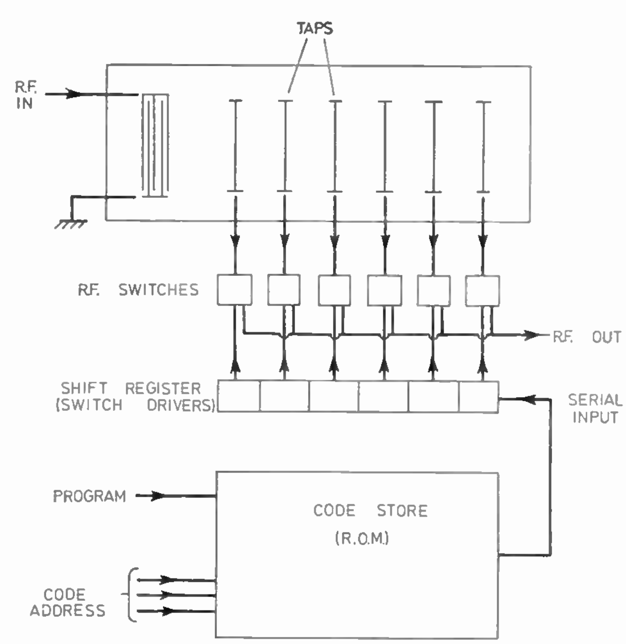


Fig. 2. Schematic of programmable a.m.f. with code storage and selection peripherals.

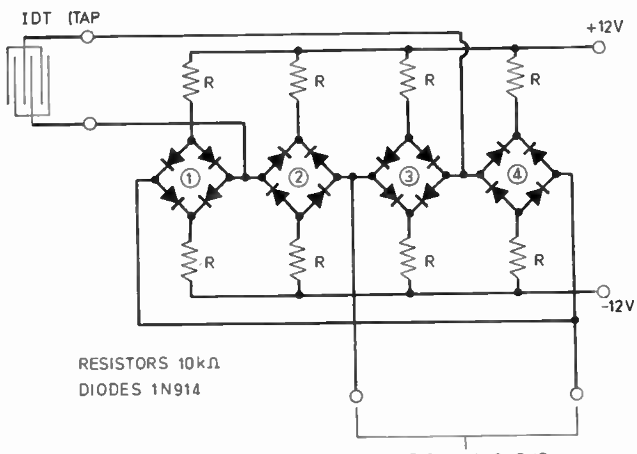


Fig. 3. Diode ring circuit for switching polarity of individual tap.

analogous to scanning the reference waveform by means of the signal. Thus the output, which can be regarded as the impulse response, corresponds to the reference waveform, with a time-contraction by a factor of 2 due to the relative motion of the waves. Further, the device is linear with respect to variations of the signal waveform. It thus acts much like a filter whose impulse response is determined by the reference input. To correlate a known signal waveform the reference needs to be the time-reversed waveform, and clearly a high degree of programmability is obtained since the reference can be varied at will. The bandwidth is limited by the sampling action of the taps, and in general the maximum signal bandwidth possible is $1/2T$, where T is the time interval corresponding to tap spacing. However a p.s.k. waveform with a larger bandwidth can be handled provided its chip period is close to T.

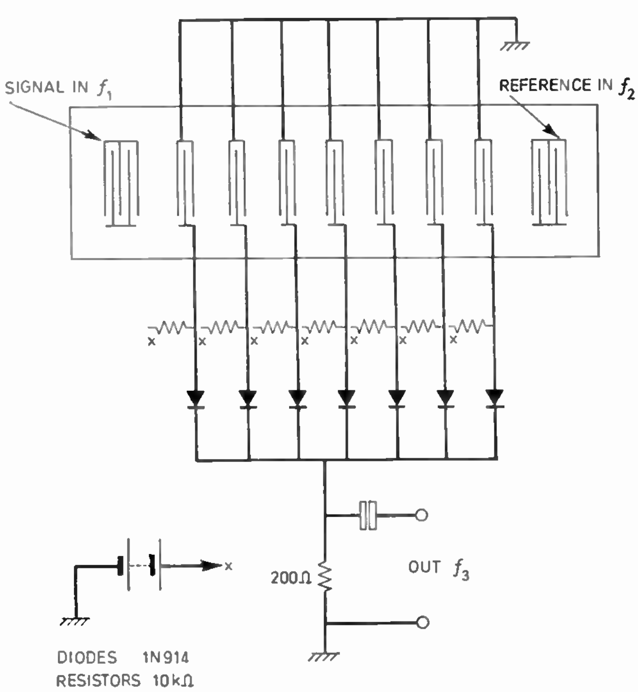


Fig. 4. Diode convolver.

Although Reeder[6] used a series chain of diodes, we have found that the arrangement of Fig. 4 gives better fidelity when the number of taps exceeds about ten. A d.c. bias of about 0·1 mA improves the non-linear efficiency of the diodes. Lithium niobate was used for the substrate, since its high piezoelectric coupling gives good transduction efficiency. The SAW velocity, 3485 m/s is similar to that of quartz. Figure 5 shows a result obtained with a 15-tap device with 200 ns propagation time between taps, using discrete diodes and resistors. The two input waveforms were p.s.k. coded signatures with carrier frequencies of 85 and 115 MHz, and bandwidth of 5 MHz. The output correlation waveform shown had a centre frequency of 200 MHz.

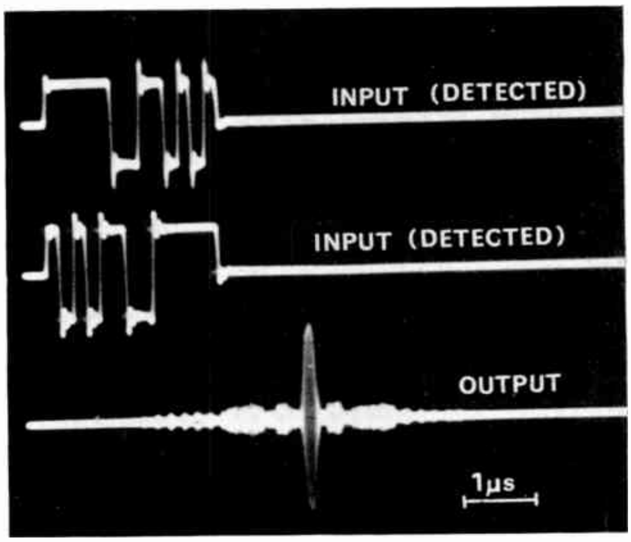


Fig. 5. Diode convolver performance. Input waveforms are modulated according to 13-chip Barker code and its time-reverse, and are synchronously detected for the display to show their structure. The theoretical peak-to-sidelobe ratio for the output is 22·3 dB.

For comparison with the a.m.f. we consider the use of a convolver to correlate a signal waveform whose timing is unknown. This can be done using a repetitive reference waveform[7,8] as shown in Fig. 6. The convolver system gives rise to a timing distortion which can only be removed at the expense of more complicated peripherals, although it can handle a greater variety of signal input waveforms. A much simpler system, using a single reference waveform, can be utilized when the signal timing is known.

3 Selection and Detail of Fabrication Techniques

3.1 Device Design Constraints

Sections 3.1 and 3.2 consider the typical requirements for a 127-tap device capable of handling codes at a 10 MHz chip rate, thus requiring a 310 μm tap spacing on a quartz substrate. Since discrete circuitry is excluded (Section 2.2) only integrated and hybrid circuitry are considered. In this context, the prohibitive size of the capacitors excludes the relatively simple Hunsinger–Franck circuit. Alternative circuits, for example that of Fig. 3, are more complex and unsuitable for hybrid construction, though this does not preclude integrated circuit (i.c.) construction.

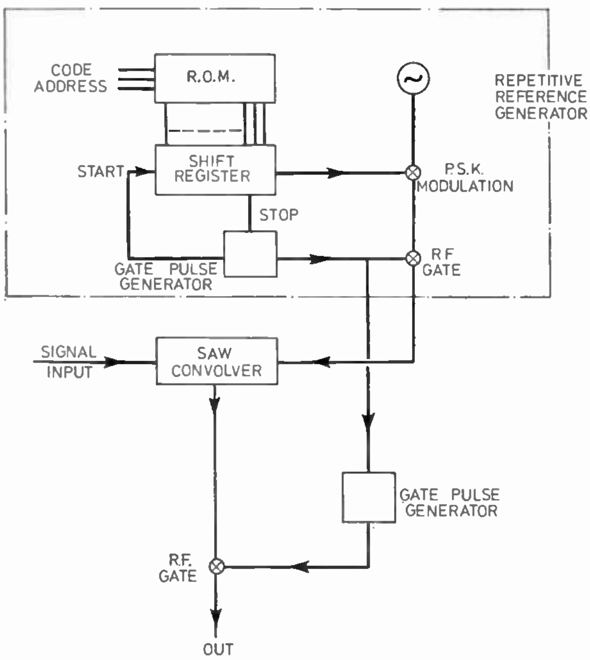


Fig. 6. Use of convolver to correlate arbitrary p.s.k. coded waveforms.

An alternative approach, however, leads to a much simpler circuit which can be implemented in either technology. If the tapped delay line (t.d.l.) is designed with a dual tap geometry (Fig. 7), where the taps simultaneously sample one half of the wavefront, suitable interconnexion can be arranged to output both 0 and π phase signals to the selective switching matrix. The switching circuit is therefore only required to select the appropriate output and isolate the other one. To minimize bonding complexities and provide more space for the switching circuit, both outputs from the dual tap are taken out on the same side of the device, with half the total taps available each side. This maintains a bond-spacing equal to the tap period (310 μm) and avoids unnecessary crossovers by accommodating switching circuits down each side of the device with a 620 μm pitch. The switching circuit, designed to interface with this new t.d.l. requires only two diodes and two resistors per tap (Fig. 7).

The similarities of this new switching circuit and the diode convolver circuitry (Fig. 4) have resulted in the concept of fabricating a compatible microelectronic network for both devices. For a programmable a.m.f. with 10 MHz bandwidth the tap spacing is 310 μm, so that the 2-diode 2-resistor circuit is repeated at 620 μm intervals on both sides of the device. The simpler convolver has one diode and one resistor per tap on a 310 μm pitch, with components on one side only. Circuit fabrication is thus identical except for different device bonding requirements. The following considerations apply for the most part to both devices, though for clarity they refer specifically to the programmable a.m.f.

3.2 Trade-off Between Available Technologies

Two features require careful attention in the design of a programmable a.m.f. First the switching matrix must have a narrow pitch between circuits to directly interface with the tap period without any fan out. Secondly the taps, which possess capacitances of $\frac{1}{4}$ pF, must not be adversely loaded by the switching circuit if insertion loss is to be minimized. The chosen technology must have diodes which exhibit a low forward-biased impedance coupled with high reverse biased capacitive reactance to permit good switching ratios (>20 dB) and thus high fidelity operation at v.h.f.

When considering the fabrication of a high technology device it is necessary to trade off the various technologies available as shown in Table 1. This Table specifically refers to the switching matrix and does not include the code storage and selection hardware. The switching matrix which requires only diodes and resistors can be realized in a high speed, low junction capacitance (0·25 pF) semiconductor bipolar process, or even more efficiently in silicon-on-sapphire (s.o.s.) technology which utilizes vertical junction diodes to obtain 0·02 pF junction capacitance with excellent isolation between diodes and ground. The latter approach has been adopted by Hagon[9] with a s.o.s. switching matrix, m.o.s.f.e.t. shift register and holding store all fabricated on a sapphire substrate. Silicon-on-sapphire, which offers the best specification for an i.c. approach, is a new technology and unfortunately is not currently available as a production process in this country. A development process was available but the costs incurred in tailoring this process to suit our requirements would have been prohibitive. Both silicon i.c. and s.o.s. technologies can easily achieve the required narrow pitch between switching circuits irrespective of the type of circuit selected. Although either approach could be used for a 127-tap, 10 MHz chip rate device, encapsulated i.c. devices are excluded by the fan-out problem. Semiconductor chips are therefore used, each chip limited to typically 16 switching circuits to obtain high yields. This construc-

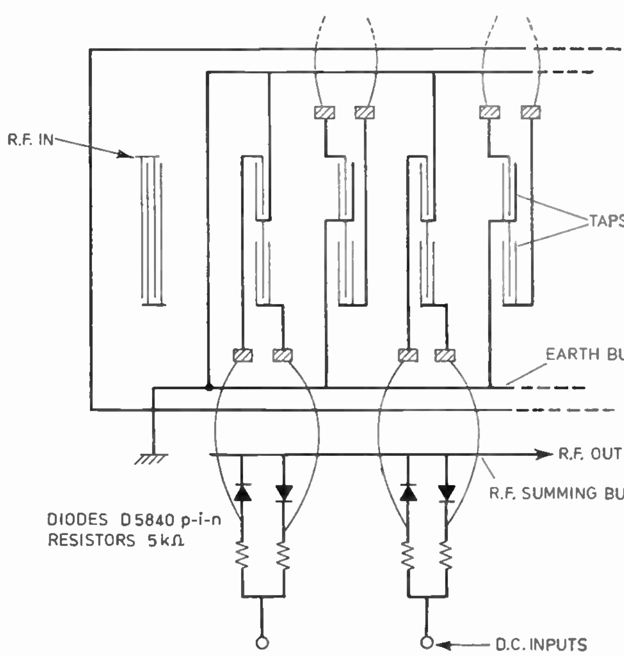


Fig. 7. Programmable a.m.f. incorporating a dual-tap structure.

Table 1. Comparison of switching-matrix technologies for two 127-tap programmable a.m.f.s

	Silicon-on-sapphire		Silicon i.c.		Hybrid	
Diodes	0·02 pF, 50 Ω at 1 mA		0·25 pF, 100 Ω at 1 mA		0·05 pF (p-i-n), 100 Ω at 0·5 mA	
Resistors	diffused		diffused		thin film	
Fabrication	few companies have technology available		industrial production process		simplest, and in-house	
Cost analysis		£		£		£
	masks	750	masks	1000	masks in-house	50
	process set-up	2500	process 1 batch	560	process in-house	50
	process 1 batch	900			diode purchase	1560
		£4150		£1560		£1660
Time scale (weeks)	mask design	16	mask design	16	mask design and fabrication	3
	mask fabrication	12	mask fabrication	12	thin film processing	6
	process set-up and production 1st batch	20	process 1st batch	12	device construction	3
	evaluation	4	device construction	3		
	device construction	3				
		55		47		12

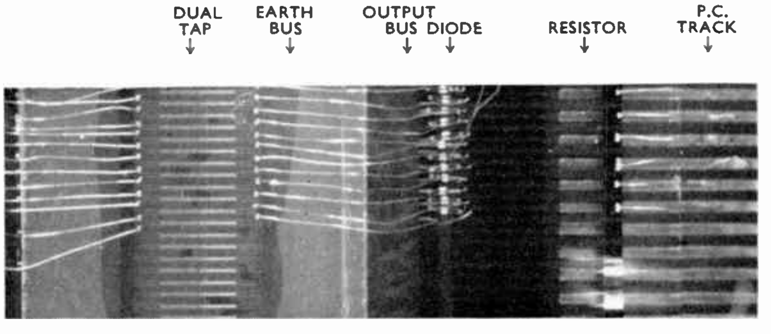


Fig. 8. Details of one side of the programmable a.m.f., highlighting tap structure, hybrid substrate and printed circuit tracks with interconnecting bonds.

tion still involves many wire bonds for chip-to-tap and chip-to-chip interconnexions, reducing to some extent the advantages of integrated construction.

A hybrid arrangement, with thin film resistors and beam lead diodes, offers a possible alternative to the silicon i.c. or s.o.s. approach. Beam lead p-i-n diodes can be obtained in a 0·17 mm × 0·9 mm package with 0·05 pF capacitance, and when combined with small area thin film resistors the 2-diode 2-resistor circuit meets the 620 μm period restriction of the latest a.m.f. design. Thick film resistors were considered as a possible alternative but the minimum reproducible linewidths of 600 μm excluded their use.

Although initial design studies were based on an i.c. approach including a tentative design, it was decided to use a thin film hybrid system in which all the fabrication could be performed in-house under direct control. As our requirements would be met by a small quantity of devices, the high component cost of the p-i-n diodes (£3 each) was offset by the relatively large mask making costs incurred in an i.c. approach. Thus for our requirements, the hybrid approach offers low capacitance diodes, acceptable cost, direct control and a much shorter time scale—9 weeks for a 31-tap device. In another context different considerations could conceivably make an i.c. approach more attractive.

3.3 Device Construction

The programmable a.m.f. consists of several separately constructed subassemblies. The hybrid microelectric circuitry was made on a 63 × 25 mm Corning 7059 glass substrate, allowing space for the SAW delay line. The latter was made on a 63 × 8 × 2·5 mm polished bar of ST-cut quartz, subsequently cemented onto the glass substrate. This assembly was then cemented on a printed circuit board carrying fan-out leads and the TTL packages which form the shift register for code selection. The r.f. parts of the device, delay line and hybrid switching matrix, were enclosed in an aluminium box which also screens the input transducer from the taps. A detail of the taps and switching circuits is shown in Fig. 8, while

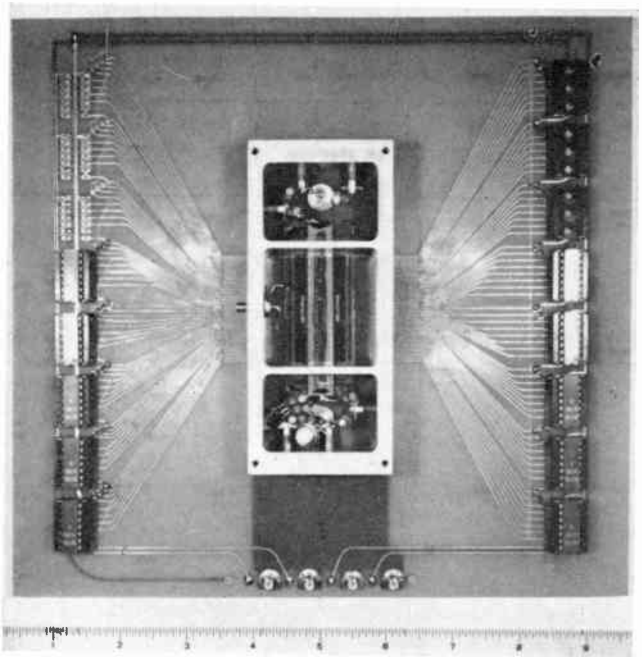


Fig. 9. Complete programmable a.m.f. with tunable matching networks and shift register code store.

Fig. 9 shows the whole device. The device had 127 taps and switching circuits, though only 31 were connected on the device illustrated.

The thin film resistors were fabricated using a nichrome-gold process. The 200 Ω/sq sheet resistance nichrome was evaporated onto the glass substrate then overlayed with gold. Both evaporations were performed with an electron-beam heated source. The resistive and conducting patterns were defined using a Shipley positive photoresist system combined with selective etches. The 5 kΩ resistors had a pitch of 620 μm per pair, with 60 × 1500 μm active areas and 200 μm wide bonding pads to accommodate the diodes. The resistor values varied by ±5% and improved photomasking combined with r.f. sputtering of the film is under consideration to improve film quality and composition. The Alpha D5840 beam lead p-i-n diodes, which were chosen for their very low capacitance (0·05 pF) and physical size, were bonded onto the gold overlay, bridging the gap between the resistor array and output summing bus as detailed in Fig. 7 and illustrated in the detail of Fig. 8. The bonding was performed with a pulse tip thermocompression bonder using a 'blanked-off' tool.

The SAW device was constructed by vacuum coating the polished cut quartz substrate with aluminium and then defining the interdigitated input transducer and taps using a positive resist system. The electrodes on each tap have a length of 1 mm and a linewidth of 6 μm with 1:1 mark/space ratio. Care was necessary to reproduce these long narrow lines over the large area of the device (40 × 3 mm) without any open or short circuits. Standard semiconductor photomasking techniques were used with special emphasis on tap placement. It is vital to the device performance that the placement of each tap should have an accuracy of at least ±0·5 μm without any accumulative error along the 127-tap array. Less accurate placement would cause a phase shift in the sampled signal degrading the coherence of signal summation over the entire tap array. Chrome-on-glass photomasks were used to define the SAW structure and the accuracy necessary is close to state-of-the-art for these masks.

The quartz substrate with SAW structure was positioned between the resistor diode networks and secured with an adhesive varnish. Connexion between the resistors and taps was made with gold wire, bonded between the gold and aluminium pads. A pulsed, thermocompression ball bonding technique was used as this does not require any steady-state heating of the device, but only an instantaneously pulsed heating of the bonding capillary when the preset bonding pressure is reached. This avoids any thermal degradation of the components and also allows wide freedom of choice for adhesives.

Interfacing the closely spaced thin film switches to standard encapsulated 8-stage TTL shift registers was accomplished by using a printed circuit board with 310 μm wide metal tracks having the same period as the resistors pairs (620 μm), which represents the state-of-the-art in printed circuit design. The tracks subsequently fan out to the shift register stages accommodated along the sides of the 9 in × 9 in (23 × 23 cm) board (Fig. 9). The pulsed thermocompression ball bonding system was again selected to complete the connexions between the thin film and printed circuit board. The copper tracks were gold plated in the bonding region to ensure good bond reliability. Soldering methods were discarded due to the close contact period and the possible problem of gold leaching from the thin film contacts. The remainder of the board was flow soldered both for protection and to facilitate connexion of the TTL packages. A ground plane on the printed circuit board formed the rear face of the screening box, and a web in the box screened the input transducer from the taps.

A similar construction is being used for the diode convolver, except that the printed circuit board will not be necessary. In this device it has been found that small variations of tap sensitivity, due to variations in the diode impedances and the taps themselves, can be corrected by trimming the resistors. For the thin film hybrid device the resolution (4 μm) of laser machining should yield a resistor tolerance of better than 0·5%.

4 Hybrid Programmable A.M.F.

4.1 Device Performance

To evaluate the feasibility of design, an initial sample device was constructed with 31 taps selected from the centre of the tap array. Programmable a.m.f. performance is summarized in Table 2, and compared with a fixed coded a.m.f. wired to the same 31-chip pseudo-noise code. The most significant difference is an increase in insertion loss to an individual tap from 66 to 74 dB, directly attributible to capacitive loading by the switching circuit. This accounts for the differences in signal/noise ratio and dynamic range between devices. The high fixed coded insertion loss of 66 dB over earlier devices (50 dB, see Section 2.2) can be directly attributed to the effects of the broadband matching network and

Table 2. Comparative performance of 31-chip fixed coded and hybrid programmable a.m.f.s

	Fixed coded	Infinitely programmable
Centre frequency (MHz)	120	120
Code chip (bit) rate (MHz)	10	10
Insertion loss to a single tap (expansion) (dB)	66	74
Signal/noise (expansion) (dB)	>30	>20
Insertion loss to the correlation peak (dB) (pulse compression)	36	44
Processing gain (dB)	15	15
Dynamic range (dB)	80	70
Loop insertion loss (dB)	102	118

dual tap geometry. The loop insertion loss quoted in Table 2 is the loss for two devices in series, one performing expansion and the other compression. The switching circuit was driven from TTL registers (see Fig. 9) with 3·5 V output swing resulting in 0·33 mW power consumption per tap for the switching matrix.

This compares very favourably with the 1 mW achieved by using an SOS switching array.[9] Trace (a) of Fig. 10 shows the impulse response for the fixed coded device. Typical tap-to-tap variation is ± 2 dB. Trace (b) shows the autocorrelation of the upper waveform, obtained by using the programmable a.m.f. Here the programmable device impulse response has been coded with a time reversed version of the 31-chip code used in the fixed device.

The ratio of the amplitude of the correlation peak to the largest of the time side-lobes is 11 dB (theoretical 15·8 dB). The discrepancy is due to internal reflexions within the programmable a.m.f. mainly caused by the loading of the reversed biased diode on the off tap. This effect, which causes the time side-lobes to loose their theoretical symmetry, can be overcome by a simple modification of the switching circuit and improving the SAW tap geometry.

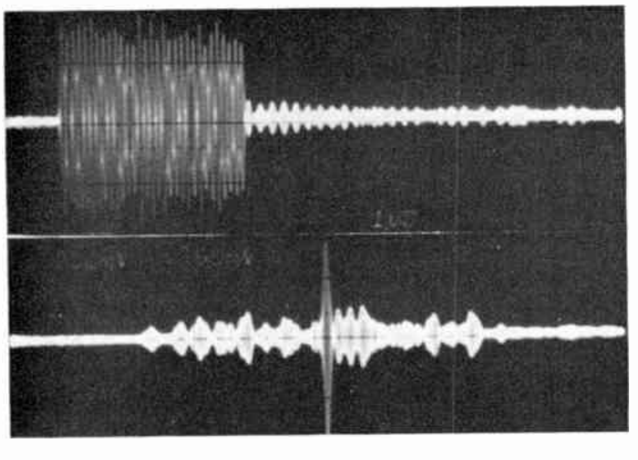


Fig. 10. SAW a.m.f. loop performance.

(a) Impulse response of a fixed coded a.m.f. passively generating the 31-chip maximum length p.s.k. sequence 1111100011011010100 00100101100 at 120 MHz. (b) Autocorrelation function displayed in a programmable a.m.f. coded with the same sequence receiving the 120 MHz burst signal shown in trace (a).

(a)

(b)

(c)

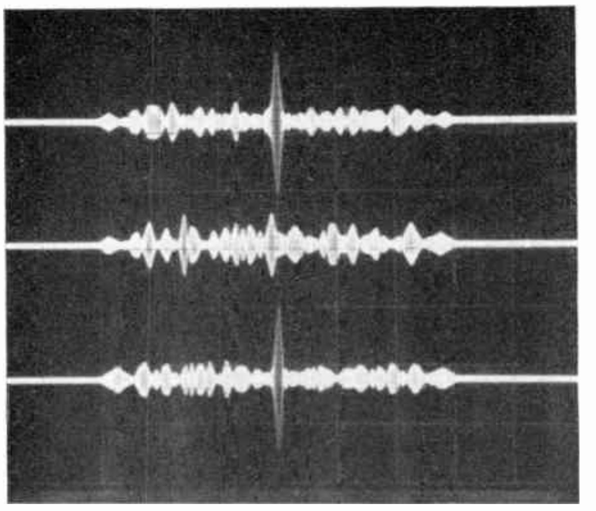


Fig. 11. Two-signature coding for bandspread communications. Traces (a) and (c) show the autocorrelation performance of the programmable a.m.f. to two 31-chip bi-phase coded signatures chosen as disjoint subsequences of a 127-chip maximum length sequence. These particular signatures were computer selected to exhibit low cross-correlation time sidelobes as shown in trace (b).

4.2 Applications

In certain multiple subscriber communication and ranging systems such as those proposed for collision avoidance surveillance,[11] user access is permitted in a truly random or uncoordinated manner. Interference between different users is avoided by allocation of a distinctive coded signature to each user. Figure 11 illustrates the performance of the programmable a.m.f. when it is coded to recognize a signature A, under conditions typical of these systems. Trace (a) shows the autocorrelation performance when receiving a signature A and trace (b) the crosscorrelation response of the same device to another signature B. The use of an envelope and threshold level detector enables a receiver to distinguish between the two signatures and decode only those signals coded with signature A. Trace (c) shows the response of the programmable a.m.f. to signature B after reprogramming the device with Code B.

Many secure communication systems[12] bandspread the baseband data by employing further modulation with a wideband binary code of very long duration prior to signal transmission. The achievement of efficient decoding in the receiver requires an identical code generator, which must be in precise time synchronization with that in the transmitter, to recover the baseband data. Synchronization of these code generators can be achieved by incorporating an infinitely programmable a.m.f. in the receiver, programmed to correlate a subsequence which will shortly occur in the transmitted code. The correlation peak provides both code recognition and accurate timing information.

Table 3. Comparison of potential performance for SAW and microelectronic programmable matched filters

	Programmable a.m.f.	Diode convolver	Charge coupled a.m.f.	Microelectronic d.m.f.
Time—bandwidth product	fair	fair	excellent	excellent
Bandwidth	medium	medium	very high	high
Dynamic range	very good	very good	good	poor
Variable chip rate	no	yes	yes	yes
Temperature dependence	medium	low	low	low
Device fabrication complexity	medium	medium	medium	complex
Reliability	modest	high	low	low
Peripheral equipment	modest	complex	modest	modest
Total power consumption	medium	medium	low	very high
Cost	high	medium	very low	high

When many subscribers are simultaneously using these systems a high co-channel interference results which renders these short code correlations (typically 127 chip) undetectable. The correlation of longer sequences with larger processing gain is often difficult due to the design, fabrication and reliability problems of a long matched filter. A programmable a.m.f., which is reprogrammable within the chip time, can however overcome this problem by correlating a code (> 1 000 chips) which is much longer than the device (127 tap) when coherent summation of the output in a recirculating delay line is employed.

4.3 Programmable Matched Filter Comparison

Several other technologies can be used to realize matched filters. The two main competitors to SAW devices are the simple digital matched filter (d.m.f.) constructed with bipolar shift registers,[13] and the truly analogue shift register realized in the new charge coupled device (c.c.d.) technology.[14]

Table 3 compares the potential performance of these devices, with the comparisons based on the 127-tap 10 MHz chip rate specification previously discussed. Both the charge coupled a.m.f. and the d.m.f. employ clocked shift register stages to achieve delay, easily permitting serial extension to handle signals with very large time bandwidth product ($>10^4$). The high clock rates of these devices also in principle permit accommodation of wideband signals (> 100 MHz), which can have a variable chip rate. The inability to achieve variable chip rate in a programmable a.m.f. limits the overall degree of programmability available. The dynamic range of a simple d.m.f. is limited, but with 8-level quantization a 50 dB dynamic range can be obtained at the expense of high device complexity and associated loss of yield and reliability. Dynamic range of the charge coupled a.m.f. is dependent on the overall charge transfer efficiency of the device. The graceful degradation characteristics of a SAW a.m.f. are not exhibited by serial devices such as the charge coupled a.m.f. and d.m.f. which have corresponding lower reliability. Active reference generation in the convolver increases the peripheral complexity compared to the other devices which only require a simple r.o.m. code store. Charge coupled device technology requires a very low power consumption when compared to high-speed bipolar logic. Cost comparisons based on the 50 dB dynamic range device tend to highlight the complexities involved in the different technologies further accentuating the necessity to evaluate accurately the hybrid device performance against other competitive technologies.

5 Conclusions

This paper has described the application of hybrid microelectronic techniques to extend the capabilities of efficient programmable SAW matched filters which will further enhance the attractiveness of SAW devices for communication systems. The particular requirement of a low-capacitance, high-fidelity switching matrix has been achieved in a size compatible with the SAW device. A detailed comparison shows that the relative attractiveness of different microelectronic technologies depends in general on the particular requirements, while for our purposes the hybrid approach was clearly the optimum choice.

A future programme is planned to extend both a.m.f. and convolver to 127 taps with a 10 MHz bandwidth and compare both fixed and programmable SAW devices under identical performance conditions. It is also considered important to design and construct a charge coupled a.m.f. to an identical specification to evaluate the performance of a competing technology. Trading off insertion loss, power consumption (device and peripherals) and achievable dynamic range in these prototype devices against precise system requirements will serve to focus future work onto the area which possesses the greatest applications potential.

6 Acknowledgments

The authors wish to acknowledge Professor P. L. Kirby, D. Willcox, B. J. Darby and J. G. Sutherland for technical contributions, I. McGee for thin film production and to the Science Research Council and DCVD (Ministry of Defence Procurement Executive) who sponsored the authors' studies.

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Manuscript first received by the Institution on 14th May 1973 and in final form on 5th March 1974. (Paper No. 1592/CC 204)

Conference Report

The Electronics Industry and Higher Education

Royal Holloway College, Egham, Surrey

28th to 31st March, 1974

Organized by the IERE with the association of the IEE, this conference aimed to bridge the gap or interface between industry and higher education. It was held at the week-end and the programme allowed for considerable time for informal discussion between the two sides. It was perhaps unfortunate that the timing of the conference was only shortly after the winter fuel crisis and the end of the 3-day working week in industry: hence, attendance, particularly from industry, was lower than might have been expected. What the conference attendance lacked in quantity, however, it made up in quality!

This review of the conference is based on reports prepared by the chairmen of the five sessions, all of whom were members of the joint organizing committee and include Professor G. D. Sims (Chairman), Mr. J. R. Thompson, Mr. C. H. G. Jones, Mr. P. G. Tanner and Professor D. S. Campbell.

The full list of titles of the papers presented at the conference was published in the February 1974 issue of The Radio and Electronic Engineer (page 104) but it should be noted that, in order to maintain the informal nature of the conference, the papers were not pre-printed and only a 28-page booklet of digests of the 31 contributions is available from the IERE Publications Department, price 50p.

The Present and Future Output of Higher Education

Dr. H. W. French (Department of Education and Science) opened the conference with a stimulating paper entitled 'Government Policy for Higher Education'. After reviewing the three main policy options open to Government he considered the 1974 situation. The financial situation is grim, drastic cuts have been made in expenditure on Higher and Further Education Projects and the only ray of sunlight in the cloud is that the figure of 750,000 Higher Education students by 1981 is likely to be a considerable over-estimate. Higher Education will need to be more economically organized and will require changes in course patterns and close liaison with Industry and Business.

Professor G. D. Sims (University of Southampton) discussed employment and qualifications in the Electronics Industry and indicated that the number of students entering the colleges was not sufficient to meet the future requirements of the country.

Both Mr. F. R. V. Langridge (Engineering Employers Federation) and Mr. J. C. Todd (EMI Electronics) in their contributions indicated that the colleges were failing to provide the right course content to satisfy industry. Academics present hotly denied this and suggested that industry when pressed were unclear on their actual requirements.

The paper by Dr. J. R. Tillman (PO Research) an 'Undergraduate Attitude to Types of Employment' outlined some of the problems of recruiting suitable men and of giving them job satisfaction during their early years in the industry.

The Conference then went on to consider types of degree courses applicable to the electronics industry including single subject courses, inter-disciplinary and multi-disciplinary courses, sandwich courses and modular degree schemes. Professor W. D. Ryan (Queen's University of Belfast) reminded the delegates that Universities were originally vocational training institutions, albeit for the church, law and medicine, and that it was not therefore unreasonable to consider this aspect today. He further pointed out that the university training cycle of from three to seven years, concerned in many cases with single discipline teaching, seemed to be able to meet the needs of an industry which can rarely look ahead more than a year. Dr. P. E. Stringer (Lanchester Polytechnic), in discussing the establishment of common ground in multi-disciplinary courses, emphasized that the typical electronics syllabus prepares a student for about seven years of practice before he needs to be retrained or further educated in order to make a further contribution. He reminded the delegates of the price industry was now paying for cutting back on graduate recruitment because of past economic fluctuations and urged industry to be more honest with students in presenting employment prospects.

Dr. N. G. Meadows (Glasgow College of Technology) brought out the point that courses were still very much Ph.D.-oriented despite a dramatic swing toward a need for graduates in the manufacturing and service functions of industry, and in developing the theme of inter-and multi-disciplinary courses postulated a clear need for more engineers with a broader academic background.

In speaking about sandwich courses, Dr. J. M. Ivison (Loughborough University) dwelt on the present advantage of the sustained contact with the industrial environment experienced by the student, and the subsequent clearer match of the needs of the individual to the needs of the employer. Mr. M. Brandon-Bravo (City of London Polytechnic) concluded the session by discussing modular degree schemes during which he commented on the lack of manpower planning at national level, the ill-defined and imprecise needs of the industry and the need for the student of today to be able to select his own degree, institution, and course so as to match his own developing sense of vocational direction and career aspirations.

Crossing the Interface and Industrial Training

Unfortunately Dr. F. W. Harrison (Philips Industries) was prevented by indisposition from making his contribution on the change for the graduate from the educational to the industrial environment. However, Mr. J. M. Sharman gave a most interesting and thorough presentation on the practical approach made by the University of Cambridge to the problem of running suitable training courses to 'match' graduate engineers to the needs of industry particularly in the fields of production methods and management.

Mr. A. Aspinall (Engineering Industry Training Board) guided the delegates' thinking to training aspects of the theme of the conference by summarizing the EITB's current grant/levy policies, and reviewing the training of professional engineers and the services which the Board is anxious to supply in order to assist industry to raise its standards.

Post Graduate Activities

Five papers were delivered in the session on Post Graduate Activities. Professor E. D. R. Shearman (University of Birmingham) described the lessons learned in providing a variety of one-year M.Sc. and short post-experience courses in electronic and electrical engineering at Birmingham University. These had been well attended but he made a point which was echoed throughout the conference that it would be more appropriate to have fewer SRC studentships but to pay them a more respectable sum. Professor H. Sutcliffe (University of Salford) analysed the number of short courses run each year by Universities and Polytechnics and the number of these which are in electrical subjects. Considering the high cost of these together with the problems encountered when busy engineers leave their work for any length of time, he questioned whether the effort of the experts involved in any one subject might not be better expended in preparing monographs for much wider circulation. In considering post graduate courses with industrially-based projects, Professor W. F. Lovering (University of Surrey) referred to the difficulty of finding projects of significance for students to undertake. He questioned whether the usual entry qualification of a good honours degree was either necessary or sufficient for post graduate courses and from his statistics he showed that professional qualifications coupled with industrial experience were often a more successful combination. Professor C. W. Turner (King's College, University of London) doubted whether the present pattern of post graduate courses in electronics satisfied the requirements of industry and said this was reflected in the fact that there was little financial advantage in an M.Sc. The M.Phil. he suggested should be scrapped. In general he considered a more highly integrated pattern of post graduate courses is needed in which private industry plays a more active role by providing man power and resources. Professor W. E. J. Farvis (University of Edinburgh) dealt with the SRC policy and post graduate degree courses, indicating the characteristics of courses and the qualities on which they are assessed. In doing this he had extracted some statistics not previously available which it is hoped will be given a wider circulation elsewhere.

The session was followed by a useful discussion in which points were raised dealing with the difficulty of obtaining useful industrial speakers on post graduate courses, the problem of obtaining recognition for staff involved on such courses and the limited financial provisions often made for these courses. The general feeling, however, endorsed by industrial speakers present, was that suitable courses were of definite value particularly where these could be used for re-training purposes.

Research

Professor E. A. Ash (University College London) opened the session with a critical assessment of the Parliamentary Expenditure Committee's third report. Given the composition of the committee which contained little experience of graduate work and none of post graduate activities, it was hardly surprising that their conclusions should have been so wrong. Ph.D. work was essential to the life and work of a university. However, there were actions that could help considerably to improve the image of a Ph.D. Firstly, SRC should provide a realistic grant; half the numbers at twice the value would be a considerable improvement. Secondly, much greater attention should be paid to the originality of the Ph.D. work and finally some way of grading Ph.D. awards would be useful. One thing was certain, that with the present cost of acquiring a Ph.D. (£3000), no student should be asked to pay this back—it would be saddling him or her with an intolerable debt.

Professor K. Hoselitz (Mullard Research Laboratories), talking from the point of view of running an industrial research laboratory, criticized universities for being too much on the defensive with regard to what they could offer. The product from the university research schools can easily be used in industrial research laboratories. However, he made one plea, and that was for a strong course work element in a Ph.D's training.

Research undertaken in Polytechnics was discussed by Dr. N. G. Meadows (Glasgow College of Technology, formerly Portsmouth Polytechnic) The nature of this research was of an applied type and as such, it was argued, needed special encouragement as it could fit in well with the needs of industry.

SRC policy on electronics research was outlined by Professor W. E. J. Farvis in his position as Chairman of the SRC Electrical & Systems Committee. SRC was involved last year with the total expenditure of £64M, of which 40% went to research grants in nuclear physics, 20% to astronomy and space work, 15% to science and 9% to engineering. (Student grants totalled 12% + other items 4%.) The Electrical and Systems Committee was only one of 9, answerable to the Engineering Board and as such was responsible for grants totalling around £700,000. In 1973, there had been 113 grants totalling £500,000. Support for Research Assistants (23) and Technicians (21) could be obtained. The success rate on grant applications was on average 40%, but varied with size of grant. (70% success if $<$ £4k; 29% success if $>$ £100k). In discussion, it became obvious that certain areas of SRC policy were not clearly solved, e.g. the compatibility of industrial relevance with that of funding from SRC rather than a company. However, it was obvious that SRC was in full agreement with the universities and polytechnics on the problem of the size of the student grants, but the solution was not in the hands of SRC. In response to criticism on the size of the nuclear physics vote, it was agreed that more money must be applied for by electronics departments and this will then cause the balance to alter!

Professor W. Gosling (University of Bath: formerly University College of Swansea), gave an encouraging description of a mobile radio project that he had run at Swansea, with co-operation between Government, University and several companies. Six engineers from industrial companies were seconded to the university for three years. This showed what could be done. A point that the papers did emphasize, and which came out clearly throughout the conference, was that an active research group cannot be run on the basis of Ph.D. student research workers, but only by using professional research scientists and engineers. Professor Gosling went so far as to call the use of students in this context, immoral.

Mr. P. G. Tanner (N.R.D.C.) spoke about the position on patents. It was noted that in electronics, 1 in 3 patents offered to N.R.D.C. was accepted. They were looking for £50,000 turnover each year from the patent to ensure success, but it could well take time to get there. Contrary to popular belief, a very large percentage of any royalties obtained were returned to the inventor.

The final paper in the session was given by Mr. K. R. Thrower (Racal). Arising from the discussions of the previous day, he had summarized what he required from new graduates entering his company as being:

(i) A basic understanding of engineering fundamentals (materials).

(ii) Circuit theory—including the ability to *use* mathematics.

(iii) A good idea of how to keep notes—he felt that the training provided in project work was not strict enough.

(iv) Experience of making measurements using equipment which had not been 'debugged'.

(v) An understanding of the easy occurrence of parasitics in circuit design.

Bridging the Interface

In this session it was broadly agreed that Universities and Polytechnics are now much better set up to work with industry than they were ten years ago, with effective group-research programmes and industrial units all aimed towards promoting the inflow as well as outflow of ideas across the interface. It was suggested, however, by Dr. J. D. E. Beynon (University of Southampton) that the flow of information about what went on in Universities and Polytechnics was insufficient and that the D.E.S. publication 'Scientific Research in British Universities and Colleges', though useful, could, to advantage, be supplemented by something resembling the Smithsonian Institution's 'Science Information Interchange'.

Mr. D. F. Dunster (NE London Polytechnic) again reminded the meeting of the misconceptions, common in the student's mind, about what industry was really like and demonstrated that with a real appreciation of the problem, and very little effort, much could be done to smooth the transition to industry—particularly by properly thought out and co-ordinated training programmes.

Professor P. L. Kirby (Welwyn Electric and University of Edinburgh) demonstrated the value of visiting professors by drawing attention to the liveliness that can be brought to such neglected areas as patent law, accounts and language teaching by intelligent industrialists. He entered a strong plea for more attention to be given by the educational sector to these, and other similarly neglected areas, and it was clear that the meeting sympathized with this point of view—while regretting that there were few people about like Professor Kirby to make them so exciting!

Mr. J. P. Moore (University of Surrey), Mr. A. P. Dorey (University of Southampton) and Mr. D. Stewart (University of Strathclyde) all spoke of the work of industrial liaison units, and companies, based on Universities and while they instanced many difficulties in such operations, it was clear that all felt that provided they contributed to the academic purpose, which was the primary concern of the higher education sector, such activities were well worthwhile. It was forcefully demonstrated that such activities were unlikely to do much more than break even in monetary terms and at the same time it was made clear that this kind of work was complementary to, rather than competing with, professional consulting or industrial organisations. The majority of problems handled were of a kind that industry either would not, or could not, take on.

The final contribution from Mr. A. J. Egginton, Director of the SRC Engineering Board, emphasized once more the relatively small SRC expenditure on engineering, compared with other science-related sectors. It was again argued that, at least in part, this was due to a shortage of good applications for support from the engineering sector and a number of reasons for the existence of this situation were advanced.

In conclusion, it was felt that the academic sector had progressed considerably in its efforts to bridge the interface with industry during the last ten years. It was far less obvious that the similar reciprocal effort was being made by much of industry. The continued improvement of relations and effectiveness needed sustained effort on both sides and the next few years could well be critical—particularly in view of DES attitudes to higher education, coupled with the underlying economic pressures which are currently affecting both sides of the interface.

UDC 621.374.32:621.317.32

A digital-analogue minimum detector for riometers

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and

D. ROUTLEDGE, Ph.D.*

SUMMARY

A circuit using commercial integrated circuits has been developed to replace the minimum-detector circuitry commonly used in swept-frequency passive ionospheric absorption measurements. The new circuit offers good accuracy and is completely free of decay error.

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1 Introduction

The method of measuring ionospheric radio absorption by swept-frequency passive observations of the galactic radio emission has been widely used since the time of the International Geophysical Year (IGY). At that time, Little and Leinbach† published detailed circuits for a standardized 27·6 MHz riometer. The operation of this riometer was based upon frequency-sweeping and minimum-detecting circuits which ensured that the system would record the minimum signal strength encountered as a narrow reception band swept periodically through a relatively broad search band. The method thereby minimized the effects of terrestrial interference on the absorption measurement.

We are building an instrument which is similar in principle to the IGY riometer, but which is to operate near 12 MHz as part of a program of flux-density measurements of discrete cosmic sources. The purpose of this contribution is to present an improvement in circuitry which will be incorporated in the 12 MHz riometer; such improvements are possible now with the use of complex but reliable integrated circuits.

2 The New Circuit

A digitally-generated ramp voltage tunes the swept oscillator in the new 12 MHz riometer. The ramp generator includes a ripple counter, a clock oscillator, and an integrated-circuit digital-to-analogue converter (d.a.c.). Figure 1 shows the pertinent section of the 12 MHz riometer.

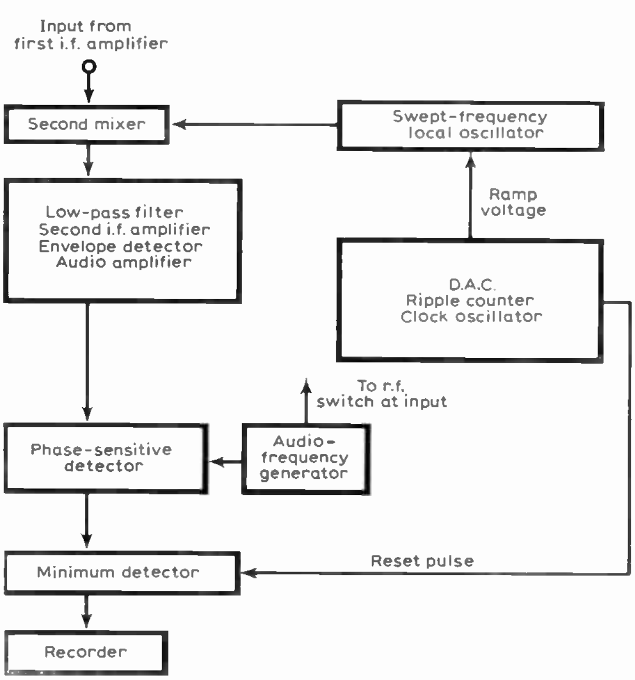

Fig. 1. Position of minimum detector in swept-frequency receiving system.

† Little, C. G., and Leinbach, H., 'The riometer—a device for the continuous measurement of ionospheric absorption', *Proc. IRE*, 47, No. 2, pp. 315–320, February 1959.

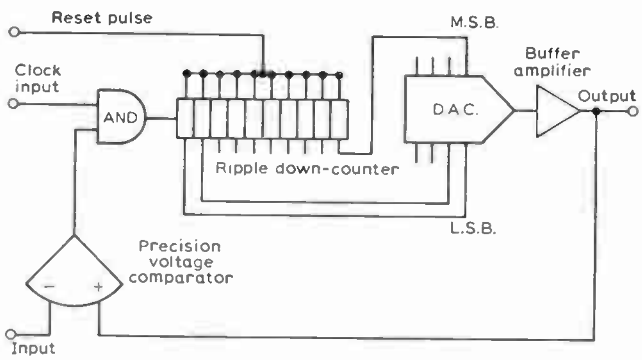

Fig. 2. Minimum-detector circuit.

In Fig. 1, the ripple counter fills and resets to zero with a period determined by the clock rate, and the d.a.c. output rises and falls periodically in response. It occurred to the authors that an excellent minimum detector could be built using similar circuitry if the count in another ripple counter could be stopped whenever the input to the minimum detector ceased to fall, and if the count could be resumed whenever the input fell once more. The output from such a minimum detector would be completely free from decay error, regardless of the duration of the frequency sweep, since the count in the ripple counter could be maintained indefinitely without error. Significant quantization errors could be avoided by using a d.a.c. of suitable precision.

Figure 2 shows a realization of the concept. The circuit may be seen to be a rudimentary analogue-to-digital converter in which the digital output is discarded. The ripple counter used is actually a down-counter so that the count decreases when clock pulses are passed by the AND gate. Thus, a count of 11...11 becomes in turn 11...10, 11...01, etc. Clock pulses are passed by the AND gate only when the voltage applied to the inverting input of the comparator is smaller than the output from the buffer amplifier, however. Thus, the circuit executes an intermittent sample-and-hold operation in which the output decreases monotonically until a reset pulse from the ramp-voltage generator sets the ripple counter to 11...11 again.

In the present circuit, the d.a.c. is a Precision Monolithics AIM-DAC100-Q1B, the ten-bit ripple-counter is composed of three Motorola MC839 four-bit ripple counters, and the precision voltage comparator is a Precision Monolithics CMP-01. The 17-Hz clock oscillator drives the minimum detector and also the ramp-voltage generator, whose period is one minute.

3 Performance

Excellent accuracy was obtained with the circuit; a simulated minimum-detector input was tracked, and the minima were detected, with 0·1% error over an input range of zero to 9 V. No decay of minima was observed. The frequency response was more than adequate, being d.c. to approximately 10 Hz.

4 Acknowledgement

This work was supported by an operating grant from the National Research Council of Canada.

Manuscript received by the Institution on 15th January 1974. (Short contribution No. 173/CC 205)

UDC 621.372.54: 621.376.52: 681.327.8

Binary transversal filters in data modems

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SUMMARY

The design of a data modulator using binary transversal filters is described. The technique used can be applied to a number of modulation schemes and leads to data modulators which are very suitable for integration. A matched filter demodulator is also described.

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1 Introduction

The transmission of data is becoming increasingly important in communications systems, particularly where the systems are computer-based. Frequently existing telephone or mobile radio voice circuits are used which have a bandpass frequency response (0·3–3·4 kHz) which is unsuitable for the transmission of the data signal $d(t)$, shown in Fig. 1(a), which has a wide spectrum (Fig. 1(c)) extending upwards from d.c. The difficulty is overcome by using a modem, in which the data signal is modulated on an audio subcarrier, using one of the many modulation schemes available, and then demodulating again at the receiver. For example, phase shift keying (p.s.k.) modulation, shown in Fig. 1(b), has a frequency spectrum restricted to twice the bit rate and has no very low frequency components (Fig. 1(d)).

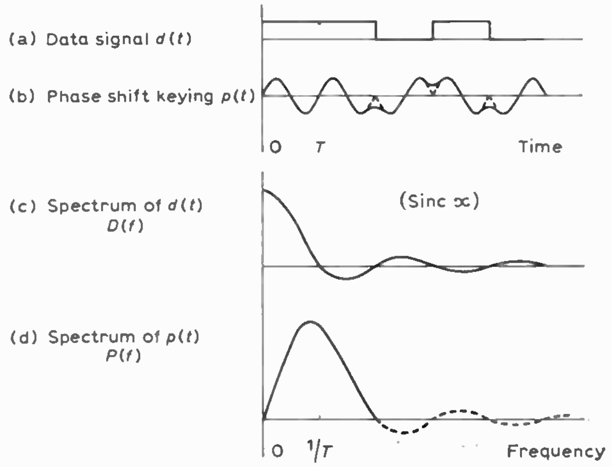


Fig. 1. Waveforms for modems.

Traditionally data modems have used analogue modulators and filters which are difficult to implement in integrated form and which are tied to particular bit rates. Recently a digital approach[1] has been described in which the modulator was replaced by a digital logic gate and a binary transversal filter used to smooth the digital waveform into the required continuous waveform needed for transmission. The binary transversal filter can also incorporate any spectrum shaping and equalization that is needed. However in many cases, particularly in mobile radio, this filtering is not used, in which case the modulated data signal can be generated directly giving an economical and easily designed solution.

2 The Data Modulator

Data modems operate at bit rates of 1·2 kb/s or more and consequently the frequency of the audio carrier is comparable and often equal to the bit rate. Because the modulated data signal has a waveform with only one or two cycles of carrier in a bit period, rather than an envelope with very many cycles, it is possible to use sampled circuits with only a modest sampling frequency and to use more direct methods of generating the modulated signal.

The new data modulator consists primarily of two filters which are stimulated by impulses under the control of the data signal, as shown in Fig. 2(a). A logical '1' steers an impulse to the first filter, which is chosen to have an impulse response $I_1(t)$ equal to the waveform which is transmitted to represent a logical '1'. A logical '0' input results in the second filter being stimulated to produce an output $I_0(t)$ equal to the logical '0' signalling waveform. The filter outputs, shown in Figs. 2(b) and 2(c) are combined in a summing amplifier to produce the modulated data signal. Any signalling scheme, even one using arbitrary signalling waveforms, can be implemented by realizing the required filter impulse responses.

It would be difficult to realize a specified impulse response restricted to a short time interval, using traditional linear filters. However the binary transversal filters (b.t.f.) which will be described here can be designed to have any impulse response required and in addition the response is automatically restricted to a given time interval, which for our purposes is the bit period.

The approach will be illustrated by a modulator designed to generate p.s.k., in which a logical '1' is represented by one cycle of carrier and a logical '0' is represented by one cycle of inverted carrier. Consequently the filter impulse responses are:

$$\left.\begin{aligned} I_1(t) &= \sin \omega_c t \\ I_0(t) &= -\sin \omega_c t \end{aligned}\right\} nT < t \leqslant (n+1)T \tag{1}$$

where ω_c is the carrier angular frequency and T is the bit period. The way in which the data signal controls the source of impulses $\sum\limits_n \delta(t-nT)$ and the way the filter outputs are combined to produce the modulated data signal is shown in Fig. 2(d)–(k).

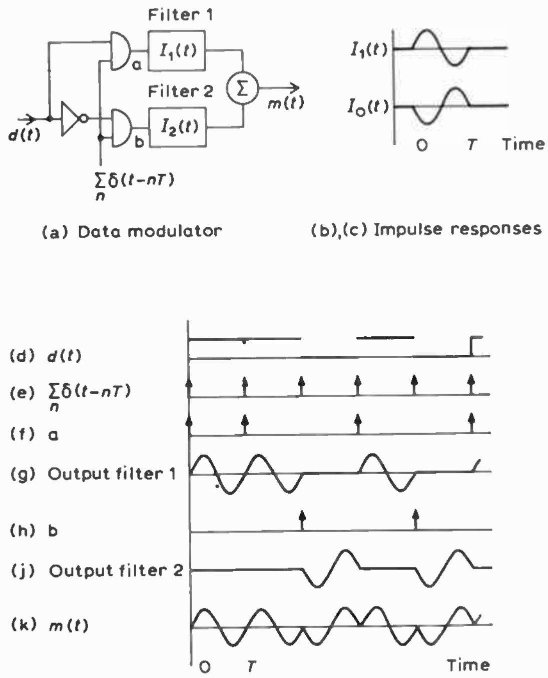


Fig. 2. Data modulator and waveforms.

3 The Binary Transversal Filter

A binary transversal filter is a digital shift register in which an output is taken from each of the n stages through a weighting resistor to a summing amplifier, as shown in Fig. 3(a). The filter accepts a binary input which is clocked through the register at a higher rate than the bit rate f_b, say at $f_k = pf_b$ (where p is an integer), and appears at each successive stage in turn. In the data modulator application the filter input should be an impulse. However, the narrowest pulse that can occur in the shift register has a duration T/p so that the input is not an impulse but a pulse $a(t)$ where

$$a(t) = \begin{vmatrix} V & 0 < t \leqslant T/p \\ 0 & 0 \geqslant t > T/p \end{vmatrix} \tag{2}$$

As this pulse is shifted through the register it occupies each stage in turn and produces an output from the summing amplifier which is determined by the size of the resistor connected to that stage. The output $b(t)$ during the jth clock period is determined by the jth resistor, so that

$$b(t) = a(t-jT/p)\,\frac{R_f}{R_j} \tag{3}$$

for $jT < t \leqslant (j+1)T$, where R_f/R_j is the gain of the amplifier.

If the correct resistor values are chosen then the pulse, as it shifts down the register, will generate the required waveform in the form of a staircase approximation; Fig. 3(b) illustrates this for a sinewave output. Note that the filter response is restricted to the time interval needed for the pulse to pass through the register. The complete filter output is found by summing the output from each stage.

$$b(t) = \sum_{j=1}^{j=n} a(t-jT/p)\,\frac{R_f}{R_j} \tag{4}$$

$$= a(t) * \sum_{j=1}^{j=n} \frac{R_f}{R_j}\,\delta(t-jT/p) \tag{5}$$

$$= a(t) * \bar{I}(t) \tag{6}$$

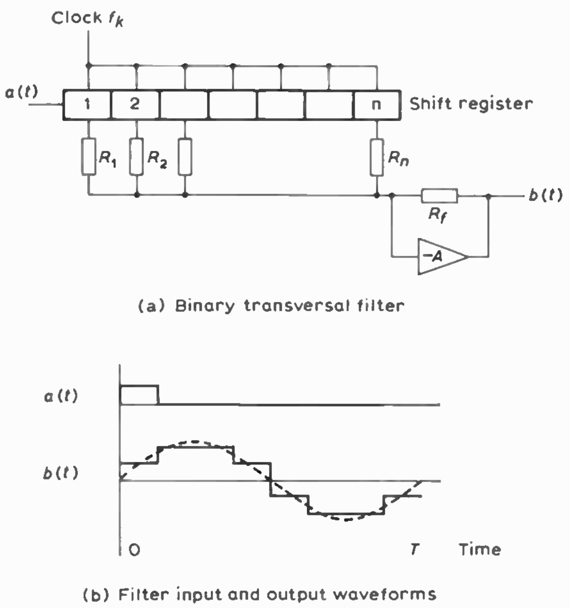


Fig. 3

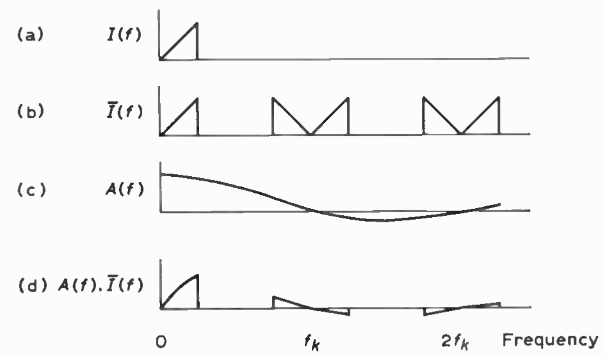


Fig. 4. Impulse response spectra.

The first term in equation (6) is the input $a(t)$ which is convolved ($*$) with the filter impulse response in sampled form. The sampling results from using a shift register clocked at p/T rather than a continuous analogue delay line. Usually $a(t)$ can be considered as an impulse at high clock rates ($p \geqslant 8$), in which case the filter is very easily designed by making the resistor ratios R_f/R_j equal to the samples of the desired waveform. However in the general case the effect of the non-negligible duration of $a(t)$ can be seen by taking Fourier transforms in equation (6):

$$B(f) = A(f).\bar{I}(f) \tag{7}$$

where $\bar{I}(f)$ is the transform of the sampled impulse response $\bar{I}(t)$, and $A(f)$ and $B(f)$ are the transforms of $a(t)$ and $b(t)$. Figure 4 illustrates the point with an impulse response spectrum $I(f)$ at 4(a) and the same spectrum in its sampled form at 4(b), which shows the spectrum repeated at harmonics of the clock frequency f_k. The spectrum of the input pulse $a(t)$ is the sin x/x function shown at 4(c), and the spectrum of the output waveform is the product of $A(f)$ and $\bar{I}(f)$, shown at 4(d). Normally the frequency components above $f_k/4$ are filtered out with a simple CR network to give a smooth waveform and the effect of the convolution with $a(t)$ has made this easier by significantly attenuating the frequency components centred at harmonics of f_k. The disadvantage is the distortion of the baseband components up to $f_k/4$, with a loss of 10% amplitude at $f_k/4$. If this is significant then a higher clock rate must be used, or the shift register outputs combined with a narrow pulse (duration say $T/32$) to give a better approximation to the sampled waveform $\bar{I}(t)$.

The filters are easily designed to generate any waveform which is restricted to the 'length' of the shift register. An interesting possibility is the pulse shape defined in reference 7 which has the narrowest possible spectrum for a given pulse width, which must be a valuable feature in data transmission.

4 Practical Modulators

Many modulation schemes result in signalling waveforms which are antipodal, in which case only one b.t.f. is needed to generate the modulated data signal. In p.s.k. for example, the logical '1' and '0' waveforms are sin $\omega_c t$ and $-\sin \omega_c t$, where one waveform is the negative of the other. In such cases the logical '1' waveform can be generated by resetting the shift register stages to '0' except for the first state which is set to 1; this 1 then shifts through the register to produce the normal waveform. The logical '0' waveform is generated by setting the register full of '1's, except the first stage which is reset to '0'; as this '0' shifts through it generates the inverted waveform.

The circuit of a p.s.k. modulator of this type is shown in Fig. 5(a); it includes a divider to generate the 'impulses' and the synchronous clock at eight times the bit rate. An extra flip flop 'D' has been included to maintain the d.c. level of the modulated data signal constant; it has its own resistor R_d connected to the summing amplifier, where $R_d = \sum(1/R_j)$. When all the shift register stages are at 1 and supplying current to the amplifier, then flip flop 'D' is at '0' and drawing the same current. Alternatively

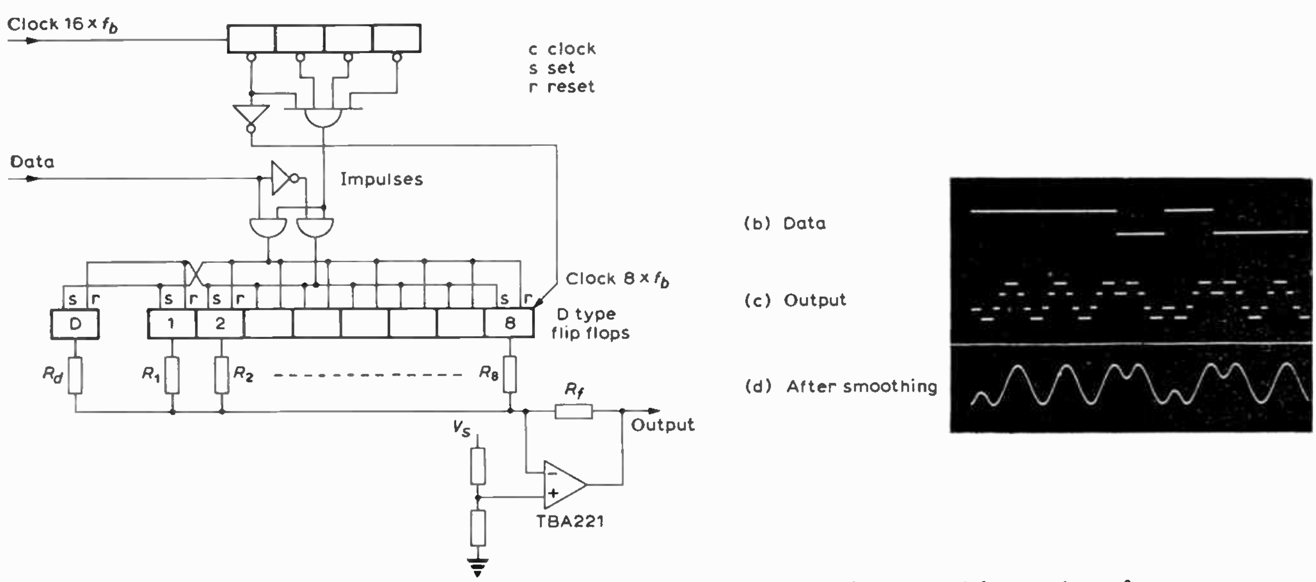


Fig. 5. P.s.k. data modulator and waveforms.

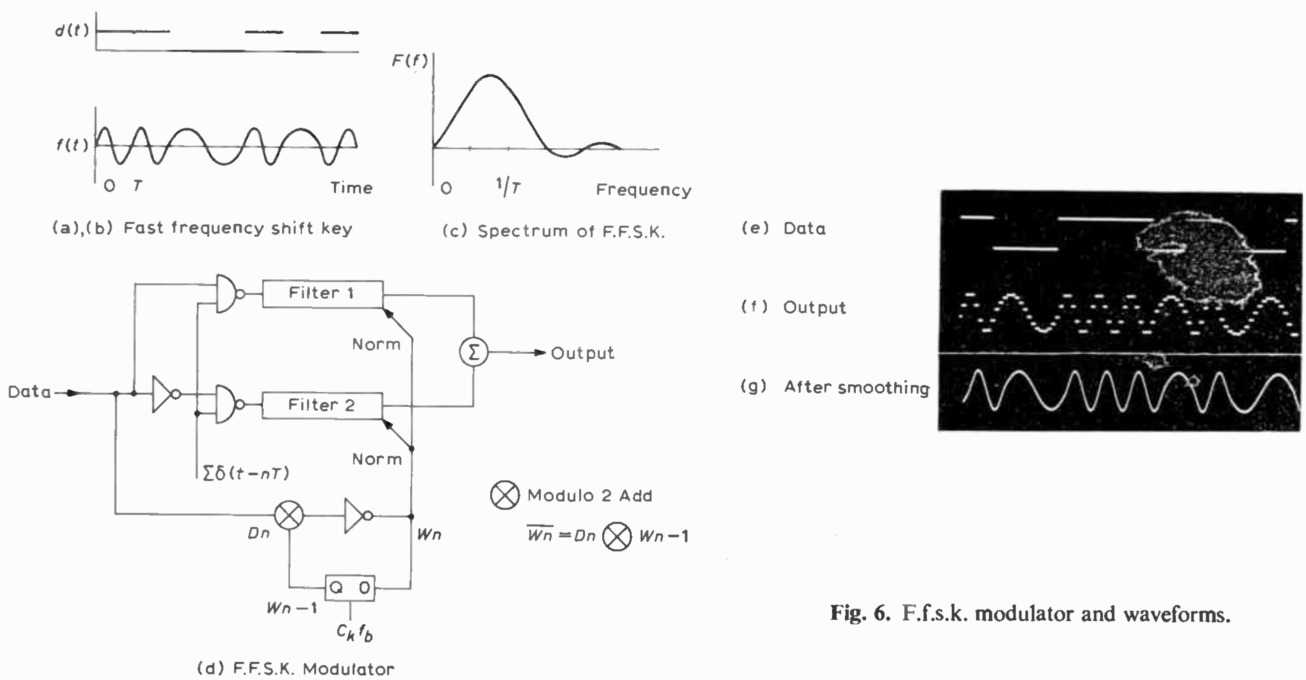


Fig. 6. F.f.s.k. modulator and waveforms.

when they are all at '0' and drawing current then D is at '1' and is supplying current. M.o.s. logic is used because it has well-defined voltage levels of zero and the supply voltage which ensures that each shift register stage contributes correctly to the output.

A modulator has also been realized which generates fast frequency shift keying[2] (f.f.s.k.), which is frequency shift keying with a modulation index of $\frac{1}{2}$. Typical waveforms are shown in Fig. 6(b) where a logical '1' is represented by a whole cycle of sinewave of bit rate frequency and a logical '0' by only half a cycle of sinewave at half the bit rate frequency. Discontinuities in the phase of the modulated data signal are avoided by using either positive or negative symbols as necessary, which means that four symbols have to be generated. However, they are in antipodal pairs so that only two b.t.f.s are required. An advantage of f.f.s.k. is the reduced bandwidth of the modulated signal spectrum,[3] which is shown in Fig. 6(c). The modulator in Fig. 6(d) includes the logic needed to decide whether a normal or inverted symbol is required, which controls the filters via their input 'norm'. The staircase type output waveform is shown in Fig. 6(f) and in Fig. 6(g) after smoothing in a CR network. The circuit can also be used to generate f.s.k. ($m = 1$) by doubling the clock frequency and allowing each symbol to be produced twice during a bit period. This results in one cycle of carrier for logical '0' and two cycles of double frequency carrier for a logical '1'.

A further saving in hardware can be made in the case of f.f.s.k. by recognizing that the four symbols to be generated are derived from a sinusoid by choosing both polarity and frequency (either $f_b/2$ or f_b). Consequently only a single filter is needed with resistor values chosen to give an impulse response of half a cycle of sinewave, and with the facility of either normal or inverted output and with a shift register clock at either eight or sixteen times the bit rate. For a logical '0' the filter is used in the usual way to produce a half cycle, either normal or inverted. For a logical '1' the clock frequency is doubled so that two half cycles are generated within the bit period, with the second half cycle arranged to be of opposite polarity to the first. Any data waveform which can be assembled from half cycles of sinewave, given a choice of polarity and frequency, can be produced by providing the necessary selection logic. The approach can be used generally by breaking the modulated waveform down into the smallest number of basic symbols which can be inverted or time compressed to generate the required waveform. There is no reason why time inversion should not be used to further reduce the filters, using bidirectional shift registers.

The above circuits were made for testing using conventional methods, but in production they could be made either as integrated or as thin film circuits. The integration technology must be able to produce resistors over a range of values of, say, 20 to one with an accuracy of 5%, to give typical figures. For smaller quantities the film circuit is attractive, particularly for making the resistors, and offers many of the advantages of the integrated circuit at a much lower initial cost.

5 An All-Digital Approach

The resistors in the b.t.f. with their analogue values can be avoided by using an all-digital approach. A longer shift register of, say, 32 bits is used, which is clocked at 32 times the bit rate. At the start of each bit period an 'impulse' parallel loads a 32-bit pattern into the register, which is then shifted out serially during the bit period. The pattern is held as hard wiring and is chosen so that it generates the required signalling waveform when it passes through a delta demodulator at the output. In the case of antipodal signalling waveforms, only one register is needed, as in Fig. 7 and the data input selects either the pattern or its complement for output to the delta demodulator.

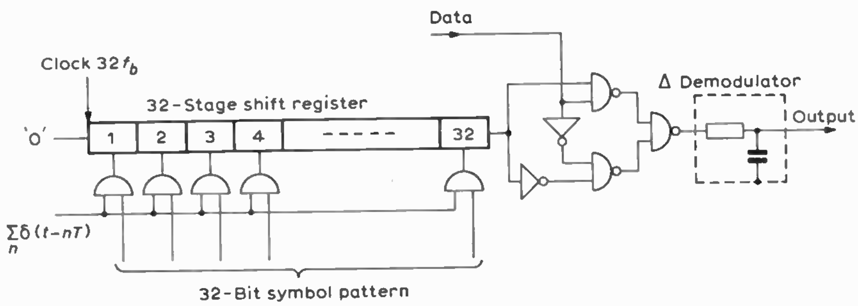


Fig. 7. All-digital data modulator.

6 Matched Filter Receivers

It is well known that a matched filteri s the optimum receiver for a signal with added stationary Gaussian noise[4], but limited use has been made of this result, perhaps due to the problem of realizing the filter. For white noise a filter is matched if its impulse response is the time reverse of the received signal waveform without noise:

$$I_m(t) = r(T_d - t) \tag{8}$$

where $r(t)$ is the received signal and
T_d is the decision time.

The time reversal is relative to the instant T_d at which the filter output is tested by a decision circuit, which is usually at the end of the bit period.

The transversal filter is very suitable for use as a matched filter because the designer is free to choose the impulse response without restriction. The filter must accept the received signal with the added noise, which is clearly an analogue signal and calls for the analogue tapped delay line shown in Fig. 8(a). The impulse response is determined by the tap weights and is in sampled form due to the spacing of the taps at intervals T/q:

$$\bar{I}_r(t) = \sum_{j=1}^{j=q} \frac{R_f}{R_j}\,\delta(t - jT/q) \tag{9}$$

The analogue delay line could be implemented with charge transfer devices[5] when they become available or with binary shift registers preceded by a delta modulator.[6,8] The delta modulator converts the analogue signal to a binary bit stream at a clock rate of qwf_b. The analogue delay elements are replaced by shift registers of w bits. The factor q (typically 8) determines the sampling rate used in the filter and the factor w (typically 4) the quantization step size in the delta-modulation process, resulting in a shift register of 32 stages. At first sight a delta demodulator is needed at each tap on the register, but because a linear delta modulator is used the demodulation can be postponed until after the summing amplifier, as in Fig. 8(b).

The matched filter is optimum in the sense that it gives the best signal/noise ratio at the times $t = nT_d$ when decisions are made. However decisions can only be made if a sampling pulse is available at the times $t = nT_d$, which means that a clock regeneration circuit is needed. Clock regenerators are normally driven by the baseband signal which results from envelope or coherent detection. The signal at the output of the matched filter has a time waveform equal to

$$u(t) = \bar{I}_r(t) * r(t) \tag{10}$$

which in general is not equal to the baseband data signal. The difference in the two waveforms is shown in Fig. 9, for the case of p.s.k.

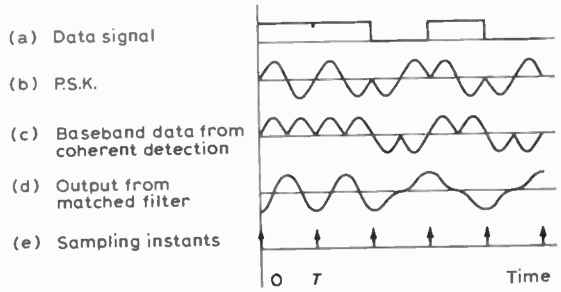
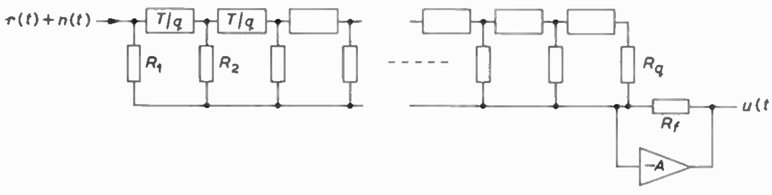
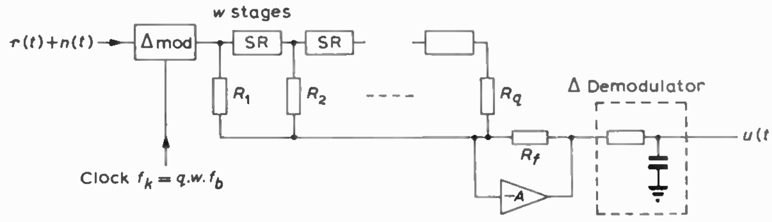


Fig. 9. Waveforms for p.s.k. in matched filter.



(a) Linear transversal filter.

Fig. 8



(b) Deltamod approximation of linear filter.

7 Conclusions

A new and useful approach has been shown to the implementation of a data modulator which is economical, easy to design and is capable of generating many modulation schemes, even one which uses arbitrary signalling waveforms. The principles have been established and extended to include the realization of a matched filter receiver.

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Manuscript first received by the Institution on 11th October 1973 and in final form on 28th November 1973. (Paper No. 1593/Com. 93)

UDC 621.317.77:621.376.22

The application of suppressed carrier double sideband signals to group delay measurements

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SUMMARY

A technique is described in which a suppressed carrier d.s.b. test signal is used to measure the variation in group delay within the baseband of a microwave video channel. The method is shown to be capable of avoiding errors caused both by differential sideband attenuation and by unwanted modulation components. A resolution of 5 ns has been achieved in the frequency range 0·5 to 10 MHz.

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1 Introduction

In order to minimize intersymbol interference in a data transmission system, knowledge of the group delay characteristic of the communication link is required. The measuring technique to be described in this paper was developed in the course of an investigation concerned with the use of a microwave television link for the transmission of data. The data, originating at the Daresbury Nuclear Physics Research Laboratory, are transmitted to Winter Hill where a repeater returns the signal to the laboratory. For experimental purposes, therefore, both ends of the channel are accessible at the same location. Information was required concerning the variation of group delay within the baseband with an accuracy better than ± 10 ns.

2 Choice of Method

A group delay characteristic may in principle be obtained by simple numerical differentiation of a measured phase characteristic. This method is commonly used in measurements on audio and r.f. networks but it is not well suited to measurements on a long link, where the absolute group delay, in our case 220 μs, is very large in comparison to the small differences to be measured.

Under these circumstances a well-known technique is to transmit an amplitude modulated signal and to examine changes in the phase of the modulation envelope as the carrier frequency is varied (see, for example, Reference 1). At a sufficiently low modulating frequency the phase characteristic within the signal bandwidth may be approximated by a straight line. The time delay suffered by the envelope is then equal to the group delay at the carrier frequency. This method, which appears to have been first used by Nyquist and Brand[2] and subsequently by van Weel,[3] essentially entails a sensitive phase comparison of the output from the envelope detector with the original modulation. As absolute group delay is not required, it is convenient to introduce a variable phase shift in one of these signals before presenting them to a phase comparator. The phase difference may then be reduced to a null at some chosen carrier frequency, whereupon changes in group delay are observed as departures from the phase null when the carrier frequency is varied. In the present application, for example, with a 100 kHz modulation, a change of 10 ns in the group delay would produce a departure from a phase null of 0·36°. This sensitivity is fairly easily achieved, especially at the relatively low and fixed modulating frequency, while the method has the advantage of being insentitive to a small drift in carrier frequency.

In applying this technique, two sources of error need careful consideration. The output from the envelope detector may contain, in addition to the true modulation, a small component at modulating frequency, arising either from transmission over the link, due to breakthrough or unbalance in the modulator, or else from capacitive pickup in the laboratory. The phase of this spurious signal will in general differ from that of the received modulation and its amplitude need therefore

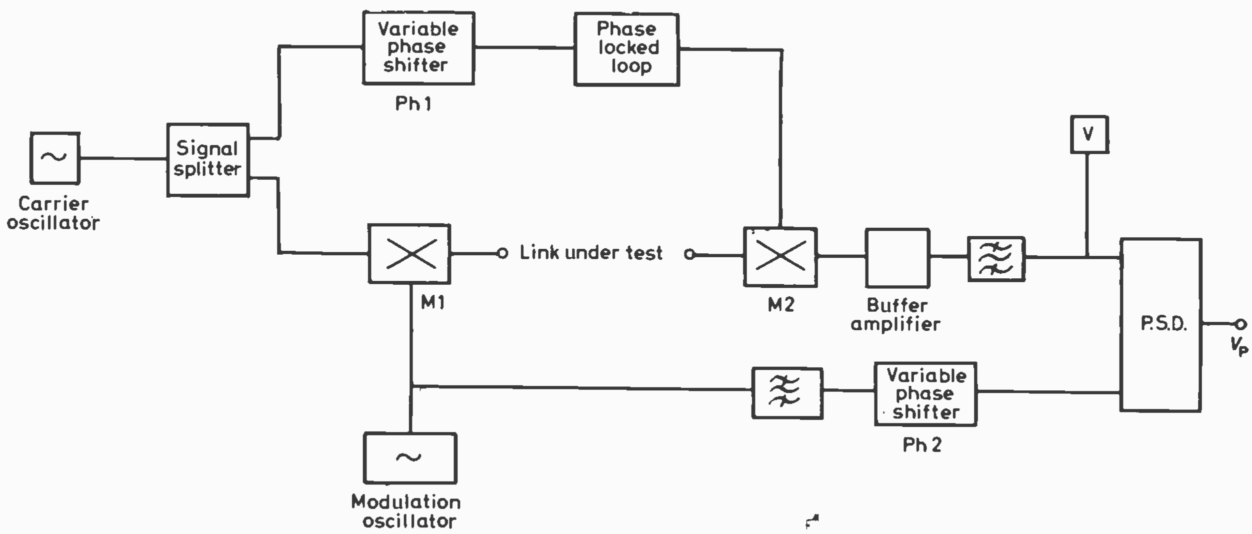

Fig. 1. Block diagram of the apparatus.

be only very small to produce a significant phase shift. It is important to ensure that phase errors from this cause are either negligible or remain constant. A second source of error may appear at carrier frequencies where the amplitude characteristic of the link varies sufficiently rapidly with frequency to produce differential sideband attenuation. This results in distortion of the envelope, so that the behaviour of the phase comparator in the presence of such distortion must be considered.

In the present work the a.m. signal and envelope detector method was not employed, as it was found that the sources of error just mentioned could be treated effectively by using a double sideband (d.s.b.) suppressed carrier signal, followed by a synchronous detector. This technique, which does not appear to have been discussed in previous literature, will be described below. A fourth technique, namely, the use of frequency modulation, was also considered in view of the insensitivity to differential sideband attenuation of narrowband f.m. signals. The final choice, however, fell on the d.s.b. method as being easier to implement.

2.1 The Principle of the D.S.B. Method

The input to the link or network under test is a suppressed carrier double sideband signal with carrier angular frequency ω_c and modulation angular frequency ω_m, thus

$$v_1(t) = 2a \cos \omega_m t \cos \omega_c t \tag{1}$$

where a represents the sideband amplitude.

If A and ψ represent the gain and phase lag introduced by the link and A' and ψ' are their derivatives with respect to ω, then it may be shown that the output signal may be written

$$v_2(t) = 2aA \cos \omega_m(t-\psi') \cos (\omega_c t-\psi) - -2aA'\omega_m \sin \omega_m(t-\psi') \sin (\omega_c t-\psi). \tag{2}$$

The first term shows the modulation to have suffered the group delay, ψ', while the second term shows an additional component which arises due to the slope A' of the amplitude characteristic. We assume that the phase of the carrier is unknown at the receiver where the demodulator multiplies the received signal $v_2(t)$ by $\cos (\omega_c t-\psi+\varepsilon)$. Performing this multiplication and selecting the low-frequency terms, the demodulator output $y(t)$ is given by

$$\begin{aligned} y(t) &= aA \cos \varepsilon \cos \omega_m(t-\psi')+aA'\omega_m \sin \varepsilon \sin \omega_m(t-\psi') \\ &= Y \cos [\omega_m(t-\psi')-\beta] \end{aligned} \tag{3}$$

where

$$Y^2 = a^2[A^2 \cos^2 \varepsilon + A'^2\omega_m^2 \sin^2 \varepsilon] \tag{4}$$

and

$$\tan \beta = \frac{A'\omega_m \tan \varepsilon}{A}. \tag{5}$$

The group delay, ψ', cannot in general be obtained simply by comparing the phase of the demodulator output with that of the input modulation because of the additional phase shift β. Further, unless $A' = 0$, β cannot be guaranteed small because the factor $\tan \varepsilon$ in equation (5) may be large. However, the condition $\beta = 0$ is satisfied if the carrier phase error, ε, is zero. This is also the condition which makes Y a maximum.

Thus the principle of the method is to adjust the phase of the re-injected carrier until the demodulator output is a maximum, whereupon the phase of the output signal gives a reliable measure of the group delay.

3 Design Considerations.

The overall system is shown schematically in Fig. 1. The heart of the system is the phase comparator for which a commercial phase sensitive detector (p.s.d.) and associated phase shifter (Ph2) were employed. The resolution of the apparatus depends on the stability of this combination in terms of the d.c. drift of the output of the p.s.d. and the phase stability of the phase shifter. It should be noted that the latter was placed in the reference channel, rather than in the signal channel, to avoid the small but significant variation of phase with input level which was observed to occur.

The modulator M1 was a wideband double balanced ring modulator. Careful attention was paid to the operating conditions so as to reduce any distortion of the modulation to a minimum and also to minimize components grouped around harmonics of the carrier frequency. These components suffer a group delay appropriate to their own part of the baseband and are synchronously demodulated by M2. The difference in group delay might be as much as 100 ns. However, the corresponding difference in the phase of the 100 kHz modulation is so small that phase errors from this source are approximately proportional to the percentage harmonic content, which can be kept small.

A monolithic switching modulator was found suitable for the demodulator M2. The required adjustment of carrier phase was accomplished using a simple three-stage CR network as the phase shifter Ph1. The output from Ph1 was taken to a phase-locked loop so as to maintain the drive to M2 at a constant level. This requirement was essential because the correct carrier phase at M2 was established, as explained in the previous Section, by adjusting Ph1 for peak amplitude of the demodulated output. This amplitude was found to depend on carrier drive.

A buffer amplifier and a band-pass filter in the signal channel of the p.s.d. ensured that phase measurement errors due to harmonic distortion of the modulation were eliminated. This filter also introduces a phase shift, which varies with modulating frequency, and so an identical filter was inserted in the reference channel to compensate for frequency drift of the modulating oscillator. Both filters were simple LC resonant circuits with a Q-factor of approximately 10.

4 Operation and Calibration

The signal input to the p.s.d. is made up of the demodulated signal, amplitude V_m and phase $\phi-\pi/2$ with respect to the reference channel of the p.s.d. together with much smaller components, also at the modulating frequency, due to transmission directly over the channel or pickup in the laboratory. The latter may be considered together as forming a single stray component of amplitude V_s and phase $\theta-\pi/2$. The d.c. output from the p.s.d. may therefore be written

$$V_p = KV_m \sin\phi + KV_s \sin\theta + V_0 \tag{6}$$

where K is a known constant of the p.s.d. and V_0 is a small adjustable d.c. offset. The amplitude V_m, which has been set to a maximum by adjusting Ph1, is measured using the a.c. voltmeter V in Fig. 1, assuming $V_m \gg V_s$. The phase shifter Ph2 may be adjusted until $V_p/KV_m \ll 1$, whereupon ϕ becomes a small angle and we may write

$$\frac{V_p}{KV_m} = \phi + \frac{V_s \sin\theta}{V_m} + \frac{V_0}{KV_m}. \tag{7}$$

When the carrier frequency is changed, any variation in group delay will change the value of ϕ, while a variation in attenuation will also change V_m. Thus the error terms containing V_s and V_0 in the above equation do not remain constant and a change in group delay cannot be derived simply from the change in V_p/V_m. With the arrangement in Fig. 1, however, it is possible to unlock the phase-locked loop. This reduces V_m to zero while leaving unaltered θ and also V_s, since the demodulator is still being switched. The offset V_0 may then be adjusted until $V_p = 0$ so that, from equation (6),

$$KV_s \sin\theta + V_0 = 0. \tag{8}$$

Equation (7) then becomes

$$\frac{V_p}{KV_m} = \phi$$

and thus the phase ϕ of the received modulation with respect to the reference channel of the p.s.d. is accurately known. Finally we may write

$$t_g = \frac{\phi}{\omega_m} = \frac{V_p}{KV_m\omega_m} \tag{9}$$

where t_g will be termed the relative group delay. When the d.s.b. carrier frequency is varied, changes in the true group delay, ψ', produce identical changes in t_g.

The validity of the procedure for obtaining cancellation of the error terms in equation (6) could be checked by reversing the connexions from the modulating oscillator to the modulator. After setting up the apparatus in the way described, this was always found to reverse the sign of V_p without affecting its magnitude.

The equipment was first tested by making relative group delay measurements on some simple networks. The measured delay of a length of coaxial cable was constant to within ± 5 ns in the 0·5 to 10 MHz range. A simple CR integrating network also gave results in agreement with theory, as did a third-order low-pass Butterworth filter whose -3 dB point lay at 5·0 MHz. Measured and predicted values agreed within 10 ns.

5 Measurements on the Microwave Link

The link installed at Daresbury is a C-band microwave link, designed for a baseband width of 5·5 MHz. Measurements of t_g, relative to 4·0 MHz, were made in the range 1·0 to 9·0 MHz. These measurements were repeated at intervals over a six-week period and the mean values plotted. The resulting graph was extrapolated to zero frequency and is shown in Fig. 2 where the group delay is shown relative to the extrapolated value at zero frequency. The vertical bars show the

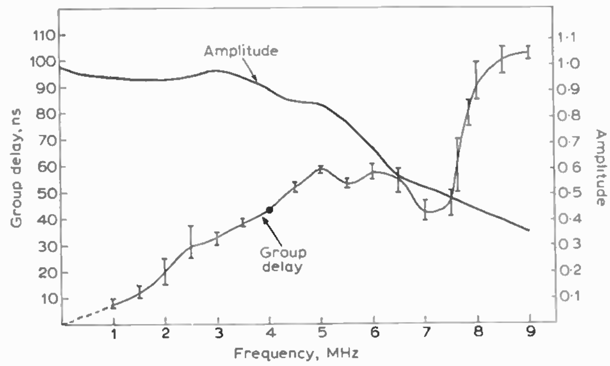

Fig. 2. Group delay and amplitude response in the baseband of the microwave link (both curves relative to zero frequency.)

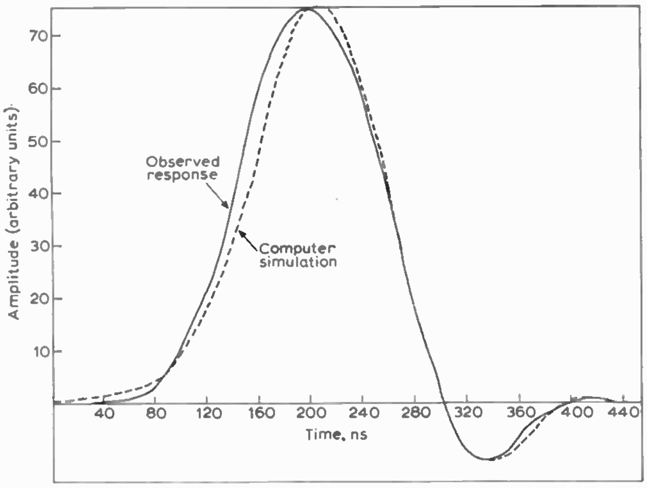

Fig. 3. Observed and predicted response of the link to a 120 ns rectangular pulse.

maximum scatter of the individual observations. This was almost everywhere less than ± 5 ns.

The amplitude characteristic, also shown in Fig. 2, was measured in the same frequency range and this information combined with the group delay measurements was used to compute the response of the link to a rectangular 120 ns pulse using a fast Fourier transform routine. The predicted pulse response is compared with the measured pulse response in Fig. 3 where, since the absolute delay is unknown, the predicted response has been time-shifted to give coincidence at the first zero crossing. Elsewhere the agreement is seen to be very close, giving further confidence in the accuracy of the group delay measurements.

6 Conclusions

A technique has been developed which uses a suppressed carrier d.s.b. signal to measure relative group delay. The method offers certain advantages over the a.m. method in that the difficulties which may arise due to modulation breakthrough and differential sideband attenuation are easily overcome. The use of balanced modulators considerably reduces direct modulation breakthrough. Differential sideband attenuation results in distortion of the demodulated signal when an envelope detector is used in the a.m. method, unless the modulation depth is small. Errors from this effect can, however, be made negligible even at full transmitted sideband amplitude using a synchronous detector in the way described.

In the frequency range up to 10 MHz a resolution of approximately 5 ns has been achieved using only standard laboratory equipment in conjunction with some of the recently developed integrated circuit packages. The method has been applied to a rather special situation in which both ends of the link were accessible in the same laboratory. On any duplex link, however, a similar situation may easily be arranged.

7 Acknowledgments

The authors are grateful to Professor J. D. Craggs of the Department of Electrical Engineering and Electronics at the University of Liverpool for the facilities of this Department in constructing and testing the equipment. One of the authors (S.R.E.) is indebted to the Science Research Council for financial support.

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Manuscript first received by the Institution on 12th December 1973 and in final form on 22nd March 1974. (Short contribution No. 174/MST3)

UDC 621.396.677.44

Variational approach to V-dipole antenna analysis

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SUMMARY

An analysis of the problem of symmetric, centre-driven V-antenna, using the variational method, is given. In deriving the variational expression for impedance the trial function in the form of a functional series with complex coefficients was assumed. The concrete evaluations were performed using two types of functional series: two-term trigonometric series and polynomials of arbitrary order.

The method was used for calculating the input impedance and current distribution for a number of representative examples. The numerical results are in very good agreement with the results obtained by the point-matching method. In addition, the variational results for admittances corresponding to the particular case of a V-dipole completely stretched out are compared with available theoretical and experimental data for the straight symmetrical dipole. The agreement was found to be excellent.

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1 Introduction

In a recent paper by the present authors[1] the problem of the centre-driven V-dipole antenna has been exhaustively treated using an integral-equation technique. The integral equation was numerically solved by the so-called point-matching method, approximating the current distribution function by a polynomial series with unknown coefficients. In that paper an announcement of the present paper was made since the main results were available at the time.

The variational approach to the problem of a thin straight dipole is due to J. E. Storer.[2,3] Later, the variational method has been successfully applied to the analysis of some other types of isolated straight cylindrical antennas[4–6] as well as to the analysis of arrays of coupled dipoles.[7–10] In all these cases a two-term trigonometric trial function for current has been adopted. Recently the polynomials were used in solving the cylindrical antenna problem by the variational method.[11]

The purpose of this paper is to provide a variational solution to the problem of impedance of the centre-driven V-antenna. After giving a brief outline of the basic theory, the variational expression for the driving-point impedance is presented. In deriving the final formula for impedance a trial function for current in the form of a functional series with unknown coefficients was assumed. The practical evaluations were performed by using two types of functional series: 1° the two-term Storer's trigonometric approximation and 2° the Mth order polynomial with complex coefficients.

The results obtained by the present variational method (using both types of trial functions) were compared with those obtained by the point-matching method, described in Ref. 1. The agreement between results was found to be very satisfactory. Due to the lack of other published data concerning isolated V-dipoles,‡ the particular results, corresponding to the straight dipole, were compared with the results obtained by Storer's [2,3] and Popović's[13] methods, and with the experimental results of Mack.[14]

2 Variational Expression for Impedance

Consider a thin symmetric V-dipole antenna, the arms of which are inclined at an arbitrary angle θ with respect to each other. Both arms have equal lengths h and the same radius a ($a \ll h$). As shown in Fig. 1, the axis of arm 1 coincides with the z-axis of the coordinate system, whose origin is at the apex of the dipole. Arm 2 is along the ζ-axis lying in the y–z plane. The antenna conductors are assumed to be perfect.

In order to simplify the analysis, an idealized model of the excitation zone is adopted. Assuming that the two halves of the dipole form an unbroken conductor, the real generator is replaced by an impressed field whose components E_z^i and E_ζ^i act axially across the narrow belts

‡ In a recent paper P. Silvester and K. K. Chan[12] used Bubnov–Galerkin method in treating the problem of the impedance of two coupled V-antennas. Unfortunately the data concerning the impedance of an isolated V-antenna are not available in their work.

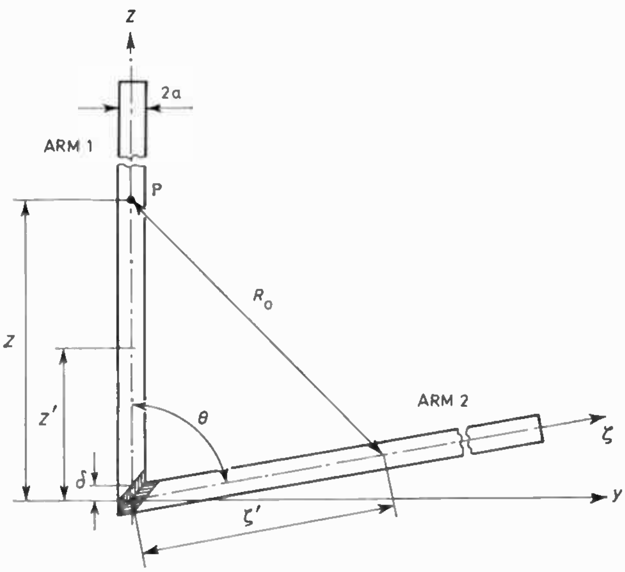

Fig. 1. V-antenna.

of the width δ on the cylindrical surfaces of the corresponding dipole-arms. The driving voltage can be expressed as the line integral

$$V = \lim_{\delta \to 0} \int_{\delta}^{0} E_{\zeta}^{i}\, d\zeta + \int_{0}^{\delta} E_{z}^{i}\, dz \tag{1}$$

calculated along the cylinder-generators lying in the plane $x = a$. Because of the symmetry the two components E_z^i and E_ζ^i must satisfy the condition

$$E_\zeta^i(\zeta) = -E_z^i(z) \qquad \zeta = z \tag{2}$$

so that

$$V = 2 \lim_{\delta \to 0} \int_0^{\delta} E_z^i\, dz. \tag{3}$$

In terms of the Dirac delta function generator E_z^i can be expressed as

$$E_z^i = V\delta(z). \tag{4}$$

This model is especially suitable in the case of a strip V-dipole, whose equivalent radius is $a = d/4$, d being the width of the strip.

Since the conductor is assumed to be perfect, the tangential component of the total electric field strength must vanish at the surface of the conductor. Denoting by

$$E_z = -j\frac{\omega}{k^2}\left[\frac{\partial}{\partial z}\operatorname{div}\mathbf{A} + k^2 A_z\right] \tag{5}$$

the z-component of the electric field strength due to the current and charges on the antenna, the boundary condition at a point P on the surface of the arm 1 can be written in the form

$$E_z^i = V\delta(z) = j\frac{\omega}{k^2}\left[\frac{\partial^2 A_z}{\partial z^2} + \frac{\partial^2 A_y}{\partial y \partial z} + k^2 A_z\right] \tag{6}$$

where $\mathbf{A}$ ($A_x = 0$) is the magnetic vector-potential, $k = \omega/c$ the free space propagation constant and ω the angular frequency.

Due to the symmetry, the currents in the two arms must satisfy the condition

$$I(\zeta) = -I(z) \qquad \text{when} \qquad \zeta = z. \tag{7}$$

In addition the current distribution function must fulfil the condition

$$I(h) = 0. \tag{8}$$

When writing the expression for the magnetic vector-potential it will be assumed that the entire current is localized in the axes of the conductors. So, we have

$$A_z = \frac{\mu_0}{4\pi}\left[\int_0^h I(z')\frac{e^{-jkr}}{r}\, dz' + \cos\theta \int_0^h I(\zeta')\frac{e^{-jkR_0}}{R_0}\, d\zeta'\right], \tag{9}$$

$$A_y = \frac{\mu_0}{4\pi}\sin\theta \int_0^h I(\zeta')\left.\frac{e^{-jkR}}{R}\right|_{y=0} d\zeta' \tag{10}$$

where

$$r = \{(z-z')^2 + a^2\}^{1/2} \tag{11}$$

$$\begin{aligned} R_0 &= \{z^2 + \zeta'^2 + a^2 - 2\zeta' z\cos\theta\}^{1/2} = R|_{y=0} \\ &= \{z^2 + \zeta'^2 + y^2 + a^2 + 2\zeta'(-z\cos\theta + y\sin\theta)\}^{1/2}|_{y=0}. \end{aligned} \tag{12}$$

$R_0 = R|_{y=0}$ in (9), (10) and (12) is the distance between $d\zeta'$ on the axis of the conductor 2 and the point P($x = a$, $y = 0$, z) on the surface of the conductor 1. The second form of R_0, containing the coordinate y, is necessary in the evaluation of $\partial A_y/\partial y$ at P.

Taking into account (7) and substituting ζ by z, eqn. (6) can be put in the form

$$V\delta(z) = j\frac{30}{k}\int_0^h I(z')K(z, z')\, dz', \tag{13}$$

where the kernel $K(z, z')$ is defined by

$$K(z, z') = \left[k^2 + \frac{\partial^2}{\partial z^2}\right]\left[\frac{e^{-jkr}}{r} - \cos\theta\frac{e^{-jkR_0}}{R_0}\right] - \\ -\sin\theta \frac{\partial^2}{\partial y \partial z}\left.\frac{e^{-jkR}}{R}\right|_{y=0} \tag{14}$$

and

$$\frac{30}{k} = \frac{\omega}{k^2}\frac{\mu_0}{4\pi}\ \text{ohms}.$$

In order to obtain the variational expression for the impedance, let us multiply (13) by $I(z)dz$ and integrate it from 0 to h:

$$\int_0^h V\delta(z)I(z)\, dz = j\frac{30}{k}\int_0^h\int_0^h I(z)I(z')K(z, z')\, dz'dz. \tag{15}$$

Having in mind that

$$\int_0^h V\delta(z)I(z)\, dz = \tfrac{1}{2}VI(0) = \tfrac{1}{2}I^2(0)Z, \tag{16}$$

where $Z = V/I(0)$ is the input impedance, we can write

$$Z = j\frac{60}{k}\frac{1}{I^2(0)}\int_0^h\int_0^h I(z)I(z')K(z, z')\, dz'dz. \tag{17}$$

Finally, if we introduce the normalized current distribution function

$$g(z) = I(z)/I(0), \quad \text{with} \quad g(0) = 1 \quad \text{and} \quad g(h) = 0, \tag{18}$$

we obtain

$$Z = j(60/k)\int_0^h\int_0^h g(z)g(z')K(z, z')\, dz'dz. \tag{19}$$

Since the kernel (14) remains unchanged when z and z' are mutally substituted, it can be shown in a manner similar to that described in Ref. 3 that the first variation of Z with respect to small variations in $g(z)$ is zero, i.e. that

$$\delta Z = 0. \tag{20}$$

The integral (19) has the stationary property. Introducing a reasonably good approximation for the current distribution function, a higher order approximation for the computed impedance will be obtained.

In order to transform (19) into a form suitable for numerical calculations on a digital computer, let us first consider the simple integral

$$\int_0^h g(z')K(z,z')\,dz' = k^2\int_0^h g(z')\frac{e^{-jkr}}{r}\,dz' - k^2\cos\theta\int_0^h g(z')\frac{e^{-jkR_0}}{R_0}\,dz' + \int_0^h g(z')\frac{\partial}{\partial z^2}\left[\frac{e^{-jkr}}{r}\right]dz' - \cos\theta\int_0^h g(z')\frac{e^{-jkR_0}}{R_0}\,dz' - \sin\theta\int_0^h g(z')\frac{\partial^2}{\partial y\partial z}\left[\frac{e^{-jkR}}{R}\right]_{y=0}dz' \tag{21}$$

After partial integrations and suitable transformations, taking into account that $g(h) = 0$, $g(0) = 1$ and

$$\sin\theta\frac{\partial R}{\partial y} = \frac{\partial R}{\partial z'} + \cos\theta\frac{\partial R}{\partial z},$$

(21) can be rearranged into the following form:

$$\int_0^h g(z')K(z,z')\,dz' = \frac{\partial}{\partial z}\left[\frac{e^{-jkr}}{r} - \frac{e^{-jkR_0}}{R_0}\right]_{z'=0} - g'(h)\left.\frac{e^{-jkr}}{r}\right|_{z'=h} + g'(0)\left.\frac{e^{-jkr}}{r}\right|_{z'=0} + \int_0^h [k^2g(z')+g''(z')]\frac{e^{-jkr}}{r}\,dz' - k^2\cos\theta\times\int_0^h g(z')\frac{e^{-jkR_0}}{R_0}\,dz' - \int_0^h g'(z')\frac{\partial}{\partial z}\left[\frac{e^{-jkR_0}}{R_0}\right]dz'. \tag{22}$$

Multiplying (22) by $g(z)dz$ and integrating it from 0 to h, and after performing some partial integrations, we finally obtain

$$Z = j\frac{60}{k}\Bigg[-g'(h)\int_0^h g(z)\left.\frac{e^{-jkr}}{r}\right|_{z'=h}dz + g'(0)\int_0^h g(z)\times\left.\frac{e^{-jkr}}{r}\right|_{z'=0}dz + \int_0^h\int_0^h g(z)\{k^2g(z')+g''(z')\}\times\frac{e^{-jkr}}{r}\,dz'dz - k^2\cos\theta\int_0^h\int_0^h g(z')g(z)\times\frac{e^{-jkR_0}}{R_0}\,dz'dz + \int_0^h g'(z')\left.\frac{e^{-jkR_0}}{R_0}\right|_{z=0}dz' + \int_0^h\int_0^h g'(z)g(z')\frac{e^{-jkR_0}}{R_0}\,dz'dz\Bigg]. \tag{2}$$

3 Variational Solution for Impedance with a Functional Series as the Trial Function

Let us now approximate the normalized current distribution function by a functional series with unknown coefficients A_m:

$$g(z) = \sum_{m=0}^{M} A_m f_m(z). \tag{24}$$

Then (23) can be written in the form

$$Z = \sum_{m=0}^{M}\sum_{n=0}^{M} A_mA_nw_{mn} + \sum_{m=0}^{M} A_mv_m = \sum_{m=0}^{M} A_mv_m + A_0^2w_{00} + 2A_0A_1w_{01} + 2A_0A_2w_{02} + \ldots + 2A_0A_Mw_{0M} + A_1^2w_{11} + 2A_1A_2w_{12} + \ldots + 2A_1A_Mw_{1M} + \ldots + A_M^2w_{MM}, \tag{25}$$

where

$$w_{mn} = j\frac{60}{k}\Bigg[-f_n'(h)\int_0^h f_m(z)\left.\frac{e^{-jkr}}{r}\right|_{z'=h}dz + f_n'(0)\int_0^h f_m(z)\left.\frac{e^{-jkr}}{r}\right|_{z'=0}dz + \int_0^h\int_0^h f_m(z)\{k^2f_n(z')+f_n''(z')\}\frac{e^{-jkr}}{r}\,dz'dz - k^2\cos\theta\int_0^h\int_0^h f_m(z)f_n(z')\frac{e^{-jkR_0}}{R_0}\,dz'dz + \int_0^h\int_0^h f_m'(z)f_n'(z')\frac{e^{-jkR_0}}{R_0}\,dz'dz, \tag{26}$$

and

$$v_m = \int_0^h f_m'(z')\left.\frac{e^{-jkR_0}}{R_0}\right|_{z=0}dz'. \tag{27}$$

Since the expression for impedance is stationary, the unknown coefficients A_m can be determined by requiring that the first variation of Z vanishes, i.e.

$$\delta Z = 0. \tag{20}$$

It should be noted that $M+1$ coefficients A_m are not independent because they are subject to two constraints:

$$\sum_{m=0}^{M} A_mf_m(h) = 0 \tag{28}$$

$$\sum_{m=0}^{M} A_mf_m(0) - 1 = 0, \tag{29}$$

expressing the conditions $g(h) = 0$ and $g(0) = 1$.

In principle we could solve the constraints for two constants in terms of other $M-1$, substitute them into (25), and arrive at an expression for impedance containing only $M-1$ unknown constants (to be determined). However it is more convenient to use the method of Lagrange multipliers. Thus we multiply (28) and (29) by Lagrange multipliers λ_1 and λ_2, respectively, and add them to the right side of (25). So, we have

$$Z = \sum_{m=0}^{M}\sum_{n=0}^{M} A_mA_nw_{mn} + \sum_{m=0}^{M} A_m\times[v_m + \lambda_1f_m(h) + \lambda_2\{f_m(0)-1\}]. \tag{30}$$

Now we arrive at the following necessary conditions for a stationary value of Z:

$$\frac{\partial Z}{\partial A_m} = 2\sum_{n=0}^{M} A_n w_{mn} + v_m + \lambda_1 f_m(h) + \lambda_2\{f_m(0)-1\} = 0$$

$$m = 0, 1, \ldots M \quad (31)$$

$$\frac{\partial Z}{\partial \lambda_1} = \sum_{m=0}^{M} A_m f_m(h) = 0 \quad (32)$$

$$\frac{\partial Z}{\partial \lambda_2} = \sum_{m=0}^{M} A_m\{f_m(0)-1\} = 0. \quad (33)$$

This gives $(M+1)+2 = M+3$ equations to solve $A_0, A_1, \ldots A_M, \lambda_1$ and λ_2.

Once the A_m coefficients are determined, Z is evaluated from (25), and $g(z)$ from (24). If current distribution $I(z)$, corresponding to the voltage V, is required, it can be obtained from

$$I(z) = I(0)g(z) = \frac{V}{Z}g(z). \quad (34)$$

The foregoing method for calculating impedance has been applied by using two different functional series as trial functions. Primarily, the use was made of the two-term trigonometric approximation,

$$g(z) = A_1 \sin k(h-z) + A_2\{1-\cos k(h-z)\}, \quad (35)$$

originally applied by Storer in his variational solution to the problem of thin straight dipole. Then, a more general functional series in the form of an Mth order polynomial

$$g(z) = \sum_{m=0}^{M} A_m z^m \quad (36)$$

was applied.

The definite integrals in (25) and (26) can be easily integrated numerically by means of an electronic digital computer, using Gaussian quadrature formulae.

4 Numerical Results

The developed variational method has been used in calculating the impedance/admittance and current distribution function for a number of representative examples. Thus, V-dipoles of three different arm-lengths $h = 0{\cdot}250\lambda$, $0{\cdot}375\lambda$ and $0{\cdot}500\lambda$ were taken. In all cases an equivalent radius $a = 0{\cdot}007\lambda$ was adopted. (Mack[14] quotes the very precise value $0{\cdot}007022\lambda$.)

In order to check the variational formula (25) and to compare the results obtained by different trial functions, primarily the admittances corresponding to the particular angles $\theta = 180°$, $135°$, $90°$ and $45°$ were calculated using polynomials of the order $M = 2$ and $M = 3$ and Storer's two-term trigonometric approximation as well. The results are summarized in Table 1. By inspection of the results a very good overall agreement between corresponding admittances can be noted, especially between conductances, irrespective of the trial function which was used.

Two other, very valuable verifications of the formula (25) and the method as a whole can be found in Tables 2 and 3. In order to compare the results obtained by present method with those achieved by other methods and by various authors and also with experimental ones, the admittances of the straight dipole, which is the particular case of the V-dipole for $\theta = 180°$, are given in Table 2. In addition to the theoretical results obtained by Storer's variational[2,3] and Popović's polynomial point-matching methods,[13] the admittances measured by Mack[14] are also included in Table 2. When comparing the admittances from Table 2, both theoretical and experimental, with corresponding values in Table 1 ($\theta = 180°$), we find an agreement between them.

Table 1. Admittances of the V-dipole, in mA/V, calculated for different trial functions ($a = 0{\cdot}007\lambda$)

Angle θ	Trial function	Dipole arm-length $h = 0{\cdot}250\lambda$	$h = 0{\cdot}375\lambda$	$h = 0{\cdot}500\lambda$
180°	Polynom. $M = 2$	8·933−j 3·445	1·575−j 0·227	1·014+j 1·704
	Polynom. $M = 3$	9·026−j 3·488	1·554−j 0·129	0·987+j 1·728
	Trigon. two-term	9·060−j 3·238	1·535−j 0·314	0·973+j 1·425
135°	Polynom. $M = 2$	10·244−j 3·983	1·534−j 0·235	1·032+j 1·820
	Polynom. $M = 3$	10·343−j 4·069	1·510−j 0·134	0·997+j 1·845
	Trigon. two-term	10·436−j 3·696	1·495−j 0·330	0·991+j 1·517
90°	Polynom. $M = 2$	18·940−j 4·671	1·377−j 0·351	0·993+j 2·223
	Polynom. $M = 3$	19·119−j 5·159	1·349−j 0·230	0·942+j 2·252
	Trigon. two-term	19·375−j 3·432	1·340−j 0·471	0·954+j 1·840
45°	Polynom. $M = 2$	10·417+j 27·567	1·051−j 1·186	0·652+j 3·046
	Polynom. $M = 3$	11·948+j 30·163	1·027−j 0·907	0·594+j 3·176
	Trigon. two-term	8·521+j 25·008	1·017−j 1·355	0·610+j 2·415

Table 2. Admittances of the straight cylindrical dipole, in mA/V, ($a = 0{\cdot}007\lambda$), calculated and measured according to some other authors

		Dipole arm-length $h = 0{\cdot}250\lambda$	$h = 0{\cdot}375\lambda$	$h = 0{\cdot}500\lambda$
B.D. Popović polynomial	M=2	8·934−j 3·445	1·575−j 0·228	1·013+j 1·701
	M=3	9·027−j 3·487	1·555−j 0·128	0·988+j 1·728
Storer variational		9·061−j 3·238	1·535−j 0·314	0·972+j 1·424
Mack experimental		8·90 −j 3·46	1·58 −j 0·17	1·02 +j 1·68

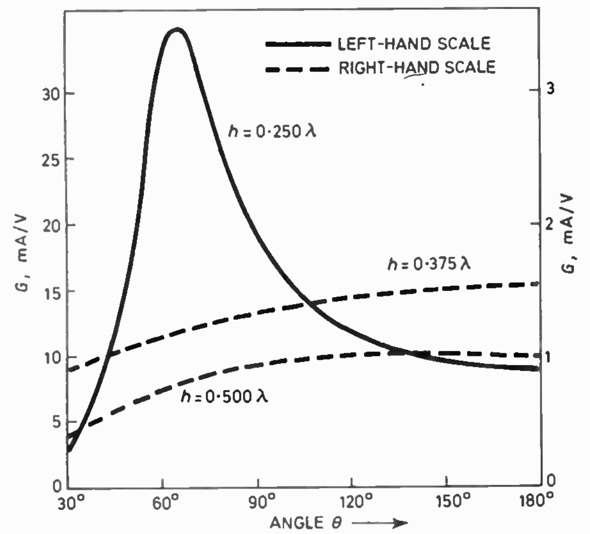

Fig. 2. Conductance of the V-dipole of arm-length h and radius $a = 0{\cdot}007$, versus angle θ.

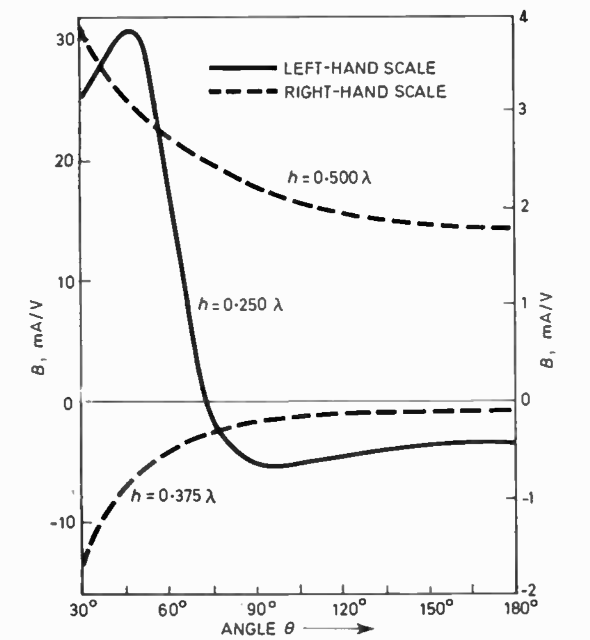

Fig. 3. Susceptance of the V-dipole of arm-length h and radius $a = 0{\cdot}007$ versus angle θ.

In Table 3 the admittances of the V-dipole according to the point-matching method, developed by the present authors,[1] are given. From Tables 1 and 2, a direct comparison between the results obtained by variational and point-matching methods is possible, with polynomial approximation for current used in both cases. Though the order of the polynomial used in point-matching procedure for the dipoles $h = 0{\cdot}375\lambda$ and $0{\cdot}500\lambda$ is one degree higher, i.e. $M = 4$, the agreement between two sets of admittances is very satisfactory.

Table 3. Admittances of the V-dipole, in mA/V, calculated by the point-matching method using a polynomial approximation for current ($a = 0{\cdot}007\lambda$)

Angle θ	Dipole arm-length $h = 0{\cdot}250\lambda$†	$h = 0{\cdot}375\lambda$‡	$h = 0{\cdot}500\lambda$‡
180°	9·16−j 3·55	1·53−j 0·11	0·97+j 1·67
135°	10·52−j 4·11	1·49−j 0·10	0·98+j 1·78
90°	19·55−j 4·80	1·33−j 0·18	0·93+j 2·17
45°	10·71+j 28·51	1·01−j 0·76	0·59+j 3·07

† Third-order polynomial. ‡ Fourth-order polynomial.

Adopting the third-order polynomial as the approximation for current, a more detailed investigation of the dependence of the input admittance on the angle θ was performed. The calculated values of conductances and susceptances for three different arm-lengths ($h = 0{\cdot}25\lambda$, $0{\cdot}375\lambda$ and $0{\cdot}500\lambda$) are shown in Figs. 2 and 3. While the diagrams showing conductances and susceptances of the dipoles $h = 0{\cdot}375\lambda$ and $h = 0{\cdot}500\lambda$ have a monotonic flow, those corresponding to the dipole $h = 0{\cdot}250\lambda$ exhibit a curious and very strong dependence on the angle θ.

The present method was also used for calculating the current distribution along the dipole-arms. As an illustration, the real and imaginary parts of the current distribution function for the dipole having arm-lengths $h = 0{\cdot}250\lambda$ and $h = 0{\cdot}500\lambda$ are calculated and shown in Figs. 4 and 5. These diagrams, corresponding to the

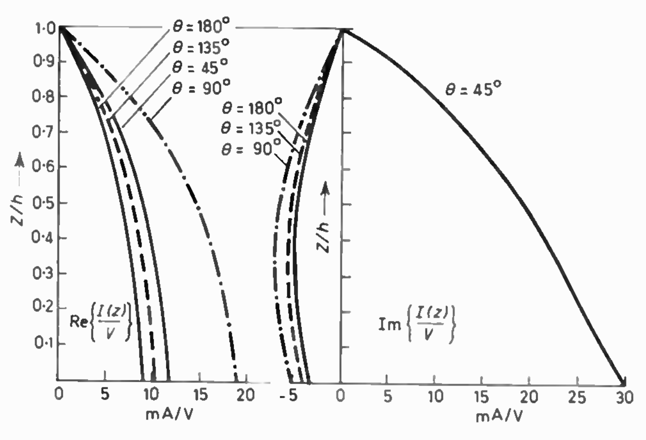

Fig. 4. Real and imaginary parts of current distribution on a V-dipole of arm-length $h = 0{\cdot}250\lambda$ and radius $a = 0{\cdot}007$ for different angles θ.

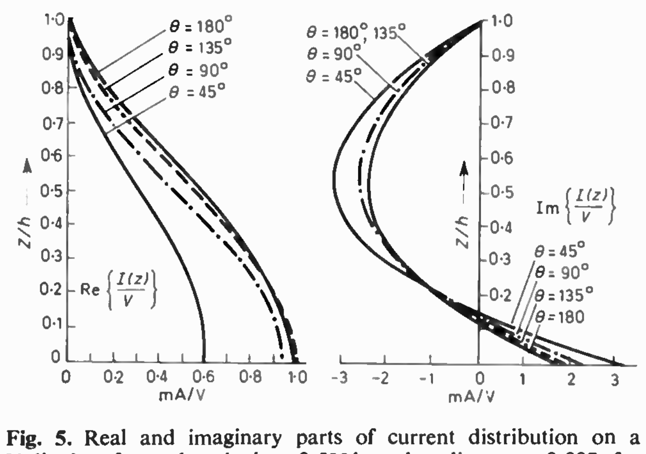

Fig. 5. Real and imaginary parts of current distribution on a V-dipole of arm-length $h = 0{\cdot}500\lambda$ and radius $a = 0{\cdot}007$ for different angles θ.

third-order polynomial approximation, are in very good agreement with those obtained by the point-matching method.[1]

5 Conclusions

A variational approach to the V-dipole antenna analysis is presented. The variational expression for the impedance was derived by using a functional series with complex coefficients as a trial function. In evaluating the input impedance and current distribution for a number of representative examples, the polynomials of the second and third order, as well as a two-term trigonometric series were employed. The results obtained were compared with those obtained by the point-matching method (developed by the same authors), with polynomial approximation for current. Although the overall agreement between the results obtained by the two methods is excellent, the variational method has the advantage that it yields the same degree of accuracy with a lower order of the polynomial.

6 Acknowledgment

This work was supported by the Serbian National Research Foundation.

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Manuscript first received by the Institution on 18th September 1973 and in final form on 4th December 1973. (Paper No. 1594/Com. 94)

UDC 621.376.56: 621.397.2

Application of p.c.m. to broadcast quality video signals

Part I. Subjective Study of the Coding Parameters

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SUMMARY

This Part describes subjective tests performed to establish the sampling rate and the number of quantizing levels or corresponding bits per sample which should be used to obtain broadcast-quality television pictures from a p.c.m. system handling PAL colour, 625-line, 5·5 MHz video signals according to the UK standard (CCIR Standard I). These tests included an investigation of adding 'dither' signals to the video signals prior to the p.c.m. encoder as a means of reducing the required number of bits per sample. An effective dither signal is proposed which breaks up coherent quantizing errors such as 'contouring' effects and replaces them with less noticeable random quantizing errors.

A theoretical estimate of the required number of bits per sample is derived by comparing calculated values of the magnitude of quantizing noise with the existing specification for random noise allowable on analogue links.

The paper also gives brief details of the p.c.m. coder and decoder developed by the author and used in the tests carried out in the BBC Research Department.

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1 Introduction

Recent advances in digital technology have made it possible, in principle, to replace much of the analogue equipment used in the processing and transmission of television signals by equivalent digital equipment. It has already found application in processing such as standards conversion. Although digital methods may not be applied extensively to television transmission for some years, the specification of p.c.m. links capable of handling digital video signals is already being given considerable attention by international committees. It is therefore essential that the optimum parameters of a digital system capable of handling colour television signals of broadcast quality should be known.

In selecting a coding system for television, both point-to-point transmission and signal processing operations have to be considered.

For transmission along national and international networks, a coding system which can accept the standard composite colour signal (as transmitted now over analogue links) is desirable, because most networks will, during their evolutionary stages, have analogue sections interspersed with the digital sections. Digital systems in which the luminance and colour difference signals are encoded separately[1,2] have the disadvantage that the additional colour coding and decoding processes required at the ends of the digital sections are likely to cause a noticeable degradation of the final picture quality.

As far as signal processing is concerned, arithmetic operations such as mixing, aperture correction and standards conversion can be effected much more easily with linear p.c.m. signals than with those provided by other coding systems such as differential p.c.m. or delta modulation.

The optimum coding law to use for p.c.m. would be one giving constant visibility of the brightness changes corresponding to each change of one quantum level in the video signal. This condition is very nearly satisfied by a linear coding law applied to gamma-corrected video signals. Since the optimum law would involve considerable extra instrumental difficulties, a linear law is the most suitable for practical purposes. Linear coding also has the advantage that quantizing errors in the colour subcarrier are independent of the luminance level.

From these discussions, it can be seen that linear p.c.m. coding of composite colour signals is one of the most useful means of providing digital video signals for general applications to television broadcasting.

This paper, in two parts, describes subjective tests carried out to determine suitable parameters for a broadcast-quality p.c.m. system handling PAL colour 5·5 MHz, 625-line video signals, according to the U.K. Standard (CCIR Standard I). The first part includes a description of the p.c.m. coder and decoder used in the tests (Section 3).

The parameters which were examined were:

(1) Sampling frequency.

(2) Number of digits per sample.

(3) Effect of transmission errors.

(4) Effect of jitter in the analogue samples obtained from a digital decoder.

These parameters fall into two groups. The sampling frequency and the number of bits per sample are fundamental coding parameters of a p.c.m. system and are considered here (Part 1), whereas transmission errors and jitter are effects which result from instrumental deficiencies and are considered in Part 2.

Before the subjective tests were undertaken,[3] an attempt was made to deduce the number of bits required per sample by comparing the theoretical quantizing noise power produced by different numbers of bits with the random noise power allowable in analogue systems. This theoretical analysis is described in Section 2.2.3 and it will be seen that the results obtained were in good agreement with the results of the subjective tests given in Section 5.

2 Factors Affecting Choice of Fundamental Coding Parameters

2.1 Sampling Frequency

By Nyquist's sampling theorem, the theoretical minimum sampling frequency which can be used without introducing unwanted aliasing components into the decoded video signal is 11 MHz for a video bandwidth of 5·5 MHz.† However, to allow for the finite rate of cut-off of the 5·5 MHz low-pass filters which are required before and after the sampling process, the minimum sampling frequency which can be used in practice is about 12 MHz. The use of sampling frequencies closer to 11 MHz places very stringent requirements on the filter design and circuit complexity becomes the limiting factor. Filters giving satisfactory results at a sampling frequency of 11·9 MHz have been constructed[4] but the considerable group delay introduced by these filters near 5·5 MHz would limit the number of codecs‡ which could be cascaded to about three or perhaps four.

In addition to aliasing components, unwanted components are also introduced as a result of the quantization process involved in p.c.m. coding. In coloured areas of a picture, the visibility of this form of distortion varies as the sampling frequency is altered and is minimized if the sampling frequency is an exact multiple of the colour subcarrier frequency. A suitable multiple for System I video signals is three since it gives a sampling frequency of 13·3 MHz which is only slightly above the minimum practical sampling frequency.

System requirements can also affect the final choice of the sampling frequency. In this connexion, a sampling frequency of three times the subcarrier frequency again has advantages since it would simplify some processing operations such as digital transcoding between primary (R, G and B) and PAL colour signals. Further advantages are to be gained if the phase of such a sampling frequency at the sampling gate is fixed relative to the subcarrier in each digital coder throughout a transmission chain; this would greatly simplify operations such as mixing which involve more than one video signal. Alternatively, digital processing operations in which the signal is required to be delayed by exact multiples of a line period are made easier if the sampling frequency is an integral multiple of line frequency. On the other hand, the need to multiplex digital video signals with other forms of data may mean that the sampling frequency will be dictated by other users of the point-to-point transmission system.

Taking into account the various factors given above, three different sampling frequencies were selected for use in subjective tests to examine the variation in the visibility of quantizing errors at different sampling frequencies; these three frequencies were:

(i) Three times subcarrier frequency, i.e. 13·301 MHz.

(ii) 851 times line frequency, i.e. 13·297 MHz. (The multiple 851 gives the line-locked frequency as close as possible to three times subcarrier frequency.)

(iii) An unlocked frequency, within 1% of (i) and (ii), which was selected to provide maximum visibility beat patterns in colour areas of pictures.

The frequencies will be denoted by $3f_{sc}$, $851f_L$ and 'unlocked' respectively.

Details of these tests are given in Section 4.

2.2 Number of Bits per Sample

2.2.1 *Effects of insufficient bits per sample*

In binary p.c.m., each sample is represented by a number containing n binary digits (bits), and the analogue signal effectively transmitted and recovered by decoding the digital signal is quantized into 2^n separate levels. The types of picture impairment caused by this quantization process can conveniently be divided into three types:

(a) Contouring effects: Areas of the picture in which the brightness varies slowly with position are represented by patches of uniform brightness separated by sharp transitions. The appearance of this effect is reminiscent of shaded contours on a map. Contouring effects tend to be masked by fine patterns which cause the signal to cross quantum levels at closely spaced intervals; since colour subcarrier has a similar effect, contours are not in general as noticeable with colour signals as with monochrome signals.

(b) Beat patterns on colour pictures. These have been discussed in Section 2.1 and result from the intermodulation of the colour subcarrier and sampling frequency components. They are most noticeable in areas of constant hue and saturation.

(c) Increase in noise. Random noise and high-frequency picture detail in the video signal at the input of the digital coder tends to break up both contouring and beat patterning effects; the quantizing errors then appear as additional noise on the picture.

† It should be noted that for monochrome video signals, the use of comb filters to remove some aliasing components from the decoded video signal enables 'sub-Nyquist' sampling frequencies to be employed with only a slight degradation in picture quality;[1] this technique was not examined in the tests described in this paper.

‡ Codec is an abbreviation for a system with one analogue-to-digital convertor (coder) followed by a digital-to-analogue convertor (decoder).

2.2.2 *Use of dither signals to reduce visibility of quantizing errors*

For a given number of bits per sample, the coherent patterning effects such as contouring and beat patterning are more visible than random errors and therefore the picture impairment produced by quantization is reduced if the errors can be transformed from coherent to random form. One method of achieving this effect is deliberately to add a 'dither' signal to the video signal at the input of the digital coder.[3,5,6,7]

A dither signal can take various forms, one being random Gaussian noise, and its purpose is to cause the video signal to cross quantum levels more frequently than would otherwise be the case. From tests on Gaussian dither signals, it was found that the noise present in the original video signal is sufficient by itself to eliminate coherent quantizing errors if the original unweighted signal/noise ratio is lower than about $(6n+5)$ dB where n is the number of bits per sample. This expression applies when the composite colour signal just fills the conversion range in the digital coder as shown in Fig. 1. (For video signal/noise ratios given in this paper, the signal magnitude is the difference in voltage between black and white levels and the noise magnitude is its r.m.s. value.)

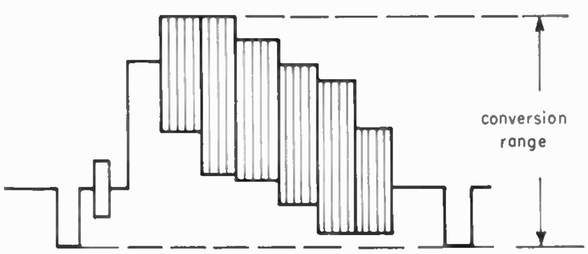
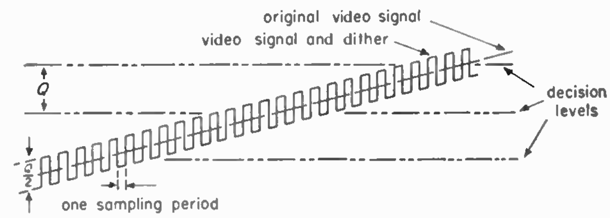


Fig. 1. Adjustment of video levels relative to the conversion range of the p.c.m. coder shown for 100% colour bar signal.

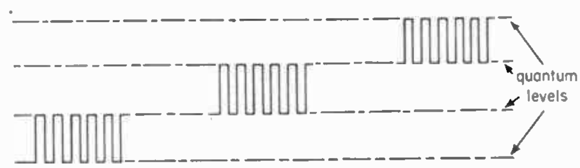
Using the number of bits per sample likely to be required for broadcast television (see Section 5), the unweighted signal/noise ratios of most present-day picture sources are, indeed, lower than $(6n+5)$ dB; the main exceptions to this rule occur during a fade or with electronically generated test signals. Thus while dither is necessary on some critical pictures, its only effect on the majority of broadcast programmes would be to slightly increase the random noise level. Therefore, to obtain maximum overall advantage by the use of this artifice, the degradation introduced when dither is not essential should be as small as possible.

Various improved forms of dither signal can be devised.

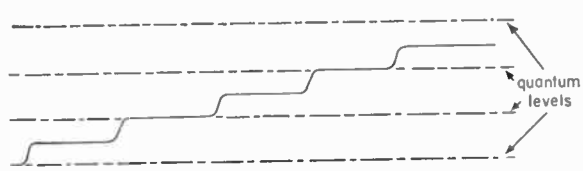
A half-sampling frequency signal with an amplitude of $Q/2$, where Q is the difference in voltage between adjacent quantum levels, is an extremely useful dither signal as it has the effect of apparently doubling the number of quantum levels in plain areas of a picture while causing negligible degradation in detailed areas. The reason for the doubling of the apparent number of levels is illustrated in Fig. 2. From this figure it can be seen



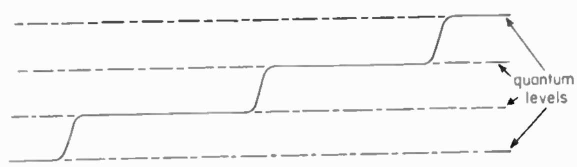
(a) Relative positions of video signal and decision levels in coder



(b) Output of decoder before filtering



(c) Output of decoder after filtering using dither



(d) Output of decoder after filtering with no dither.

Fig. 2. Effect of adding half sampling frequency dither to slowly changing video signal

that as long as the original video signal remains within $\pm Q/4$ of a decision level (see Fig. 2(a)), alternate samples of the resulting quantized signal lie on adjacent quantum levels (see Fig. 2(b)). In these circumstances the quantized signal has a mean level lying half-way between quantum levels; moreover, the frequency of alternation between quantum levels is at half sampling frequency which is above the cut-off frequency of the video low-pass filter following the decoder. As a result, in plain areas of a picture, the output of the decoder after filtering (see Fig. 2(c)) has twice the number of levels obtained without this form of dither (see Fig. 2(d)).

In principle, coherent patterns such as contouring can be removed with no further degradation at all in the signal/noise ratio by adding a pseudo-random dither signal before coding, and subtracting an identical signal after decoding.[7] It has been found, however, that by using suitable forms of dither signal, the resulting increase in noise power could be so small that subtraction is not worthwhile. For the subjective tests described in Section 4, this result was obtained by adding a dither signal consisting of a half-sampling frequency component having a peak-to-peak amplitude of $Q/2$ plus a random noise component having an r.m.s. amplitude equal to $0{\cdot}18Q$, where Q is the spacing between adjacent quantum levels. This form of dither signal causes a reduction in the overall signal/noise ratio of about 1·5 dB below that given by quantizing noise alone.

2.2.3 *Estimation of number of bits per sample from specifications for the allowable noise in analogue systems*

Since dither can cause quantizing errors to look like random noise on an unquantized picture, it is possible to deduce the number of bits per sample required for a p.c.m. system using dither, by comparing the allowable noise on analogue links with the theoretical values of the quantizing noise power produced by different numbers of bits per sample. This estimation assumes that incoherent quantizing errors having a given r.m.s. value cause the same subjective impairment on a picture as random Gaussian noise having the same r.m.s. value, an assumption which is supported by experiment.

Using the specification for a CMTT, System I, hypothetical reference circuit (CCIR, Recommendation 451-1, New Delhi 1970) to give the noise allowable in an analogue link, the weighted signal/noise ratio should not be less than 52 dB in the luminance channel or 46 dB in the chrominance channel.

Considering now the noise due to quantizing, the weighted signal/noise ratios can be obtained theoretically on the assumption that it has a uniform spectrum.[8] The results for 6, 7 and 8 bits per sample, allowing 1·5 dB degradation due to dither, are as follows:[3]

Bits per sample	S/N in luminance channel	S/N in chrominance channel
6	48·7 dB	47·2 dB
7	54·7 dB	53·2 dB
8	60·7 dB	59·3 dB
Analogue specification	52 dB	46 dB

These figures are based on the assumption that the entire video signal including synchronizing pulses and 100% saturated colours are being encoded as shown in Fig. 1. It has also been assumed that the sampling frequency is 13·3 MHz and that the noise has been measured in the frequency range 0 to 5 MHz, this being the range normally used for analogue measurements.

By comparing these quantizing noise figures with the requirements for an analogue system, it can be seen that, in the absence of instrumental imperfections, 7 bits per sample exceeds the specified requirement by 2·7 dB in the luminance channel and 7·2 dB in the chrominance channel.

Assuming that there is no correlation between the quantizing errors introduced by different codecs, the total quantizing noise obtained from 4 codecs in tandem each using 8 bits per sample would again exceed the specified requirement by 2·7 dB in the luminance channel. This result is obtained because the 6 dB reduction in quantizing noise given by the use of an extra bit per sample is exactly offset by the 6 dB increase given by four codecs instead of one.

3 P.C.M. Coding and Decoding Equipment

A block diagram of the 8-bit p.c.m. coder[9] used in the subjective tests is shown in Fig. 3. This coder has two similar stages, the first giving the four most significant bits and the second giving the four least significant bits.

In the first stage, the sampled and held video signal is fed simultaneously to 15-level comparators supplied with reference voltages equally spaced over the conversion range. The first four bits are obtained from a simple TTL logic circuit supplied with the outputs of these comparators.

To determine the lower-order bits, the first four bits are decoded to give a 16-level quantized signal which is subtracted from the original sampled-and-held signal. A second set of 15-level comparators divides this difference signal into 16 parts to give the state of the last four bits.

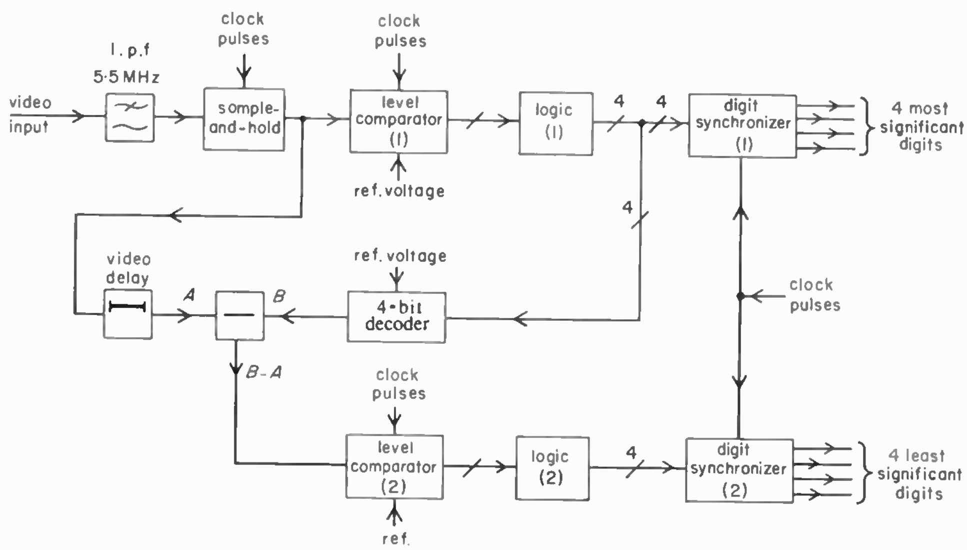


Fig. 3. Block diagram of video p.c.m. coder.

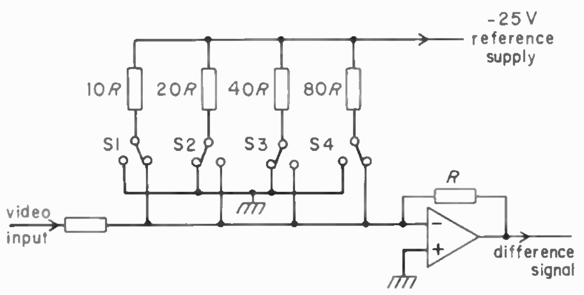


Fig. 4. Subtractor and quantized signal generator (4-bit decoder).

Finally, the digital signals are re-timed in the digit synchronizers so that all 8 digits corresponding to a given sample appear simultaneously at the output.

In order to maximize the settling time allowed in each part of the conversion process, each level comparator is followed by a D-type flip-flop supplied with clock pulses timed to occur at the end of the held portion of each sample. As a result, the speed of operation demanded from most of the circuit components is relatively slow compared with that required in most other methods of p.c.m. coding of video signals,[10,11,12,13] all the digital processing being performed by high-speed (6 ns delay per gate) TTL circuits.

The main details of the 4-bit decoder and subtraction circuits are shown in Fig. 4. The 16-level quantized signal is obtained by adding binary weighted currents corresponding to the four most significant digits. These weighted currents are supplied via diode switches S1, S2, S3 and S4 which are operated by the digital signals in such a manner that the quantized signal obtained is of opposite polarity to the unquantized video signal. The resulting output signal from the amplifier therefore represents the difference between the quantized and unquantized video signals. The input impedance to this amplifier had to be made very low compared with the resistance R to prevent significant interaction between the quantized and unquantized input signals. Additional requirements of this amplifier were that its settling time to within 0·1% of its final level should be about 30 ns or less and the d.c. drift in its output should be less than about 2 mV over its working temperature range.

In the 8-bit p.c.m. decoder used in the subjective tests,[9] the digital input signals control currents flowing into or out of the appropriate nodal points of a resistive ladder network. The output from this network gives a quantized version of the original video signal. This form of decoder has been described in more detail elsewhere.[14]

4 Subjective Tests

4.1 Experimental Arrangement

A block diagram of the equipment is shown in Fig. 5. The signals were displayed on a high-quality 19-in colour monitor having a peak brightness of 55 candelas/metre^2 (16 ft-L); with zero beam current, the brightness of the screen resulting from ambient illumination was 0·1 candela/metre^2 (0·03 ft-L). Switch S2 was used to by-pass the digital equipment when the unquantized video signal was required to be displayed. In the coder, the video signal was clamped with the tips of synchronizing pulses at the bottom of the conversion range and its amplitude adjusted so that 100% saturated subcarrier just reached the top of the conversion range (see Fig. 1).

As indicated in Fig. 5, arrangements were made to insert unquantized synchronizing pulses and colour bursts into the decoded signal; this process removed picture impairments caused by timing and clamping errors due to the quantization of this information. It should be noted that errors of this type do not cause any noticeable picture impairment when the number of bits per sample is sufficient to give broadcast quality pictures (see Section 5); they would only have affected the results of the tests for low numbers of bits per sample. If necessary, timing and clamping errors could be avoided when no unquantized signal is available by applying relatively simple processing techniques to the decoded signal.

4.2 Signal Sources

Both moving and still pictures were used in the tests, these being selected to provide both critical and non-critical picture material.

The sources for moving pictures consisted of:

(a) A colour camera which was panned slowly across a studio set. This set included a large plain white area, representing a fairly critical object for showing contouring effects, and also coloured curtains of various hues to provide large areas of high colour saturation.

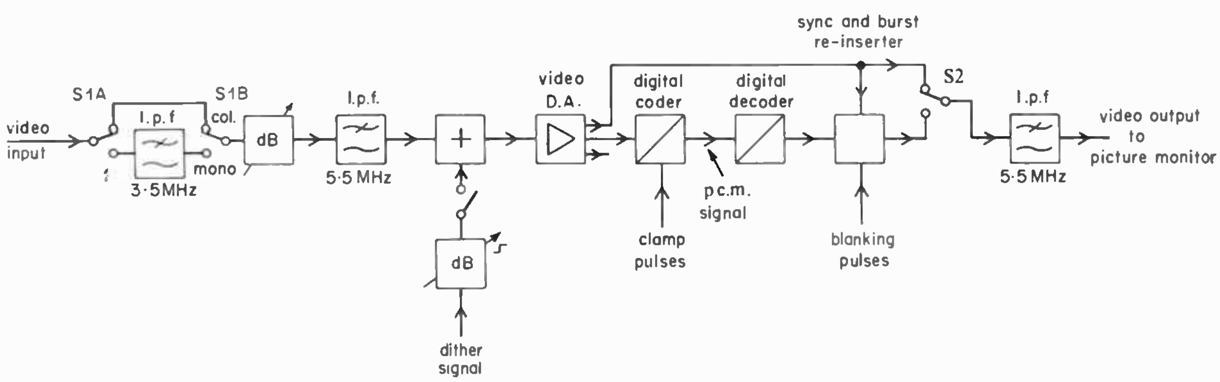


Fig. 5. Block diagram of equipment used for subjective tests.

Fig. 6. Monochrome version of slide used in tests.

(b) A high quality video tape recording of a musical variety programme.

(c) A high-quality receiver providing live pictures of an outside broadcast of a cricket match. The signal/noise ratio of the receiver alone was about 40 dB unweighted.

(d) A 16 mm colour film scanner showing a film of a musical variety programme.

The sources for still pictures consisted of:

(a) A 35 mm colour slide scanner containing a slide of which a monochrome version is shown in Fig. 6. This slide was selected as it contains relatively large plain coloured areas of low colour saturation and therefore provides a fairly critical picture for showing contouring and colour beat patterning effects.

(b) A colour bar generator adjusted to give 75% saturation. This type of signal is very critical for showing beat patterning effects.

The inherent signal/noise ratios of all sources, apart from the 16 mm film scanner and the off-air receiver, were typical of the best equipment at present available.

For some of the tests, the video signal was attenuated by 8 dB prior to the digital coder in order to obtain very critical pictures.

Monochrome pictures were obtained by inserting a 3·5 MHz low-pass filter in the video path prior to the coder. Since the tests on monochrome signals were connected only with determining the effect of varying the number of bits per sample and were not concerned with the optimum sampling frequency, this bandwidth limitation had virtually no effect on the results obtained.

4.3 Test Procedure

Each test condition was shown twice to a group of six observers seated at a distance of six times picture height from the picture monitor. The observers were all engineers or technicians having previous experience of assessing picture quality. A total of about 20 observers were used in the complete series of tests.

Picture quality was graded using the 6-point impairment scale:

Grade	Degree of Impairment
1	Imperceptible
2	Just perceptible
3	Definitely perceptible but not disturbing
4	Somewhat objectionable
5	Definitely objectionable
6	Unusable

The tests were divided into groups of 12 assessments in which the unquantized picture and quantized pictures using 4, 5, 6, 7 and 8 bits were shown twice in random order while all other test conditions remained constant. Before each group was assessed, the various types of impairment were pointed out on a 4-bit quantized picture. In addition, the unquantized picture was shown before each test picture.

The possible variants in the test conditions apart from the number of bits were as follows:

(a) Picture source.

(b) Colour or monochrome.

(c) Three different sampling frequencies, i.e. three times subcarrier frequency ($3f_{sc}$), 851 times line frequency ($851\ f_L$) and 'unlocked' as discussed in Section 2.1.

(d) Dither or no dither. Dither consisted of a half sampling frequency component plus random white noise as discussed in Section 2.2.2.

(e) Input signal to coder at normal level or attenuated by 8 dB.

It was not thought practicable or necessary to examine all possible combinations of these test conditions and

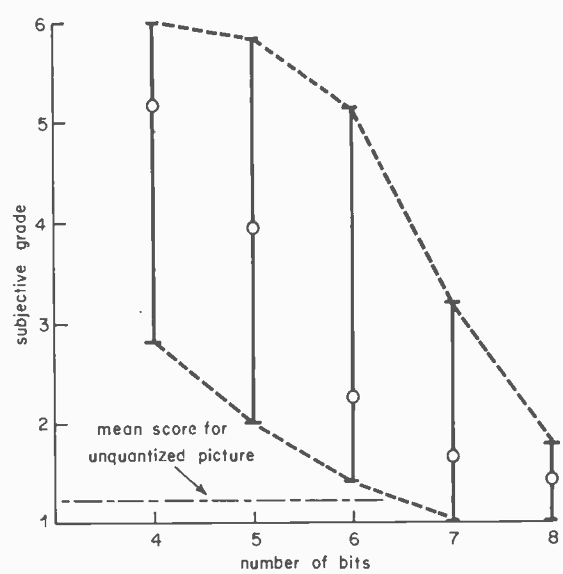


Fig. 7. Variation in subjective impairment at different numbers of bits per sample with no dither

I variation in grade for different picture sources

○ mean grade for all picture sources.

therefore a representative selection of critical and uncritical conditions was chosen for examination.[3]

5 Results of Subjective Tests

5.1 Effect of Varying the Number of Bits per Sample

Figure 7 shows a summary of the results obtained from all the tests in which no dither was used. The upper and lower curves show the limits in the mean grade given for different picture sources and different test conditions. It can be seen that the most critical pictures required 8 bits per sample to reduce the impairment below grade 2, while the least critical pictures required only 5 bits per sample for a grading of 2.

Figure 8 gives the results of the tests using dither signals. From this figure, it can be seen that there is negligible advantage to be gained by using more than 7 bits per sample with dither in use.

A comparison of Figs. 7 and 8 shows that the addition of dither signals reduces the variation in the results for different test conditions and, in general, either lowers, or has no effect on, the subjective impairment for a given number of bits, being most effective with the most critical types of picture.

Using 4 and 5 bits, the least critical pictures were rated slightly worse with dither than without. This was because the general noise added by dither produced greater impairment than the slight contours visible on these pictures without dither signals.

Figure 9 shows the difference between monochrome and colour pictures with no dither. It can be seen, with small numbers of bits per sample, monochrome pictures are in general degraded more than colour pictures; this is because colour subcarrier tends to act like a dither signal in breaking up contouring effects. The results plotted in Fig. 9 represent the mean scores obtained from five test conditions which were identical for both colour and monochrome pictures apart from a 3·5 MHz low-pass filter inserted in the video channel prior to coding for the monochrome signals.

A comparison of the results for monochrome and colour pictures showed that quantization causes a similar degradation to both types of picture when dither is in use.

The results given above apply to the subjective impairment produced by only one codec. In deciding the number of bits required for a complete broadcasting system, some allowance has to be made for the use of several codecs in tandem. With dither in use, it is reasonable to assume that a system including 4 codecs in tandem would require 8 bits per sample to give the same picture quality as one codec using 7 bits per sample (see Section 2.2.3).

With no dither in use, the overall effect of several codecs in tandem will depend on the number of bits per sample and the characteristics of the video signal. Using 8 bits per sample, most picture sources will provide sufficient inherent random noise to make the quantizing noise random in nature and therefore the overall effect will be similar to that obtained with dither in use. If, however, non-random quantizing effects such as contouring are visible, the overall effect of several codecs in tandem will depend on such factors as the relative positioning of quantizing levels in the separate codecs and it is difficult to deduce any general conclusions except that at least 8 bits per sample would be required for broadcast quality pictures.

5.2 Effect of Varying the Sampling Frequency

With dither in use, the impairment due to beats between sampling frequency and colour subcarrier was almost imperceptible for all numbers of bits examined and as

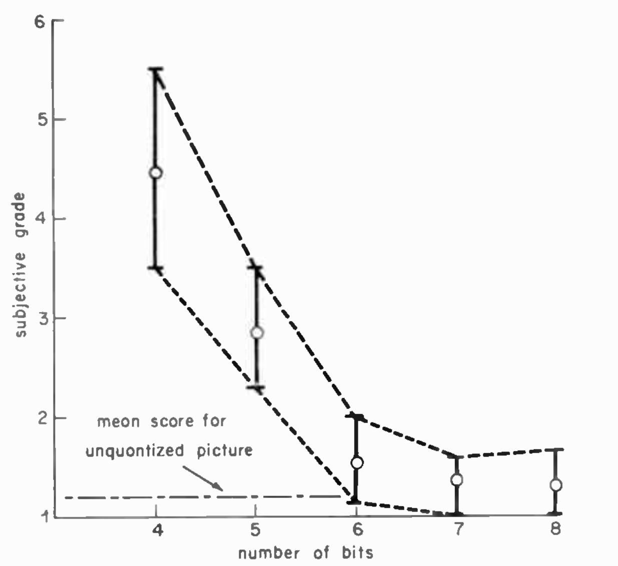


Fig. 8. Variation in subjective impairment at different numbers of bits per sample with dither in use
I variation in grade for different picture sources
○ mean grade for all picture sources.

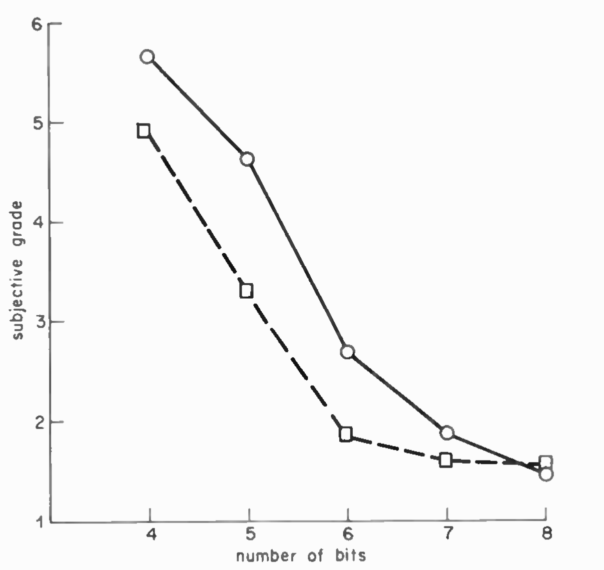


Fig. 9. Difference in subjective impairment of colour and monochrome pictures with no dither
—— mean grade for monochrome pictures
--- mean grade for colour pictures.

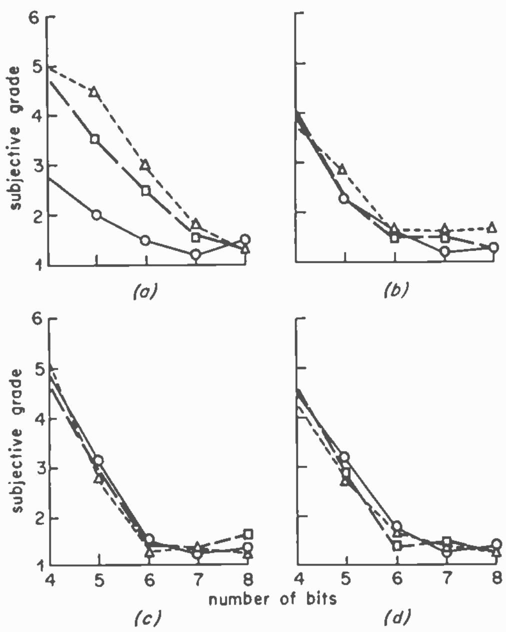


Fig. 10. Effect of different sampling frequencies on a critical picture (colour bars) and on non-critical pictures (picture from off-air receiver)

(a) Colour bars; no dither (b) Colour bars; with dither
(c) Off-air pictures; no dither (d) Off-air pictures; with dither
—— $f_s = 3 \times$ colour subcarrier frequency
—— $f_s = 851 \times$ line frequency
—— f_s unlocked.

indicated in Fig. 10 (b) and (d) there is no significant advantage in using either a locked or an unlocked sampling frequency.

With no dither, it was again found, with most types of picture, that there was little difference in the results obtained with locked or unlocked sampling frequencies. With 4 and 5 bits per sample, although there were obvious differences in the beat patterns occurring in plain coloured areas of uncritical pictures, these different effects were in general rated equally annoying. With 6 or more bits per sample it was very difficult to see the effect of changing the sampling frequency except with very critical pictures containing large uniform coloured areas such as a colour bar signal. The results for colour bars, given in Fig. 10 (a), show that there is some advantage in sampling at exactly three times colour subcarrier frequency if less than 8 bits per sample are being used without dither. Colour bars tend to be slightly misleading, however, since each coloured strip has constant luminance, hue and saturation and therefore there is no obvious degradation due to contouring effects. These somewhat non-typical conditions account for the abnormally low grades obtained using 4 and 5 bits per sample with sampling locked to 3 times subcarrier frequency.

With all types of picture, the use of a sampling frequency locked to a multiple (851) of line frequency gave very similar results to those obtained with the unlocked frequency.

Sampling at 3 times subcarrier frequency has an advantage which is not apparent from the results of the tests. Such a sampling frequency minimizes residual patterning caused by instrumental imperfections, or by overloading of the system if the video signal accidentally exceeds the conversion range of the coder, thereby causing clipping of colour subcarrier components.

6 Conclusions

The tests described in this report show that, if normal subjective standards of broadcast quality are to be maintained then not less than 7 bits per sample may be used for p.c.m. encoding of 625-line PAL or monochrome television signals. Seven bits are not adequate unless

(a) dither of a form such as that described in Section 2.2.2 is used,

and

(b) the video signal is subjected to only one coding and decoding operation.

If these conditions are not satisfied, then at least 8 bits per sample must be used. Using 8 bits per sample and dither, it is estimated that the video signal could be subjected to 4 coding and decoding operations without significant degradation.

A comparison of the theoretical signal/noise ratio of a p.c.m. system using dither with the signal/noise ratio required on analogue links (see Section 2.2.3) leads to the same conclusions regarding the number of bits per sample.

The subjective tests also indicated that for minimum picture impairment, the sampling frequency should be locked to three times colour subcarrier frequency if less than 8 bits per sample are used without dither, but need not be locked with dither in use or with 8 or more bits per sample.

Other practical considerations concerning the optimum sampling frequency are discussed in Section 2.1.

The conclusions given above are largely substantiated by the results of similar tests carried out by the RAI Laboratories in Turin.[15]

7 Acknowledgments

The author wishes to acknowledge the contributions made by other members of the BBC Research Department to the work on which this paper is based.

The paper is published by kind permission of the Director of Engineering, BBC.

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Manuscript received by the Institution on 2nd November, 1973. (*Paper No. 1595/CC206*)

STANDARD FREQUENCY TRANSMISSIONS—May 1974

(Communication from the National Physical Laboratory)

May 1974	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	0	0	−0·1	701	603·5
2	0	0	−0·1	702	603·9
3	−0·1	−0·1	−0·1	703	604·5
4	0	0	−0·1	703	604·6
5	0	0	−0·1	703	604·9
6	0	0	−0·1	703	604·8
7	0	0	−0·1	703	604·9
8	0	0	−0·1	703	604·9
9	0	0	0	703	604·8
10	0	0	−0·1	703	605·1
11	−0·1	−0·1	−0·1	704	605·7
12	0	0	−0·1	704	605·9
13	−0·1	−0·1	−0·1	705	606·5
14	0	−0·1	−0·1	705	607·0
15	0	0	−0·1	705	606·9
16	0	0	−0·1	705	606·8
17	0	0	−0·1	705	606·6
18	0	+0·1	−0·2	705	606·1
19	0	0	−0·1	705	606·3
20	−0·1	−0·1	−0·1	706	607·0
21	−0·1	0	−0·1	707	607·4
22	0	0	−0·1	707	607·1
23	+0·1	+0·1	−0·1	706	606·7
24	+0·1	+0·1	−0·1	705	605·9
25	+0·1	+0·1	−0·1	704	604·8
26	+0·2	+0·1	0	702	603·7
27	+0·1	+0·1	−0·1	701	602·9
28	0	+0·1	−0·1	701	602·1
29	0	0	−0·1	701	601·7
30	0	+0·1	−0·2	701	600·8
31	+0·2	+0·1	−0·2	699	600·3

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

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

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

Contributors to this issue*

Mr. V. G. Devereux received a degree in mechanical sciences at Cambridge in 1955. He then spent two years as a graduate apprentice with the BBC and joined the BBC Research Department in 1957 where he is still employed. He is currently a member of the Baseband Systems Section where his work is mainly concerned with the processing and transmission of digital video signals.

Dr. R. C. French (Member 1964, Graduate 1959) studied telecommunications at Norwood Technical College and after two years national service in the RAF joined Muirhead and Co to work on facsimile equipment. In 1960 he joined Mullard Research Laboratories, Redhill, to work on projects in the Communications Group. In 1973 he was awarded a Ph.D by the CNAA for work on Speech Scrambling and Synchronization techniques undertaken at MRL and Brighton Polytechnic. His current work is on data transmission in mobile radio networks.

Professor Jovan V. Surutka received the B.E. and D.E.E. degrees from the University of Belgrade, in 1947 and 1957 respectively, and from 1947 to 1951 he served as a Research Assistant at the Institute for Telecommunications, Serbian Academy of Science, Belgrade. He then joined the Faculty of Electrical Engineering at the University of Belgrade, in 1959 he was appointed an Associate Professor of Electromagnetics, and in 1968 a full Professor of Electrical Engineering. For the past seventeen years Professor Surutka has been a Consultant to the Radio-Televizija Beograd.

Dr. Dragutin M. Velickovic received Bachelor's, Master's and Doctor's degrees in electrical engineering from the University of Nish, Yugoslavia, in 1965, 1969 and 1973 respectively. Since 1965 he has been a Teaching Assistant at the Department of Electronics, University of Nish. His main research interest is in the field of linear antenna theory.

Mr. S. R. Ely studied electronics at the University of Liverpool from 1969 to 1972 and received his B.Eng. degree with first class honours in 1972. As a postgraduate student he is now engaged in experimental work both at Liverpool and at the Daresbury Laboratory and is working towards a Ph.D. thesis on very high speed data transmission.

Dr. C. C. Goodyear obtained his B.Sc. degree in physics at the University of Birmingham in 1956 and his D.Phil. at the University of Oxford in 1960. His first appointment was as Visiting Assistant Professor in the Physics Department at the University of Oklahoma and he returned to the UK in 1963 to take up an appointment in the Department of Electrical Engineering and Electronics at the University of Liverpool where he has been a Senior Lecturer since 1970. He has published papers in the fields of electrical discharges and atomic collision processes, while his current interest in telecommunications has been stimulated by teaching this subject in his present appointment.

Dr. H. J. Sherman received his B.Sc. and Ph.D. degrees in physics from London University in 1958 and 1961 respectively. From 1962 to 1963 he was a research assistant at the Imperial College of Science and Technology engaged in elementary particle physics research using bubble chamber techniques. For the next six years he was a lecturer in the Physics Department of Queen Mary College, London University, and during this time he worked on counter and spark chamber experiments in elementary particle physics. Since 1970 Dr. Sherman has worked in the Computer and Electronic Systems Division at the Daresbury Laboratory. His main interest is in the use of computer systems in physics research.

Dr. B. Zacharov received his B.Sc. and Ph.D. degrees in physics at the University of London in 1952 and 1960 respectively. He worked as a Staff Physicist at CERN, Geneva, from 1960 until 1965 and was appointed Reader in Experimental Physics at Queen Mary College, London, in 1965. He has been Head of the Computing and Electronics Group at the Daresbury Laboratory since 1966 and Visiting Professor of Physics, Queen Mary College, London, since 1968. His main interests lie in the fields of nuclear instrumentation, computers and computing.

*See also page 342.

IERE News and Commentary

New Appointments to Advisory Council on Calibration and Measurement

Mrs. Shirley Williams, Secretary of State for Prices and Consumer Protection, has made two appointments to the Advisory Council on Calibration and Measurement. Mr. Richard Foxwell, C.B.E., formerly Deputy Chairman, becomes Chairman of the Council, succeeding Sir Harold Bishop, C.B.E., who retired on 31st March 1974. His successor as Deputy Chairman is Mr. R. S. Medlock. Both appointments run initially from 1st April 1974 to 31st December 1975.

Mr. Foxwell is a Director of Marconi Instruments Ltd and was formerly Chairman of the Wayne Kerr Co. Ltd from which he retired in 1973. Mr. Medlock has been with George Kent Ltd for 39 years. He was made a Director of the company in 1956.

Three new members have recently been appointed to the Council and will serve until 31st December 1975: Mr. J. A. Edwards, B.Sc., A.R.C.S., D.I.C., C. Eng., F.I.E.E. (Technical Director of Taylor Instrument Companies (Europe) Ltd); Brigadier E. Holland, B.Sc., C.Eng., M.I.E.E. M.I.Mech.E., (Quality Assurance Manager, Decca Radar Ltd) and Professor F. W. Spiers, C.B.E., (Chairman of the British Committee on Radiation Units and Director of the Bone Dosimetry Research Unit at the University of Leeds). Further appointments are to be made with the intention of strengthening the industrial representation on the Council.

The Advisory Council on Calibration and Measurement was set up in 1966, by the then Minister of Technology. Its functions include advising on the development and operation of the British Calibration Service and it now reports to the Secretary of State for Prices and Consumer Protection. It comprises 13 members in addition to a Chairman, Deputy Chairman, Secretary, Deputy Secretary and ten Assessors; all are appointed by the Secretary of State.

The British Calibration Service (BCS) provides authenticated certification of the accuracy of measuring instruments. About 70 laboratories, each specially approved for particular types of measurement, operate within the Service.

The Council is assisted by seven technical panels of experts each dealing with a particular field of measurement. The Secretary to the Council is Mr. J. W. McHugo of the Department of Prices and Consumer Protection's Standards, Weights and Measures Division, 26 Chapter Street, London SW1.

RECMF Officers for 1974

At the recent Annual General Meeting of the Radio & Electronic Component Manufacturers Federation, the following Officers were re-elected:

Chairman: Mr Ronald A. Bulgin, M.A., C.Eng., M.I.E.R.E; Deputy Chairman: Mr. C. Robert Jennings (Companion I.E.R.E.); Treasurer: Mr. Cedric M. Benham, C.B.E.; Trustees: Mr. Richard Arbib, Mr. K. G. Smith, Dr. G. A. V. Sowter.

Presidents of the Institutions

Professor J. H. Merriman, C.B., O.B.E., D.Sc., M.Sc., F.K.C., M.Inst.P., C.Eng., F.I.E.E., who is Board Member for Technology and Senior Director, Development, of the Post Office, will take office as President of the Institution of Electrical Engineers in October.

Professor Merriman, who recently received an honorary doctorate from the University of Strathclyde, joined the Post Office Research Station in 1936 after graduating from the University of London. Until 1948 he was mainly concerned with research and after a period at Post Office Headquarters, was seconded to the Treasury from 1955 to 1959 as Deputy Director, Organisation and Methods, with particular responsibility for computer and automation policy. In 1963 he became Assistant Engineer-in-Chief and two years later Deputy Engineer-in-Chief. He was appointed Senior Director of Engineering in 1967 and when the Post Office became a Public Corporation, he was appointed to its Board as Member for Technology. In December 1968 he was appointed by the University of Strathclyde, Glasgow, as Visiting Professor in the Department of Electronic Science and Telecommunications. Professor Merriman is a past Chairman of the Electronics Division of the I.E.E. and he has represented that Institution on the National Electronics Council since 1970.

Mr. A. T. Morris, C.Eng., F.I.C.E., F.I.Mun.E., M.R.T.P.I., County Surveyor and Engineer of Oxfordshire, has been installed as President of the Institution of Municipal Engineers.

Mr. Morris began his municipal career in 1931 and during the next twenty years held appointments in Shrewsbury, Leicester, Oxford and Bath. During the war he was seconded to the Air Ministry Works Directorate and in 1953 he returned to Shrewsbury as Borough Surveyor and Planning Officer.

In 1965 Mr. Morris was appointed City Engineer and Surveyor of Oxford, and following local government reorganisation in April of this year became County Surveyor and Engineer for new Oxfordshire.

At the 64th Annual General Meeting of the Institution of Structural Engineers held in London recently, **Mr. Derek Dick, B.Sc., C.Eng., F.I.Struct.E., F.I.C.E.,** was elected President for 1974/75 to succeed Mr. L. R. Creasy in October next. Mr. Dick is a director of the W. S. Atkins Group of Epsom, one of the largest consultancy firms in the construction industry.

Dr. A. W. Davis, C.Eng., F.I.Mar.E., has been elected President of the Institute of Marine Engineers. Educated at Glasgow Academy and Glasgow University, he obtained a D.Sc. degree in engineering for contributions to marine engineering design, development and manufacture of gears for marine applications.

For five years he was Marine Mechanical Manager with Westinghouse Electric in California and he is now Technical Director for their Marine Turbine Division in Europe.

Canadian Council of Professional Engineers

Mr. J. N. Gilles Tanguay, B.A., B.Sc.A., P.Eng., the President for 1974 of the Canadian Council of Professional Engineers, ɪs a graduate of Laval Unversity, Quebec. He joined Bell Canada in 1958 and has held a number of senior engineering posts with the organization prior to his appointment as Plant Extension Engineer, Montreal Area, in 1972. The CCPE is the national co-ordinating body for the eleven provincial and territorial associations of Canadian engineers

IERE Visit to ESTEC—New Date

The Institution visit to the European Space and Technology Centre (ESTEC) at Noordwijk in the Netherlands planned for last May had to be postponed owing to unforeseen circumstances. A new date has now been arranged—Thursday, 26th September—and the programme will be substantially as planned for the original visit, details of which were given on page 173 of the March 1974 issue of the Journal. Members in the British Isles or Europe who are interested in taking part in this visit are urged to get in touch with Mr. R. C. Slater, Deputy Secretary, IERE, 8–9 Bedford Square, London WC1B 3RG as quickly as possible.

National Engineering Laboratory Director

Mr. D. H. Mallinson, Director-General Engines in the Ministry of Defence Procurement Executive, will become Director of the Department of Industry's National Engineering Laboratory at East Kilbride on 1st September. He succeeds Mr. R. H. Weir, who is retiring from public service.

The National Engineering Laboratory undertakes research and development on a wide range of mechanical engineering problems. It aims to establish more efficient design procedures and new or improved manufacturing processes. By promoting rapid introduction of the results of its work it strives to enhance profitability and competitiveness of British industry.

Institution Correspondence and Address Changes

Letters to the IERE headquarters can be dealt with more easily if members quote in all correspondence both membership grade and registration number.

If a change of address is being notified, please remember that Journals are despatched according to computer file, and that the Institution's Master File is up-dated once a month. Therefore, if at least five weeks' notice of a change of address cannot be given, please make arrangements for forwarding of mail. Overseas members will be aware that delivery of Journals can take several weeks.

Leeds Electronics Exhibition

At the 11th Leeds Electronics Exhibition held in the Department of Electrical and Electronic Engineering at the University of Leeds from 25th-27th June, about 160 firms from all parts of the United Kingdom took part and the attendance was in the region of 11,000. The emphasis of the exhibits was on instrumentation and control rather than on communications. However, since the latter formed the theme of a commercially organized exhibition and conference held a month earlier in Brighton, the reason for this response is understandable and indeed practical from the point of view of the electronics industry.

Traditionally, the Leeds exhibition has been supported by a concurrent series of lectures sponsored in alternate years by the Institution of Electronic and Radio Engineers and the Institution of Electrical Engineers. This year, the North Midland Centre of the IEE organized the seminars at which eight lectures, dealing mainly with measurement techniques and with industrial applications, were presented.

The exhibition was opened by His Royal Highness the Duke of Kent, who is Deputy Chairman of the National Electronics Council and a member of the Council of the IERE. He was welcomed by Professor Geoffrey Carter, head of the department which organizes and accommodates the exhibition.

Replying to Professor Carter's welcome, the Duke referred to the great strides which had been made in the 11 years the exhibition had been held. This was proof of the wisdom of the Department's staff in foreseeing the need for a major electronics exhibition in the North of England. He continued:

'The idea of holding a technical exhibition in a University Department, attractive to industry yet commercially neutral, is a fairly unusual one but I believe extremely valuable. For one thing it can only help to remove that mutual mistrust which still I fear persists in some quarters between the University researcher and engineers in industry—the former being accused of working in a vacuum and without regard for what is commercially practical, and the latter suspected of not being interested in research for its own sake but eager to seize other people's ideas if they show signs of becoming profitable. Any move to bring the university graduate closer to his more market-oriented counterpart in industry can only be welcomed. Contacts established during an event like this should help to achieve that aim.

'The other reason that I believe an exhibition in this particular environment is important is that it provides an essential means of exchanging ideas and of keeping up to date. Such is the pace of development in electronics, above all other technologies, that no one, and probably no single company can hope to stay abreast of all activities across the whole spectrum of this science—so that managers, researchers, salesmen and accountants all need opportunities at fairly frequent intervals to escape from the office or the laboratory and see what is going on "down the road"—or maybe even across the corridor! This cross-fertilization and brain-picking is absolutely vital if we are to avoid the fragmentation and isolation of different disciplines which are such a danger in modern technology.

'I find it is highly encouraging that so many firms are exhibiting here at Leeds, and particularly so many small ones. A few of them, I understand, manufacture a single product and some are unique in their own particular fields—taken together they represent a formidable body of expertise and flexibility incidentally of export potential—I wish them every success and hope this exhibition will have been a profitable exercise for them. I am also impressed by the idea of holding a seminar in parallel with the exhibition. This is, I know, a well-tried concept and I am certain it is a thoroughly sound one—and what better place for such a series of lectures than at a great University.

'Those who have had the courage and foresight to arrange these annual exhibitions and those who now regularly use them to display their goods deserve the fullest encouragement and support. A vigorous, confident and forward-looking electronics capability is vital for this country, for it is going to be the key technology of the future and already the impact of electronics on other sections of industry as well as on defence, transportation, medicine and in very many different fields is enormous and out of all proportion to the value of the products or the number of people engaged in making them. Exhibitions of this kind have a great part to play in helping to promote and sustain that capability.'

Members' Appointments

CORPORATE MEMBERS

Group Capt. R. Morris, O.B.E., M.A., RAF (Fellow 1966, Member 1953, Associate 1945) has been appointed Assistant Director of Signals Policy on the Central Staff of the Ministry of Defence. He was previously head of the Maintenance Data Centre at RAF Swanton Morley.

Mr. S. J. H. Stevens, B.Sc.(Eng.) (Fellow 1964, Member 1952) has been appointed Director of Aircraft Quality Assurance, Ministry of Defence Procurement Executive. Mr. Stevens was previously an Assistant Director in the Electrical Quality Assurance Directorate. Since entering Government service in 1949 after some 15 years' experience in industry, he has held various appointments in Government Production and Inspection Departments and from 1958 to 1961 he was attached to SHAPE as Production Specialist on the Tropospheric Scatter Communications System.

Mr. Stevens served on the Institution's Technical Committee from 1952 to 1956, on the Examinations Committee from 1963 to 1968 and he has been a member of the Membership Committee since 1968. Between 1957 and 1958 he was a member of the Council.

Lt. Cdr. S. M. Bruce, RN (Member 1969, Graduate 1962) has been posted to HMS *Leander* as ADA Weapon Systems Officer. He was previously Section Officer in the Avionics Systems Training Group in HMS *Daedalus*.

Mr. P. G. Evans (Member 1967, Graduate 1963) will take up the appointment of Lecturer in Aeronautical Engineering at the W. R. Tuson College, Preston, in September. After retiring from the RAF in the rank of Squadron Leader in 1971, Mr. Evans was with the Department of Aeronautical Engineering at Southall College of Technology.

Mr. I. Frogley (Member 1974, Graduate 1967), who has been with the Marconi Research Division since 1967, has joined Plessey Avionics and Communications as Project Administration Manager for *Ptarmigan* SCRA.

Mr. Abdul Hamid (Member 1973, Graduate 1971) has joined the Libyan Broadcasting and Television Organisation as Television Transmitter Engineer at the Derna television station. He was previously Senior Engineer with Radio Pakistan, Quetta.

Mr. B. R. Heasman (Member 1973, Graduate 1971), who was a Development Engineer with Edwards Instruments, Eastbourne, has joined the IBM Research Laboratories, Winchester, as an Associate Engineer.

Mr. C. P. C. Heightman (Member 1971, Associate 1969) has returned to the Plessey Company as a Design Assurance Review Engineer in the *Ptarmigan* Project Office at Roke Manor, Romsey, after a period at Camberley with EASAMS, where he was a member of the Central Design Management Team on the MRCA Project. Until 1972 he was with Plessey at the Braxted Park Laboratory as a Principal Electronic Development Engineer.

Mr. P. J. Marchant (Member 1973, Graduate 1967), who has been with Independent Television News since 1969, latterly as Senior Engineer, has been appointed Supervisory Engineer in charge of Vision Maintenance.

Flt. Lt. M. A. Nixon, RAF (Member 1972, Graduate 1970) has been appointed Officer Commanding Electrical Engineering Squadron at RAF Leeming. For the past two years he was Engineering Officer attached to the Air Staff Division of the Royal Malaysian Air Force.

Mr. B. O'Brien (Member 1966) who joined the Educational Technology Unit of the University of Witwatersrand, Johannesburg, a year ago, has been appointed Deputy Chief Engineer, Rhodesia Broadcasting Corporation. Before joining Uganda Television in 1967, Mr. O'Brien was with the Independent Television Authority.

Sqn. Ldr. D. R. West, RAF (Member 1970) has been appointed Senior Engineer Officer at RAF Masirah. He has just completed a two-year tour of duty at the Aeroplane and Armament Experimental Establishment, Boscombe Down.

Wing Cdr. A. G. Wilmot, RAF (Member 1968) has been appointed to the Headquarters of RAF Strike Command as Electrical Engineer (Telecomms.)

NON-CORPORATE MEMBERS

Mr. A. Broadbent (Graduate 1968), previously a Development Engineer with Telefunken Computer at Konstanz, has moved to Litef, Freiburg, West Germany, in a similar capacity.

Mr. J. Bushby (Graduate 1968), who is with the International Telecommunications Union, has been appointed Head of the Switching Section at the Nigeria National Posts and Telecommunications Training Centre under the ITU/United Nations Technical Aid programme. He was previously Telecommunications Training Officer for the Ministry of Posts and Telecommunications in Lesotho.

Mr. O. O. Gbeminiyi, M.Sc. (Graduate 1969) who has been Experimental Officer in the Department of Electronic & Electrical Engineering at the University of Ife, has been appointed to a Lectureship in Electronic Engineering in the School of Engineering at Midwest Polytechnic, Auchi, Nigeria.

Mr. P. L. Haworth (Graduate 1974) has joined Geophysical Services Ltd, and is concerned with the operation and maintenance of seismic instruments. He was previously a Development Engineer with Marconi-Elliott Avionics at Rochester.

Mr E. J. Hooper, M.B.A. (Graduate 1970) has been appointed Senior Lecturer in Management and Computing at Blackburn College of Technology and Design. He was previously Assistant General Manager, B. Rhodes & Son Limited.

Mr. H. C. Parker (Graduate 1969) who has been with Avimo Ltd as a Senior Electronic Design Engineer, is now Chief Designer at the Taunton works of the Company.

Mr. S. J. Piper (Graduate 1973) is now Computer Field Engineer with Rank Xerox Data Systems Limited.

Mr. B. B. Streeter (Graduate 1972) who is with Diablo Systems Limited, has joined the Company's Brussels Office as Product Manager, European Operations.

Mr. D. B. Webb, B.Sc. (Graduate 1967) who was previously a Principal Engineer with Marconi Space and Defence Systems, has joined the Scientific Civil Service and is now a Senior Scientific Officer at the Royal Radar Establishment.

Mr. S. P. Wong, B.Sc. (Graduate 1967) is now Public Lighting Engineer with China Light & Power Co. Ltd, Hong Kong. He was previously Telecommunications Planning Engineer with the Transmission and Distribution Department of the Company.

Mr. M. J. Valliant (Associate Member 1974) has been appointed Professional and Technology Officer III at the Hartland Geomagnetic Observatory of the Institute of Geological Sciences.

Mr. G. H. Bates (Associate 1961) has joined the International Civil Aviation Organisation Headquarters at Montreal as a Simulator Engineering Consultant. He previously held a similar appointment at the Civil Aviation Safety Centre in Beirut.

Forthcoming Institution Meetings

London Meetings

Wednesday, 25th September

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

Colloquium on RADAR FOR VEHICLE GUIDANCE

IERE Lecture Room, 2 p.m.

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

Thursday, 3rd October

ANNUAL GENERAL MEETING

London School of Hygiene and Tropical Medicine, Keppel Street, London WC1, 6 p.m. (Tea 5.30 p.m.)

Wednesday, 9th October

COMPONENTS AND CIRCUITS GROUP

Colloquium on H.F. HEATING CIRCUITS AND TECHNIQUES

IERE Lecture Room, 10 a.m.

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

Wednesday, 16th October

JOINT MEETING WITH IEE
EDUCATION AND TRAINING GROUP

The Technician Education Council

By F. Fidgeon (*Technician Education Council*)

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

Thursday, 17th October

JOINT IEE/IERE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP

Colloquium on ELECTRONICS IN AUDIOLOGY

IERE Lecture Room, 10 a.m.

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

Wednesday, 23rd October

AUTOMATION AND CONTROL SYSTEMS GROUP

Colloquium on IMPLEMENTATION AND RELIABILITY OF AUTOMATED SYSTEMS

IERE Lecture Room, 10 a.m.

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

Tuesday, 29th October

JOINT MEETING OF COMMUNICATIONS GROUP AND AUTOMATION AND CONTROL SYSTEMS GROUP

Colloquium on SIGNAL PROCESSING IN COMMUNICATIONS SYSTEMS

IERE Lecture Room, 10 a.m.

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

Kent Section

Wednesday, 25th September

Flight Simulation

By G. R. Wilson (*Redifon Flight Simulators*)

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

Thursday, 17th October

Modern Colour Television Systems

Speaker to be announced

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

East Anglian Section

Thursday, 10th October

ANNUAL GENERAL MEETING

The Saracen's Head Hotel, High Street, Chelmsford, 7 p.m.

Thames Valley Section

Wednesday, 16th October

Colour Television

By A. C. Maine (*IoW Technical College*)

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

This lecture will review the progress of colour television from the early Baird experiments to the present day quality transmissions. After an outline of the principles involved, the advances made in receiver design will be examined, with particular reference to the various types of display tube and electronic devices available. The lecture will be illustrated by slides and demonstrations.

South Western Section

Tuesday, 8th October

JOINT MEETING WITH IEE

Seminar on ADVANCES IN TELECOMMUNICATIONS

The Seminar will begin with an introductory lecture at 6 p.m. followed by a buffet dinner in the Senior Common Room at the University of Bath at 7.15 p.m. and finally by a discussion with a panel of experts from 8.15 p.m.

There will be a charge of £1.00 per head for dinner, payable in advance. Tickets can be obtained from Mr. B. Bolton, School of Electrical Engineering, University of Bath, Claverton Down, Bath. Payment by cheque would be preferred and these should be made payable to B. Bolton. The closing date for dinner bookings will be Wednesday, 2nd October 1974.

The Senior Common Room Bar will be open for those who do not wish to dine.

It should be stressed that these seminars are not intended to be research colloquia. The Committee hopes that members who are new to the field will look upon this as an opportunity to find out about advances in telecommunications and that members who are more familiar with the subject will take the opportunity to share their experience and to put questions to the panel.

It may be necessary to place a limit on the numbers attending the dinner and early booking is advisable.

Wednesday, 23rd October

JOINT MEETING WITH IEE

The Digital Data Network

By M. Foulkes

Westinghouse Canteen, Chippenham, 6 p.m. (Tea 5.30 p.m.)

West Midland Section

Wednesday, 2nd October

JOINT MEETING WITH R.AE.S.

Redundancy in Aviation Systems

By R. K. Barltrop (*Marconi-Elliott*)

R.A.F. Cosford, 7.15 p.m. (Tea 6.30 p.m.)

East Midland Section

Thursday, 17th October

Digital Differential Analysers and Analogue Computers

By W. Forsythe (*Loughborough University*)
Leicester University, 7 p.m. (Tea 6.30 p.m.)

Southern Section

Wednesday, 16th October

JOINT MEETING WITH IEE

Chairman's Address—Gigawatt Power Transmission

By Dr. B. M. Weedy (*Southampton University*)

Lanchester Theatre, Southampton University, 6.30 p.m. (Tea from 5.45 p.m. in Senior Common Room)

Thursday, 17th October

Underwater Acoustic Imaging

By S. O. Harrold (*Portsmouth Polytechnic*)

South Dorset Technical College, Weymouth, 6.30 p.m.

Thursday, 24th October

JOINT MEETING WITH IEE
Automatic Weather Stations

By H. R. S. Page (*Plessey Radar*)

Farnborough Technical College, 7 p.m. (Coffee and biscuits available in Refectory from 6.30 p.m.)

The Thames Valley Section reports—

How one Local Section serves its members' needs

The principal activity of most IERE Local Sections is the presentation of a regular programme of technical meetings and the Thames Valley Section has been fortunate in attracting audiences of more than 100 to many of its evening meetings at Reading University. In addition to this work the Section Committee has on several occasions discussed other ways of interesting members in the activities of the Institution. Two meetings held at a local hostelry to discuss the organization and future of the Institution were well attended and revealed a latent interest in its future.

In spite of the difficulties through which the country has been passing during the 1973–74 Session, the Section has continued to flourish and expand its service to members. During the session there have been six technical meetings (the January one being cancelled) with an average attendance of about one hundred (a record for any session in the history of the Section). Moreover Professor Fellgett's lecture on 'Ambisonic Sound', held jointly with the IEE, attracted the incredibly large record audience of 184, exceeding the previous record (achieved at the previous meeting) by over forty.

Before settling on further action the Local Section Committee decided to carry out a survey of members living within the area of the Thames Valley Section, so as to give them an opportunity to comment on its activities. The questionnaire was returned by more than 200 members, including 67 graduates, but only five students.

Although the meetings at Reading University are well attended, half of the replies were from members who had never attended one of these meetings. Some were too busy, while others who worked in London found it more convenient to attend the meetings at Bedford Square. Some 40 replies were from members who found Reading inaccessible, though since the survey was made, travelling may have become less of a problem for many of those who live around High Wycombe and Maidenhead due to the opening of new trunk roads.

More than 75% of members live in the relatively small part of the Thames Valley between London and the Chilterns. The remainder are spread over a much wider area to the west and north of the Chilterns, extending from Wantage to Bletchley, a distance of some 40 miles. These members are scattered and isolated from Reading, the centre of activity.

The replies gave almost unanimous support for the less formal gatherings, where there is an opportunity to question senior members of the Institution and discuss their replies.

Technical visits are favoured, but there remains the problem of a very wide diversity of interests.

As a result of the survey a small sub-committee has been set up to establish contact with Graduate members to help them with their enquiries about transfer to Chartered Engineer status. In addition the Section Committee is examining ways of overcoming the isolation of members to the North of the Chilterns.

There have been many offers to display posters and the increased publicity that this will provide should help the trend towards the involvement of members in the sectional activities and in particular arouse the interest of potential student members.

The achievements of the Section can probably be attributed to various causes. The Local Section Committee is undoubtedly the most enthusiastic and active committee to hold office in the history of the Section. It is also a very experienced committee and has selected the correct balance in the programme of meetings to attract large and professional audiences. Also the venue with its adequate parking facilities and excellent lecture theatre provides the ideal background for meetings.

The future policy of the Committee is that of providing maximum service for the membership in terms of technical meetings at a professional level giving coverage to both 'state of the art' technology and new developments in popular consumer fields such as hi-fi and colour television.

P. Atkinson
Chairman,
Thames Valley Section.

Forthcoming Institution Meetings *(cont.)*

Yorkshire Section

Thursday, 19th September

Custom v. Standard, M.O.S. and Bipolar Integrated Circuits

By T. Everest (*ITT*)

Leeds Polytechnic, 6.30 p.m.
(Refreshments 6 p.m.)

North Western Section

Thursday, 17th October

Current Trends in Semiconductors

By Dr. K. J. Dean (*SELTEC*)

Bolton Institute of Technology, 6.15 p.m.
(Tea 5.45 p.m.)

South Wales Section

Wednesday, 9th October

Charge Coupled Devices

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

Department of Applied Physics and Electronics, U.W.I.S.T., Cardiff, 6.30 p.m.
(Tea 5.30 p.m.)

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

Wednesday, 23rd October

JOINT MEETING WITH IEE

What are the Wild Waves Saying?—an early History of Radio Detection

By V. J. Phillips (*University College of Swansea*)

University College of Swansea, 6.30 p.m.
(Tea 5.30 p.m.)

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 28th December 1973, 4th and 20th June 1974 recommended to the Council the election and transfer of 47 candidates to Corporate Membership of the Institution and the election and transfer of 22 candidates to Graduateship and Associateship. In accordance with Bye-law 23, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

Meeting: 28th December 1973 (Membership Approval List No. 186)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Fellow

BRIGHT, John Alexander Sydney. *Weybridge, Surrey.*

Transfer from Graduate to Member

BIRKBY, Arthur James. *Knaphill, Surrey.*
CARTER, David John Robert. *Henley-on-Thames, Oxfordshire.*
CLARKE, Ernest Trevor. *Eastleigh, Hampshire.*
CROWTHER, Nigel Joshua. *Moston, Manchester.*
DAS, Hironmoy, B.Sc. *Morden, Surrey.*
DAVIES, Bryan. *Stoke-on-Trent, Staffordshire.*
EVANS, Denis. *Neath, Glamorgan.*
FINDLAY, James Stuart. *Marlow, Buckinghamshire.*
GHOSH, Asish, B.Sc. *Knutsford, Cheshire.*
GORVETT, Leonard. *Oldham, Lancashire.*
HAWES, Robert Ian. *Carshalton, Surrey.*
JENKINS, William Richard Scott, Lieutenant R.N., B.Sc. *Hamble, Southampton, Hampshire.*
KIRKHAM, Richard Joseph. *Caterham, Surrey.*
MUNRO, Ian. *Aberdeen.*
OAKES, Patrick Andrew Benjamin. *Royston, Hertfordshire.*
PARKER, Henry Charles, Lieutenant R.N. *Low Ham, Langport, Somerset.*
PAUL, Samuel Leslie Rankin. *Leatherhead, Surrey.*
POLLOCK, Ian, Charles Burns. *Kings Somborne, Hampshire.*
POWELL, Victor Cecil. *Bolton, Lancashire.*
PRESTON, John Richard. *Fingland, Manchester.*
RICHARDSON, John Derek. *Ascot, Berkshire.*
SAYERS, James Francis. *Glengormley, County Antrim, Northern Ireland.*
SHEEHAN, Michael. *Edgware, Middlesex.*
SIMPSON, Roger John. *Buntingford, Hertfordshire.*
SMART, Peter Britton. *Isle of Wight.*
SNEDDON, John Leishman. *Midlothian, Scotland.*
STEVENS, John Norman. *Fleet, Hampshire.*
STRIPPLE, Robert Phillip. *Romford, Essex.*
VUCEVIC, Svetozar. *Chelmsford, Essex.*
WALLER, George. *Cheriton, Alresford, Hampshire.*
WESTCOTT, John Jennings. Flight Lieutenant, R.A.F. *Salisbury, Wiltshire.*
WHITBY, Anthony Maurice. *Baldock, Hertfordshire.*
ZOTOV, Nikolai Victorovich, Flight Lieutenant. *Dereham, Norfolk.*

Direct Election to Member

ANDERSON, Anthony John, Squadron Leader, R.A.F. *Maidenhead, Berkshire.*
DAVIS, Alan Huw, B.Sc. *Penclawdd, Swansea.*
MOFFATT, Harry St. Clair. *Dunbartonshire, Scotland.*
NICOL, Peter John Patrick. *Hoddesdon, Hertfordshire.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

AJILOGBA, Adetayo Alamu. *Lagos, Nigeria.*
CALEY, Geoffrey Victor. *Antigua, West Indies.*
DE, Subrata Kumar. *Dhanbad, Bihar, India.*
EWINGTON, Paul Anthony. *Ontario, Canada.*
LOWING, John Edward. *Glenbrook, New South Wales, Australia.*
NWANKWO, Godfrey Irodinaulo. *Lagos, Nigeria.*
QUINN, Peter Felix. *Victoria, Australia.*
SOIN, Sarvnandan Singh. *Singapore 20.*

Meeting: 4th June 1974 (Membership Approval List No. 187)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBER

Transfer from Student to Member

BUSBY, John Lawrence, B.Sc. *Crawley, Sussex.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

WALTERS, Philip William, B.Sc. *London, N.2.*

Direct Election to Graduate

STEWART, Peter Jackson. *Southampton, Hampshire.*
TYLER, Clifford Ronald, B.Sc. *Penarth, Glamorgan.*

Transfer from Graduate to Associate Member

ELSON, David. *Warrington, Lancashire.*

Direct Election to Associate Member

AUSTIN, Anthony John. *Sleaford, Lincolnshire.*
KHAN, Akber Ali. *Newcastle upon Tyne.*
WILLIAMS, Geoffrey Michael. *London, E.4.*

STUDENT REGISTERED

LAWLER, Richard Kieran. *Beaumont, Dublin 9.*

OVERSEAS

NON-CORPORATE MEMBERS

Direct Election to Associate Member

ENEZE, Angus Ozoemena. *Ibadan, Nigeria.*
MA, Man Kwong. *Kowloon, Hong Kong.*
RAJAH, T. Thanga. *Kuala Lumpur, Malaysia.*
TAN, Eng Seng. *Singapore 12.*

Direct Election to Associate

TREMLETT, Lewis Reginald. *Ontario, Canada.*

STUDENT REGISTERED

TAN, Lee Tee. *Singapore 3.*

Meeting: 20th June 1974 (Membership Approval List No. 188)

GREAT BRITAIN AND IRELAND

NON-CORPORATE MEMBERS

Direct Election to Graduate

PATEL, Bhasker, B.Sc. *London, S.W.8.*
TSIFTSIS, Constantine Demetre, B.Sc. *Leeds 7.*

Transfer from Graduate to Associate Member

DOMICAN, Joseph. *Plymouth, South Devon.*

Direct Election to Associate Member

PARSONS, Terence. Swanage, Dorset.

Direct Election to Associate

LEE, Archibald Leonard, B.A. *Stockton-on-the-Forest, York.*

STUDENT REGISTERED

ILLINGWORTH, Graham. Potton, Bedfordshire.

OVERSEAS

NON-CORPORATE MEMBERS

Direct Election to Graduate

HINCHCLIFFE, John David, B.Sc. *Manila, Philippine Islands.*
VISWAMBHARAN, K. *Trivandrum, Kerala State, India.*

Direct Election to Associate Member

AL-KINDY, Nassir Issa. *Lusaka, Zambia.*
LEE, Ah Yew. *Kuala Lumpur, Selangor, Malaysia.*

Direct Election to Associate

GOHAR, Nazir Ahmed, Lieutenant P.N. *Karachi, Pakistan.*

STUDENTS REGISTERED

KAN YEW TUCK, Patrick. *Jalan Bukit Merah, Singapore 3.*
KOH, Kian Kok. *Singapore 3.*
LAW, Wai Kuen. *Hunghom, Kowloon, Hong Kong.*
LIM, Seng Fatt. *Singapore 8.*
MUTHUKUMAR, Ratnaswamy. *Colombo 6, Sri Lanka.*
TOYE, Dele Olawoye. *Western State, Nigeria.*