FOUNDED 1925
INCORPORATED BY
ROYAL CHARTER 1961

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

THE RADIO AND ELECTRONIC ENGINEER

The Journal of the Institution of Electronic and Radio Engineers

VOLUME 40 No. 1 JULY 1970

A Policy for Planning Research

THE WISDOM of a Solomon, let alone a government department, board of directors or university committee, would be sorely taxed in trying to solve the problem of allocation of resources, both of money and of manpower, to research. Competing calls of rival factions have to be pondered upon and inevitably there has to be a policy of selectivity and concentration. This is the theme of a recent pamphlet issued by the Science Research Council of the British Government's Department of Education and Science* and it first shows, briefly, what the Council has done in supporting research in major fields of science and engineering. On the strength of this experience a list of principles has been drawn up which it proposes to follow in exercising its policy of greater concentration and selectivity.

In surveying over the past two years the achievement of its main objectives of creating scientific and technical assets, as represented by the training of highly skilled manpower and the support of work of promise, the Council has come to certain conclusions: selectivity must be applied both in choosing fields of science on which to concentrate and in choosing the scientists to do this research and the laboratories in which they will work. This last criterion in particular implies that more than average support will be given to certain university departments who will be thus selected on the basis of their leadership, past achievement, present expertise, or factors such as ability to collaborate with industry. Further, concentration of resources will be planned by shifting from less promising areas. It is however emphasized at several points in the pamphlet that it will be an essential part of S.R.C. policy to keep some support always available to any outstanding individual in any part of any subject for work of sufficient 'timeliness and promise'. Similarly the pattern of preferred topics and places will be continuously reviewed and just as programmes will be discontinued or run down as they reach completion or lose impetus, so new centres will be supported if the university itself shows sufficient initiative. The Council's Boards will have to keep continually aware of the risk of failing to support an outstanding piece of research because it originates in a minor area and, conversely, of looking with greater benevolence on perhaps less outstanding work which has arisen within one of the large centres.

Areas in which selectivity and concentration have already been evident include computing science—at Cambridge, Edinburgh and Manchester—and control engineering—at Cambridge, Imperial College and Manchester. Special problems arise in some other areas where large capital installations are necessary, for instance astronomy, space and nuclear physics, and consideration has then to be given to creation of regional or national facilities or participation in international organizations. A larger measure of collaboration between the universities and S.R.C. laboratories is implicit in the planning of such areas and it is acknowledged that this greater influence of the S.R.C. over university research calls for especial care in fully discussing the proposals before decisions are taken.

In his foreword, the Chairman of the S.R.C., Sir Brian Flowers, F.R.S., stresses that the intention of the pamphlet is to encourage just such wide discussion of the basic policy and he specifically invites constructive criticisms. These are attitudes warmly to be applauded and could with advantage be adopted in many other places.

F.W.S.

* 'Selectivity and Concentration in support of Research'. Published by the S.R.C., State House, High Holborn, London, WC1R 4TA.

INSTITUTION NOTICES

Birthday Honours List

Congratulations on behalf of the Council have been sent to the following members of the Institution whose appointments to the Most Excellent Order of the British Empire were announced in Her Majesty's Birthday Honours List published on 13th June 1970.

OFFICER OF THE MILITARY DIVISION (O.B.E.):

Wing Commander Ronald Herbert Smith, R.A.F. (Member 1956)
(Wing Cdr. Smith is on the Electrical Engineering Staff of R.A.F. Strike Command.)

MEMBER OF THE MILITARY DIVISION (M.B.E.):

Major John Drennan, R.E.M.E. (Member 1969)
(Major Drennan is Senior Engineering Officer with 21 Joint Services Trials Unit, South Australia.)

MEMBERS OF THE CIVIL DIVISION (M.B.E.):

David Robinson Cockbaine (Member 1961)
(Mr. Cockbaine is a British Technical Assistance Officer in Turkey, attached to the Turkish Atomic Energy Commission as Electronics Adviser and Director of the A.E.C. Nuclear Electronics Laboratory.)

Douglas Herbert Allenby Scholey (Fellow 1964, Member 1951)
(Mr. Scholey recently retired from the position of Engineer in Chief of the East African Posts and Telecommunications Administration having completed over twenty-one years with the Administration; he now holds a senior appointment with the Telecommunications Group of the Plessey Company. For a number of years, Mr. Scholey was the Institution's representative in East Africa.)

Courses in Electronic Engineering

A supplement to the June 1970 issue of *The Radio and Electronic Engineer* was issued entitled 'Courses in Electronic Engineering and Allied Subjects in Universities and Colleges of Technology 1970–71'. This booklet which contains announcements by 35 British Universities, Polytechnic, Colleges of Technology and Technical Colleges gives details of university and C.N.A.A. degree courses, college diplomas, full-time courses for the C.E.I. Examination as well as a wide range of special advanced courses. Copies were sent only to members and subscribers in Great Britain, Northern Ireland and the Republic of Ireland. However, members in other countries may obtain a copy of this advertisement supplement from the Institution price 2s. 6d. post free; extra copies may be obtained at the same rate.

Dinner of Council and Committees

Members who are serving or who have served on the Institution's Council and its Committees (Standing, Group and Local Section), or as I.E.R.E. representatives on other bodies, are given this early advice that the 13th Dinner of the Council and Committees will be held at the Savoy Hotel, London, on Thursday, 5th November next. The Dinner will provide an opportunity of thanking Sir Leonard Atkinson for his services to the Institution during his term as President in 1968 and 1969.

This is traditionally an Institution function to which members may bring their ladies. Announcements will shortly appear in the *Journal* and *Proceedings* regarding the price of tickets, etc.

Institution Activities in Israel

Following the agreement reached between the Director of the Institution, Mr. Graham D. Clifford and Mr. E. Pelles, Director of the Association of Engineers and Architects of Israel, a further twenty-three Corporate Members of the I.E.R.E. have now been accepted as full Members of the Association, bringing the total number of the I.E.R.E. members in the Association to twenty-six.

Groups of members have recently visited places of interest to electronic engineers. These included the terminal station of the submarine telephone cable linking Israel to France and the Nahal Soreq Experimental Nuclear Centre.

The Section has recently published a further issue of its 'Newsletter' which has been distributed free of charge to members residing in Israel. The contents include a short description of a computing counter by Mr. Z. Glaser, Director of Monsel Electronic Instruments of Haifa, and Institution notices and news items. Members residing in other parts of the world may obtain copies of the 'Newsletter' from the Institution's offices in London.

Members' Appointments

Mr. G. K. Lev (Member 1959) has been appointed Director of Engineering Services, Ministry of Posts, Israel.

Mr. M. Parann (Fellow 1956, Member 1963) is now a Lieutenant Colonel on the Israeli Army Reserve and has recently been appointed Material Control Manager of Tadiran Ltd.

Index to Volume 39

The June issue of *The Radio and Electronic Engineer* completed Volume 39 and an Index will be sent out with the August issue.

U.D.C. 621.397.62

Solid-State Television Receivers—A Pattern of Second Generation Design for Monochrome and Colour

By

P. L. MOTHERSOLE,

C.Eng., F.I.E.E., F.I.E.R.E.†

Presented at the I.R.E.E. 12th National Radio and Electronics Engineering Convention, held in Sydney in May 1969, and reprinted from the Proceedings of the Institution of Radio and Electronics Engineers, Australia (Volume 30, No. 12, December 1969).

Semiconductor technology has now reached a point where high voltage transistors, integrated circuits and thyristors may be used with confidence in consumer equipment. In this paper solid-state monochrome and colour receivers are described that exploit the latest semiconductor devices to provide receivers of a high performance standard.

For horizontal deflection, high-voltage transistors are used and regulation is obtained by the use of a thyristor employed as a phase-controlled rectifier in the power supply. Integrated circuits are extensively employed in both receivers, typical applications being for sound i.f. amplification, in the video signal processing sections and in the RGB matrixing and demodulator circuits of the colour receiver.

1. Introduction

In order to be a viable proposition the overall economics of a solid-state receiver must be comparable with a hybrid one. The cost comparison must take into account factors such as ease of manufacture and mechanical simplification together with the overall reliability and the subsequent reduction in service costs. Broadly speaking, the signal processing and amplifying circuits equate and hence the major cost differences occur in the timebase and power supply areas.

The conventional line deflection circuit requires an output valve and booster diode operating from an h.t. line of 250 to 285 V and in Europe such a potential is obtained with a half-wave rectifier operated directly from the mains supply. The circuit is a high-voltage/low-current configuration employing a series booster diode and a peak potential of about 6 kV appears at the anode of the pentode valve. Furthermore, by operating the valve above the knee of its anode characteristic, stabilized operation is possible by feeding back a d.c. control potential to the grid that is proportional to the peak anode potential. The basic circuit is shown in Fig. 1. This configuration is not suitable for semiconductors, which by comparison to a valve are higher current/lower voltage devices and the parallel efficiency diode circuit is therefore used. In such a configuration the peak voltage applied to the output device is approximately 10 times the h.t. potential. However, a stabilized mode of operation is not possible with semiconductors due to the limited dissipation and hence a stabilized h.t. line is necessary.

† Pye TVT Limited, Cambridge, England; formerly with Mullard Central Applications Laboratories, Mitcham.

2. Power Supply and H.T. Potential

A transistor television receiver is normally characterized by a mains transformer followed by rectifiers and an electronic stabilizing circuit. The h.t. potential is normally some 30 to 50 V with an h.t. current of between 1·5 and 5 A. The peak potentials occurring in the line timebase are typically about 400 V and hence the peak collector current flowing in the line output transistor is typically 5 A for a monochrome receiver and 10 A in the case of a colour receiver. These high currents flow through associated components and the resistive losses give rise to non-linearity of the scan current and hence relatively high-cost components have to be used. A separate efficiency diode is also required since the collector/base junction and reverse base emitter voltage drops give unacceptable linearity. A further by-product of these high currents is voltages induced into the signal amplifying and processing circuits so that the video signal at the detector is often characterized by overshoots that coincide with the start and finish of the line flyback. These small overshoots, which are particularly noticeable on weak signals when the receiver gain is high, are extremely difficult to eliminate and often give rise to poor horizontal synchronization and poor vertical interlace.

The video circuits of a television receiver must operate from an h.t. line that is sufficiently high to

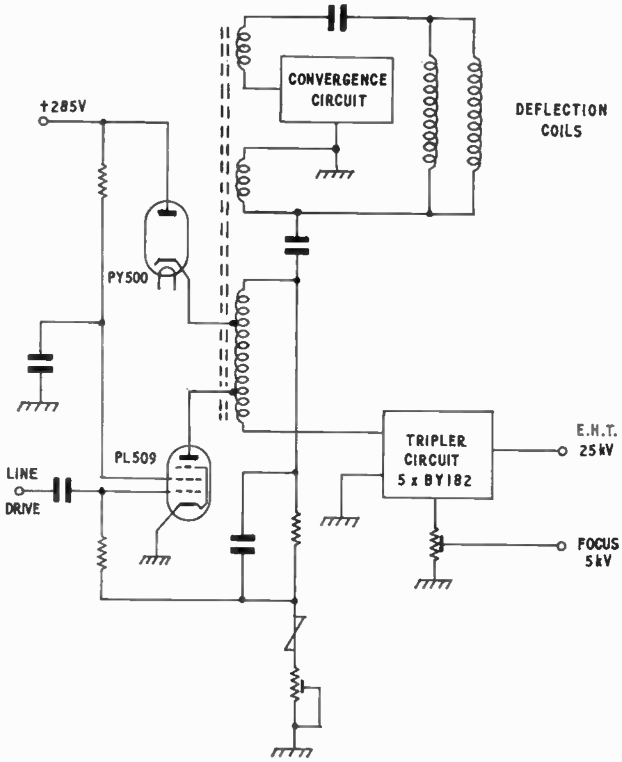


Fig. 1. Basic tripler line output circuit.

provide adequate drive to the picture tube. For monochrome receivers an h.t. line of about 150 V is required but for colour the corresponding figure is 200 V providing RGB drive is used. For colour difference drive, where a signal of some 220 V must be produced, an h.t. line approaching 300 V is necessary. At the moment, the maximum peak voltage that is envisaged for a line output transistor such as the BU105 is 1500 V and this device has a peak current rating of 2·5 A. Hence a single transistor can be used in a monochrome receiver where the scanning and e.h.t. requirement is approximately 3500 VA and two devices may be used for colour where some 5000 VA is required. The most convenient configuration is to connect the two BU105 transistors in series and operate them from the 200 V line.

The conventional electronic regulated power supply employs a rectifier followed by an electronic stabilizing circuit. The stabilizing circuit consists of a transistor in series with the h.t. supply which operates as a variable resistor controlled by a differential amplifier. The series transistor dissipates considerable power, particularly when the mains input voltage is high and at the same time, in the event of it going short circuit, the h.t. line in the receiver is increased considerably. An alternative approach is to replace the mains rectifier diode with a thyristor used as a phase-controlled rectifier. In this way the rectified d.c. potential is maintained constant by varying the conduction angle of the thyristor. The main advantage of this technique is that the circuit is virtually dissipationless and, in the event of the thyristor going short circuit, no d.c. output is produced. Unfortunately, however, the phase-controlled rectifier cannot be used for smoothing. In Europe where transformerless techniques are commonly used for television receivers, the thyristor is driven directly from the 220/240 V mains input. A typical circuit is shown in Fig. 2.

The mains waveform is delayed by the network R1C1 and applied to the trigger diac. When the potential reaches the breakover voltage, the diac conducts, triggering the thyristor. The charging current for C1 is controlled by the transistor, the base

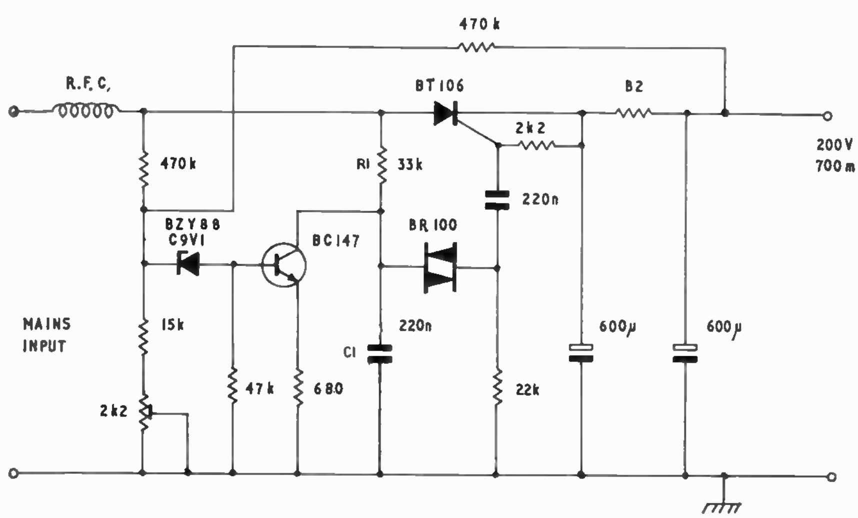


Fig. 2. Half-wave thyristor power supply.

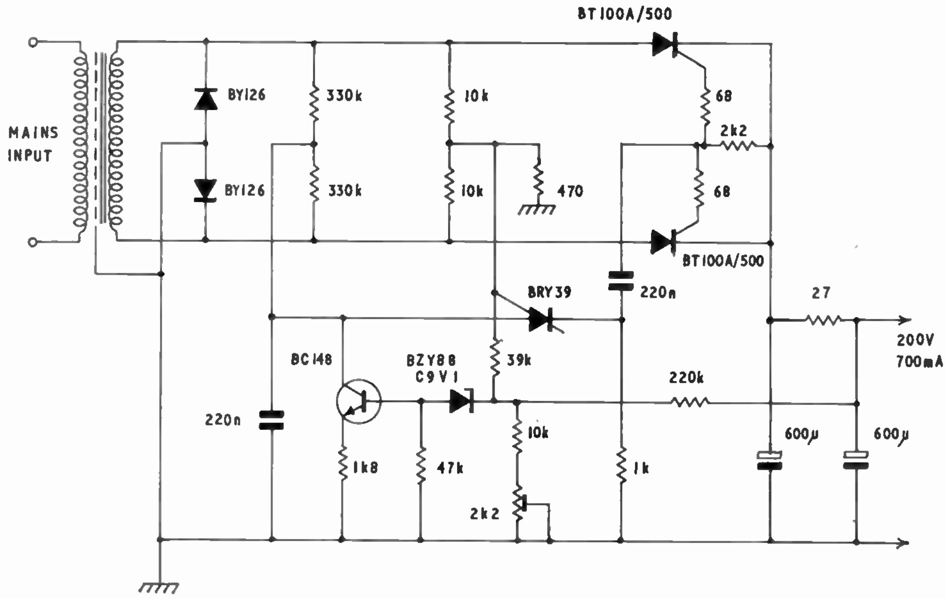


Fig. 3. Full-wave thyristor power supply.

circuit of which is connected to a feedback and reference network designed to achieve a constant output.

When a mains transformer is required for isolation purposes, two thyristors may be used in a bi-phase circuit (Fig. 3) or alternatively a bridge rectifier used, feeding a single thyristor phase-controlled rectifier.

The vertical timebase and audio output circuits can also operate from h.t. potentials of about 130 to 200 V but a transformer is necessary to match to the low impedance of the deflection coils and loudspeaker. When transformerless output circuits are used the required h.t. line is determined by the load.[1] Whilst high resistance loudspeakers are available, to enable relatively high potential h.t. lines to be used, high impedance scanning coils are not very practical due to cross-talk difficulties. Existing coil designs require h.t. lines of some 30 to 40 V. The signal amplifying and processing circuits require potentials in the range 20 to 30 V.

These low voltage circuits can be conveniently driven by rectifying the scanning waveform obtainable from a tap on the horizontal output transformer or alternatively from a winding on the mains transformer.

3. Picture Tube Signal Driver Requirements

The video output stage fulfils the difficult function of linking the low-level transistor circuits to the thermionic picture tube which requires a drive of some 80 to 120 V.[2] Since its only load is that of the picture tube, the collector load resistor can be quite a high value, typically 4·7 kΩ, and simple emitter compensation used. When a d.c.-coupled video circuit is used, some form of mean current limiter is required in the e.h.t. system. A typical circuit is shown in Fig. 4.

A major hazard to the video output transistor is the transient energy produced at the cathode of the picture tube in the event of an internal flashover.[3] Experience has shown, however, that the transistors situated in remote areas of the receiver are often damaged or destroyed by the circulating currents associated with flashovers. To prevent flashover difficulties, a series impedance must be included between the picture tube cathode and the video transistor with spark gaps connected between the tube electrodes and the external

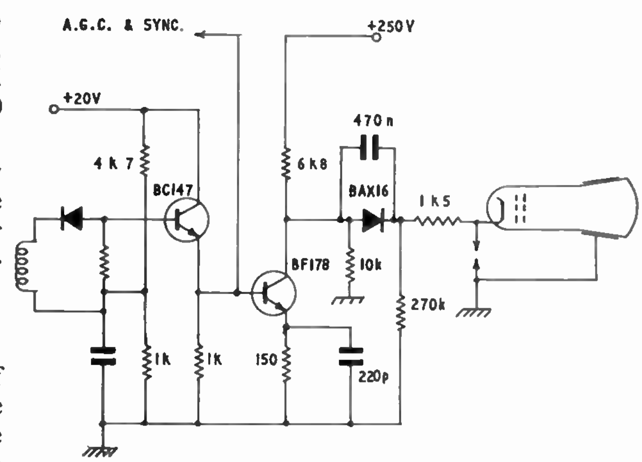
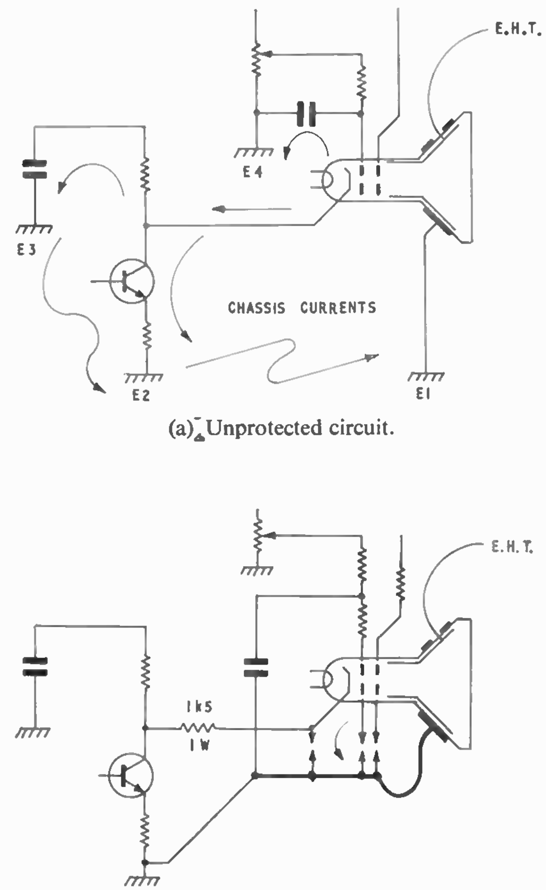


Fig. 4. Typical video amplifier with beam current limiter.



(a) Unprotected circuit.

(b) Protected circuit.

Fig. 5. Transient current path.

conductive tube coating by means of a low inductance lead. In addition a single earth connexion should be made between the cold end of the spark gaps and the chassis. This circuit arrangement ensures that the transient current flows only in the low inductance loop and virtually no transient energy is applied to the video or receiver circuits. The current paths are illustrated in Fig. 5.

In a valve or hybrid receiver, colour difference drive for the picture tube is normally employed. With such a technique the luminance signal is applied to the three cathodes of the picture tube and the colour difference signals are separately applied to the control grids. The amplitude of the colour difference signals is required to be some 220 V peak-to-peak in the case of the (B-Y) signal and hence an h.t. line of about 285 V is necessary. The alternative technique is to employ separate RGB drive in which the luminance or Y signal is matrixed with the two demodulated colour difference signals, (R-Y) and (B-Y), to form the individual R, G and B signals at low level. Traditionally receivers that employed RGB drive have had a tendency for black

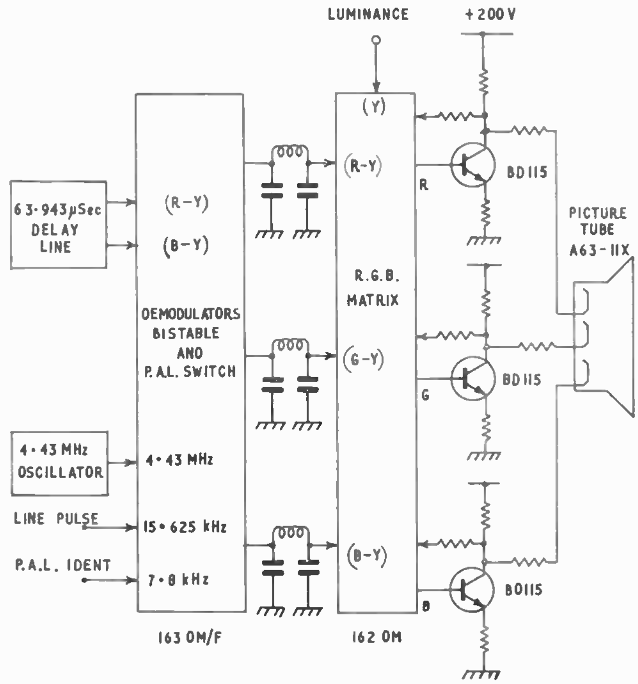


Fig. 6. RGB drive system using integrated circuits for colour signal demodulation and feedback amplifier matrix.

level instability, which results in coloration of a monochrome picture, and chromaticity errors in the case of a colour picture.

In order to obtain adequate stability it is essential that the RGB amplifiers should have identical transfer characteristics and at the same time the d.c. stability must be of a high order. The black level of the output signals should have a differential shift of less than 1 V for all causes, that is, mains input variations and temperature changes. Conventionally such stabilities have only been achieved by the use of fairly elaborate clamping circuits. The use of integrated circuits enables more complex systems to be economically built and at the same time individual amplifiers can be thermally linked by manufacturing them on the same chip. Hence the matrix circuitry can be integrated together with the first stages of the individual RGB amplifiers to obtain the required thermal stability and overall d.c. feedback employed to stabilize the individual transfer characteristics. Such a system is illustrated in Fig. 6. When such amplifiers are operated from a stable 200 V supply, excellent signal drive stability results.

4. Areas of Possible Integration

At first sight all the low-level signal amplifying and processing circuits lend themselves to integrated techniques. Unfortunately integrated circuits have certain restrictions and these tend to define the areas of integration. The main restrictions arise because it is virtually impossible to integrate inductors or capacitors and hence circuits that involve selectivity are not economically integratable. Furthermore the cost of an integrated circuit is to some extent a function of the 'pinning' since the number of connexions tend to impose a cost penalty and the operating levels are limited to about 12 V or so which precludes integration of the output circuits. From an economic point of view an integrated circuit requires to replace a reasonable number of active and passive devices in order to make it a viable proposition but at the same time it must not be unduly restrictive as it will be required to be used in various receiver designs.

As a result of a detailed system study it was apparent that integration of the video signal processing functions following the video detector would be a logical starting point. In the decoding circuits, signal matrix and demodulators with the associated PAL switch were again viable propositions. At first sight the i.f. amplifiers were attractive propositions but in practice only the sound i.f. amplifier was a viable one providing the expensive ratio detector transformer could be eliminated. This can be achieved with an integrated circuit since several cascade amplifier stages provide a high value of a.m. suppression which enables a simple quadrature detector to be used in place of the conventional ratio detector.

5. Complete Monochrome Receiver

The block diagram of a 625-line monochrome receiver using the techniques outlined above is shown in Fig. 7.

The three transistor, v.h.f. tuner employs three varicap diodes for tuning purposes and has a power gain of about 25 dB with an associated noise figure of about 6 dB. Switching between bands 1 and 3 is effected by the five switching diodes and the h.t. supply to the varicap tuning circuit is stabilized by a Zener diode.

The main advantages that accrue from a varicap tuner is that the tuner itself can be located on the i.f. amplifier panel with only d.c. wiring to the station selector mechanism. This facilitates cabinet styling and remote control.

The output from the tuner feeds a three-stage vision i.f. amplifier employing four transistors, two of them forming a cascode middle stage. This configuration provides adequate gain and excellent stability.[4]

The sound i.f. amplifier employs an integrated circuit, TAA570, the output of which drives a transformerless output circuit. Incorporated into this integrated circuit is a d.c.-operated volume control. This again enables the volume control function to be performed by a d.c. potential from the control panel.

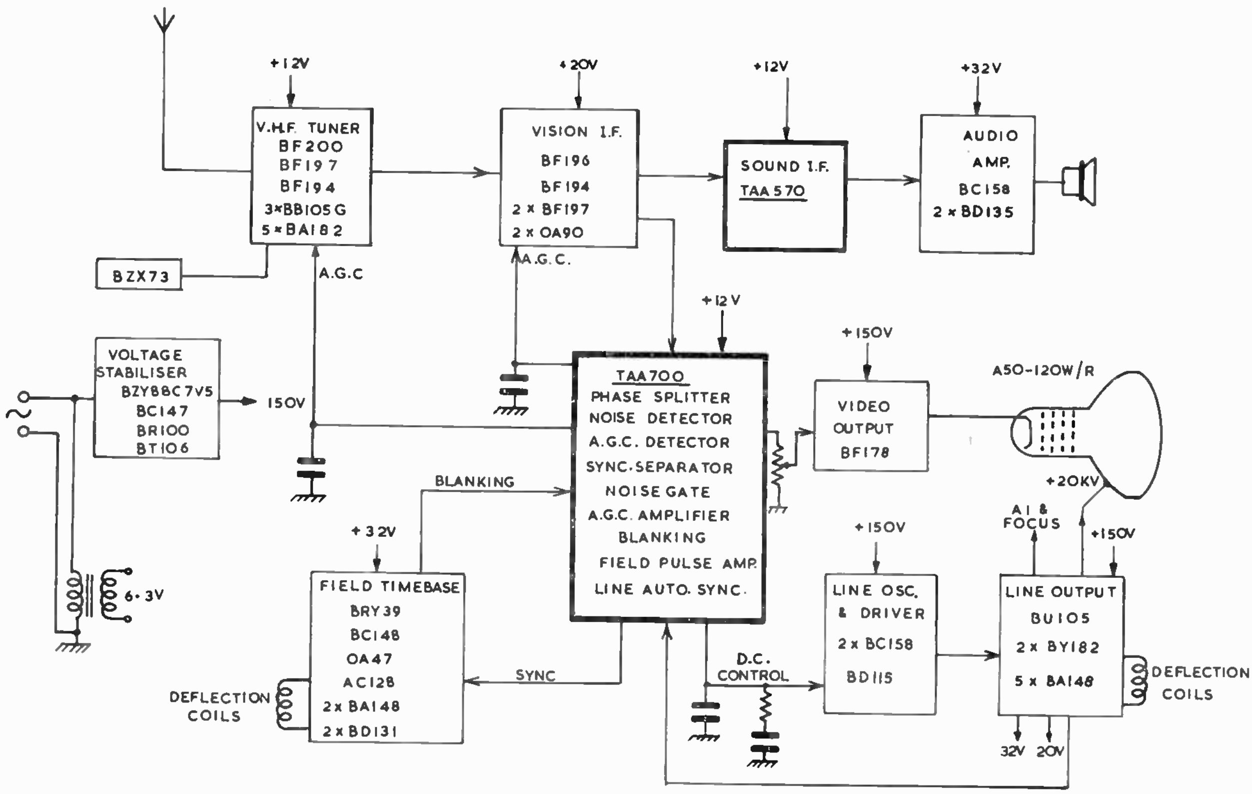


Fig. 7. 625-line transistor television receiver.

The a.g.c., synchronizing pulse separation and video drive functions are carried out in the integrated circuit type TAA700. The outputs from the circuit are either d.c. potentials or low impedance waveforms and hence no difficult layout problems are introduced for the receiver designer. By providing some voltage gain in the circuit, the choice of operating level for the video detector can be optimized. A signal level of about 2 V peak is a reasonable compromise for a normal detector from considerations of linearity and harmonic distortion. A low impedance video output signal of about 6 V peak-to-peak is provided which is more than adequate for driving a video output stage. The d.c. level of the video signal in the integrated circuit is carefully maintained and hence, if a bridge type contrast control is used, the black level of the displayed picture is held constant when the contrast is varied. A blanking facility is also incorporated within the integrated circuit which largely removes the necessity for high voltage blanking pulses to be applied to the picture tube.

The a.g.c. system controls the gains of the r.f. stage and the i.f. amplifier so that the detected video signal is held at a constant amplitude. For correct a.g.c. operation the initial control must be on the i.f. stage up to a predetermined attenuation and then, after this, control must be transferred to the r.f. stage. Such an a.g.c. distribution network can be readily incorporated into the integrated circuit with the required a.g.c. gate.

A horizontal gated sync. tip a.g.c. system has been chosen with an internally fixed threshold level to hold the video signal at 2·2 V amplitude. Line gating is used to minimize the variation of a.g.c. output potential which occurs during the field pulse period and hence enables a short a.g.c. time constant to be chosen. The signal level at which a.g.c. action is switched from the i.f. amplifier to the tuner is determined by the bias potentials applied to the controlled transistors and also cross-over potential applied to the integrated circuit. Hence this aspect of a receiver design is determined by the receiver designer and is completely external to the integrated circuit.

Because of the higher voltage and power levels encountered in approaching the deflection output stages and the necessity to maintain flexibility in design and choice of devices in the timebase oscillator circuits, the boundary of the integrated circuit in the synchronizing area must be carefully considered.

The synchronizing pulse separating function together with the subsequent pulse clipping circuits are well suited to integrated circuit techniques. A convenient boundary for the integrated circuit in the line synchronizing channel is immediately after the line flywheel phase discriminator. At this point in the system a large group of components are encountered that cannot be integrated (flywheel filter and phase reference shaping networks). For synchronizing the vertical timebase a pulse of some 10 V amplitude is provided by the integrated circuit which is an identical waveform on both odd and even fields.

It is important that both the a.g.c. and synchronizing channels should be protected against impulsive interference. A desirable feature of such a protection system is that it should require no pre-set controls. This is possible by means of a technique of separating interference from the video signal on a combined amplitude and frequency basis and using the resulting information to gate the a.g.c. and synchronizing channels. The phase detector in the integrated circuit is also preceded by a horizontal-pulse gated circuit. This gate is operated by the externally applied flyback pulse and only allows synchronizing pulses that occur during the flyback period to be applied to the phase detector. This gate has a comb filtering action and reduces the amount of noise applied to the phase detector when the timebase is in synchronism and only allows approximately coincident synchronizing pulses to be applied to the phase detector when it is out of synchronism. This latter action improves the catching range of the flywheel system since only pulses which are approximately correctly phased can influence the phase detector. The comb filter therefore improves the noise performance of the horizontal oscillator without significantly reducing the catching range.

A conventional video output stage is used employing a BF178 transistor operated from the +150 V h.t. line.[2]

A transformerless field timebase is employed utilizing two BD131 output transistors in a single-ended push-pull configuration. A silicon controlled switch is used as the vertical oscillator, directly synchronized by a pulse from the TAA700. The sawtooth output drives the output stage through a current amplifier.[1]

A sinewave line oscillator is employed, controlled in frequency by a reactance stage. The drive waveform required by the BU105 transistor is substantially a square-wave and hence the drive waveform is obtained from the centre of the line oscillator waveform. A single BU105 is used for the horizontal output operating from the 150 V h.t. rail. No efficiency diode is required with the BU105 since the forward-biased collector junction performs the function of a parallel efficiency diode. The e.h.t., A_1 and focus potentials are obtained by rectifying flyback pulses on the line output transformer and the low voltage supplies for the vertical timebase, audio output and signal processing circuits are obtained by rectifying a scan waveform from the line output transformer.

The thyristor (type BT106) is used as a phase-controlled rectifier in the power supply and enables a

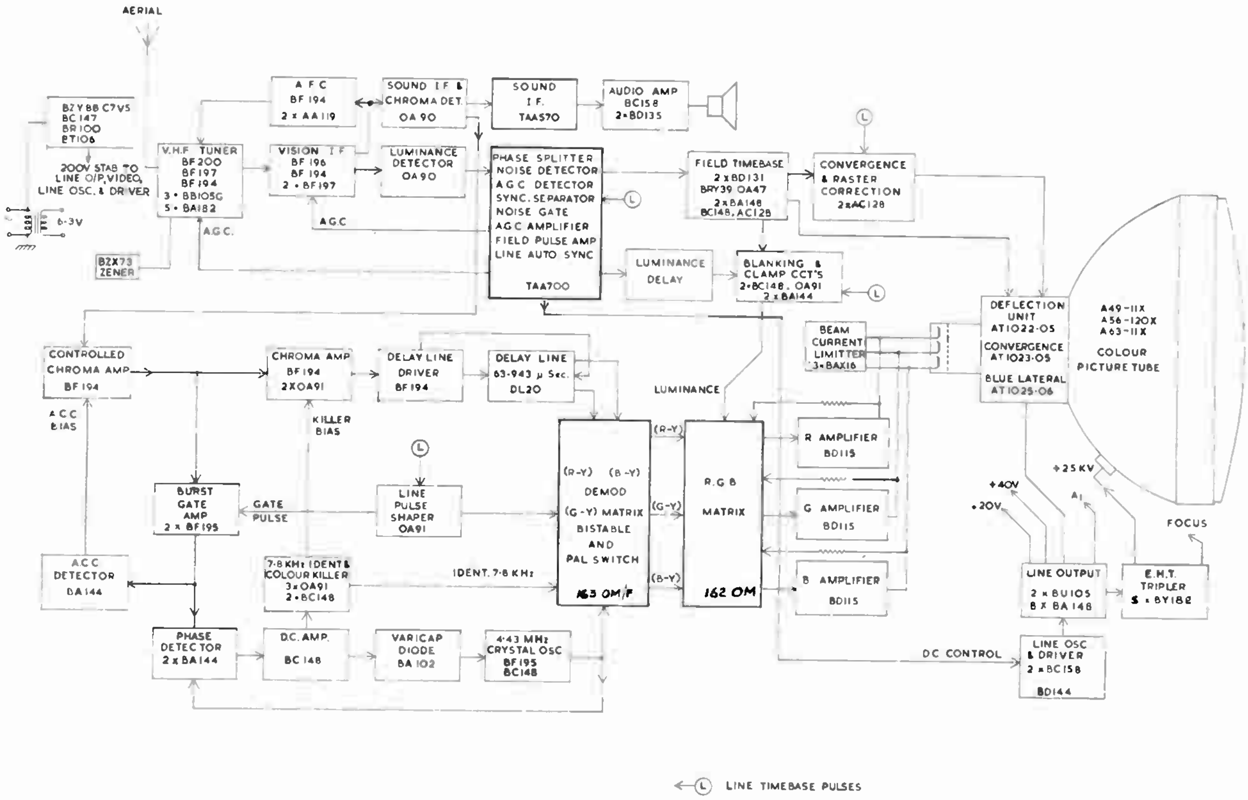


Fig. 8. 625-line transistor colour receiver (RGB drive).

well-regulated 150 V potential to be obtained suitable for driving the line output stage and video amplifier directly with a minimum of heat dissipation. The total power dissipation in the receiver described is typically 60 W.

6. Complete Colour Receiver

The block diagram of a 625-line PAL colour receiver using the techniques outlined earlier in this paper is shown in Fig. 8.

The signal amplifying and luminance processing circuits are similar to those described for the monochrome receiver; the major difference in the colour receiver is the additional colour decoding circuits and higher power timebases. The tuning of a colour receiver is more critical than a monochrome one and a.f.c. tends to be a desirable feature.[4]

In the colour receiver the luminance output from the TAA700 is current matched into the 600 ns luminance delay network. The output from the luminance delay network feeds into a blanking and clamping circuit. This circuit removes the transmitted d.c. component and reinserts it at a level determined by the brightness control. The output from the luminance clamp drives the RGB matrix integrated circuit, 162 OM. This circuit is also driven by the demodulated colour difference signals.

Three identical output circuits drive the three individual cathodes of the picture tube with the R, G and B signals respectively. Individual overall feedback is effected by resistors connecting the output of the amplifiers with the RGB matrix circuit. Brightness control is effected by variation of the d.c. level of the luminance signal applied to the matrix circuit. The brightness control is applied to the picture tube through the RGB amplifiers and hence excellent grey scale tracking is maintained at all brightness settings since the individual amplifiers are aligned for a particular picture tube.

It is very desirable in a colour receiver to prevent excessive mean current flowing in the picture tube. Under normal signals conditions the total picture tube current will be typically 1 mA or so but on high key scenes this could easily increase three or four times. Such high currents would give rise to very poor definition pictures due to the excessive spot size and at the same time the picture tube may be damaged by the excessive power dissipation. Some form of current limiter is therefore required and two techniques can be employed. One is to monitor the e.h.t. current and when this exceeds a pre-determined value to feed a bias potential back to the brightness or contrast circuit to reduce the drive. The main disadvantage of this technique is a tendency for brightness flicker on the threshold of beam current limiter action. To minimize this, a long time-constant for the brightness control must be employed. The alternative technique, which can be used with an RGB drive system, is to employ separate beam current limiting diodes in series with each cathode as shown in Fig. 4. These circuits do not have the same tendency for brightness flicker on the threshold of operation as do feedback types but the grey scale of the picture is disturbed when they operate since the picture is essentially a.c. coupled. However, since the beam current limiters only operate on very high key scenes the loss of low level grey scale tracking is relatively unimportant.

The chrominance signal is applied to a gain-controlled amplifier, the output of which drives the burst gate and second chrominance amplifier. The burst gate is operated by a shaped pulse from the line timebase and its output feeds the a.c.c. detector and a.p.c. phase detector. A crystal oscillator is used in a conventional a.p.c. system, its frequency and phase being controlled by a varicap diode.

In the PAL system the colour burst is phase-modulated $\pm 45°$ on alternate lines. The mean phase is constant and hence the phase of the 4·43 MHz oscillator is maintained constant. The output of the phase detector alternates at a half line frequency and this waveform, known as the 7·8 kHz identification waveform, is used to lock the PAL switch in the correct phase. The switch is contained in the integrate circuit 163 OM/F and is driven with a pulse from the line timebase. Since the line timebase is synchronized by the line flywheel system, the PAL switch is automatically protected from the effects of interference. Synchronization of the switch is effected by applying the 7·8 kHz waveform to the integrated circuit.

The second chrominance amplifier is preceded by the saturation control and it is also controlled by the colour killer bias. The colour killer is operated from the 7·8 kHz identification waveform which is only produced when the sub-carrier oscillator is phase locked with the colour burst. Hence a chrominance output is only produced in the event of firstly, a colour burst being present and secondly, the reference oscillator being correctly synchronized. The colour burst itself is used to drive the a.c.c. detector which controls the gain of the first chrominance amplifier. Hence the a.c.c. circuit maintains the colour burst at a constant amplitude at the input to the phase detector which ensures that the chrominance signal at the input to the chrominance amplifier is also of constant amplitude.

Variation of saturation is effected in the second chrominance amplifier independent of the colour burst. The output of the second chrominance amplifier drives the 64 μs chroma delay line. The output from the glass delay line is mixed with an undelayed signal in a transformer to produce separate (R-Y) and (B-Y) signals for the colour demodulators which are contained in the 163 OM/F. The (G-Y) matrix is also contained in the integrated circuit and the three colour

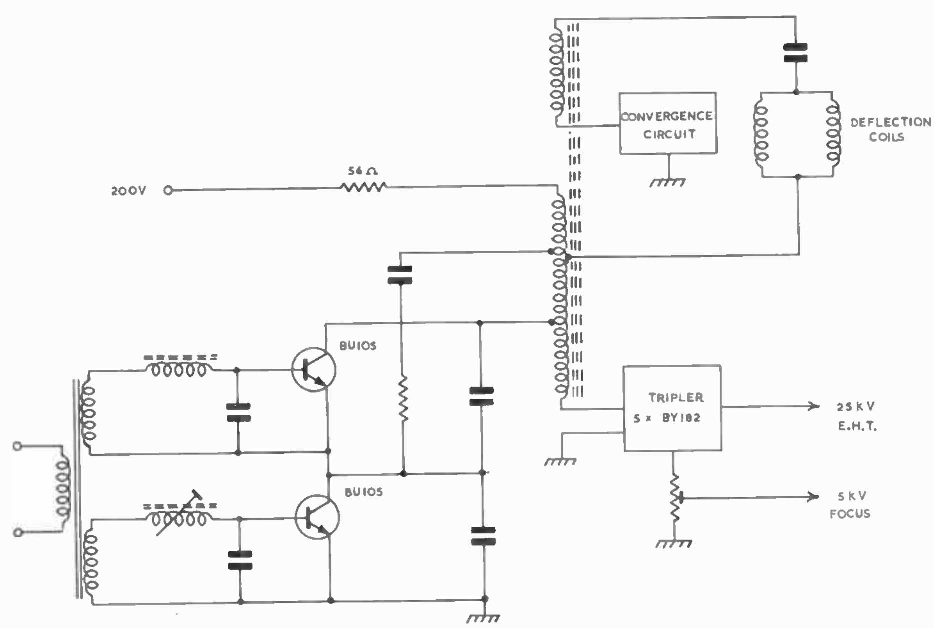


Fig. 9. Basic line output circuit using two BU105 transistors.

difference output signals drive the RGB matrix integrated circuit 162 OM via simple π filters as shown in Fig. 6.

The line oscillator and driver are similar to those employed in the monochrome receiver except that a higher power driver stage is required capable of driving two BU105 transistors.

The basic line output circuit is shown in Fig. 9. The two BU105 transistors are series connected and the output circuit operates from a stabilized 200 V potential. The peak voltages that occur across each transistor during flyback are equalized by tapping the junction of the two transistors into the line output transformer. It is, however, necessary to ensure that the storage time of the two transistors is matched and this is achieved with the variable inductance in the base of the lower transistor. This inductor is set for minimum circulating current in the voltage equalizing tap. The 56 Ω resistor in series with the h.t. line limits the peak current and voltage that can occur during a flashover and the small capacitors connected between the base and emitter of the transistors are used to prevent voltage transients being applied to the driver circuit and triggering the transistors inadvertently. In order to ensure that the output circuit is safe in the event of picture tube flashover, it is essential that considerable attention be paid to the earth loops associated with the drive circuits, tuning capacitors and h.t. decoupling points. The e.h.t. potential is obtained by voltage tripling employing five BY182 diodes and the focus potential for the picture tube is obtained from the first stage of the tripler.

The 200 V stabilized supply for the line oscillator, driver and output and the video circuits is supplied by a thyristor acting as a phase-controlled rectifier (Fig. 2) and the low voltage supplies are derived from the line output stage by scan rectification. The performance of the receiver is therefore virtually independent of the supply mains over the range 185 to 265 V. Below this level the h.t. potential falls as the thyristor operates as a normal rectifier.

The convergence and raster corrector circuits are similar to those used in a hybrid receiver.[5] Experience has shown that the current-driven convergence circuits are extremely stable and consistent in operation.

7. Conclusions

In this paper all-solid-state monochrome and colour receiver concepts have been described that exploit semiconductor devices to provide very stable high quality pictures. For horizontal deflection, high voltage transistors are used with a thyristor employed as a phase-controlled rectifier in the regulated power supply. In the signal processing area, integrated circuits are extensively employed. An important feature of the receivers described is the low total power consumption, typically 60 W for a monochrome and some 130 W in the case of the colour receiver. This low dissipation enables compact and economical receiver designs to be achieved.

8. Acknowledgments

The author would like to acknowledge the contributions made by his colleagues, in particular Mr. A. Ciuciura, Mr. M. C. Gander, Mr. D. S. Hobbs, Mr. D. J. King and Mr. K. E. Martin to the work described in this paper and to the Directors of Mullard Limited for permission to publish it.

9. References

1. Attwood, B. E., 'Transistor field timebases for colour and monochrome receivers', *Mullard Tech. Commun.*, **9**, No. 89, p. 246, September 1967.
2. Gander, M. C., 'Transistor video amplifier design for monochrome television receivers', *Mullard Tech. Commun.*, **10**, No. 96, p. 203, November 1968.
3. Ciuciura, A., 'Flashover in picture tubes and methods of protection', *The Radio and Electronic Engineer*, **37**, No. 3, p. 149, March 1969.
4. Bissmire, P., 'Single system i.f. amplifier for U.K. 625 line system', *Mullard Tech. Commun.*, **10**, No. 98, p. 234, March 1969.
5. King, D. J. and Thomson, E. C., 'Pincushion correction and convergence for 625 line colour receivers', *Mullard Tech. Commun.*, **70**, No. 97, p. 217, January 1969.

Manuscript received by the Institution of Radio and Electronics Engineers, Australia, on 30th May 1969. (Paper No. 1328/ Com. 30.)

Contributors to this Issue

Professor D. G. Tucker (F.1953) is head of the Department of Electronic and Electrical Engineering at the University of Birmingham, an appointment he has held since 1955. Professor Tucker's career began in the Post Office Engineering Department and he was at the Dollis Hill Research Station for sixteen years, reaching the position of senior executive engineer in charge of transmission measurements. During this period he obtained his Ph.D. and D Sc. degrees from the University of London. From 1950 to 1955 Dr. Tucker was with the Royal Naval Scientific Service, being concerned principally with underwater acoustics. He received a special merit promotion to senior principal scientific officer in 1952.

Professor Tucker served on the I.E.R.E. Council from 1959–1962 and 1965–1966 as well as on the Education and Research Committees; he is currently chairman of the Organizing Committee for the Conference on Electronic Engineering in Ocean Technology.

Author of several books and many papers in engineering and scientific journals, Professor Tucker's papers in the Institution's Journal have gained him both the Clerk Maxwell Premium (in 1961) and the Brabazon Award (in 1958).

S. D. McLaughlan graduated in applied physics at the Royal College of Science and Technology, Glasgow, in 1956. He then studied theoretical physics at the University of St. Andrews and was awarded an M.Sc. degree for work on electron interactions in solids. Since 1960 he has been employed at the Royal Radar Establishment, Malvern, where he has been working principally on electron spin resonance studies of materials.

Dr. M. J. Taylor graduated in physics at Oxford University in 1957 and was awarded the degree of D.Phil. in 1961 for research on radio-frequency resonance in gases. After a period of two years at the National Research Council in Ottawa, working on gas lasers, he joined the Royal Radar Establishment, Malvern, in 1963 to work on materials for solid-state lasers. He currently heads the project on colour centre displays.

Dr. D. J. Marshall received the degree of B.Sc. from Imperial College, London, in 1959 and was awarded his Ph.D degree in 1962. After a period of postdoctoral work at Imperial College on the hydrothermal synthesis of alumino-silicates he worked for two years at Bell Telephone Laboratories in the U.S.A. on the growth of various oxide single crystals. He joined the Royal Radar Establishment, Malvern, in 1966 where he is engaged in the growth of crystals by hydrothermal and flux techniques.

Dr. P. A. Forrester received his B.Sc. from Manchester University in 1952 and the degree of Ph.D from Kings College, Durham University in 1957. He has worked at the Royal Radar Establishment, Malvern, since 1956, apart from a two-year leave of absence spent at the Bell Telephone Laboratories in the U.S.A. Dr. Forrester has published papers on electron spin resonance, microwave and optical masers, and colour centres.

A biographical note on Mr. P. L. Mothersole was published in the June 1969 issue of the *Journal*.

ELECTRONIC ENGINEERING IN OCEAN TECHNOLOGY

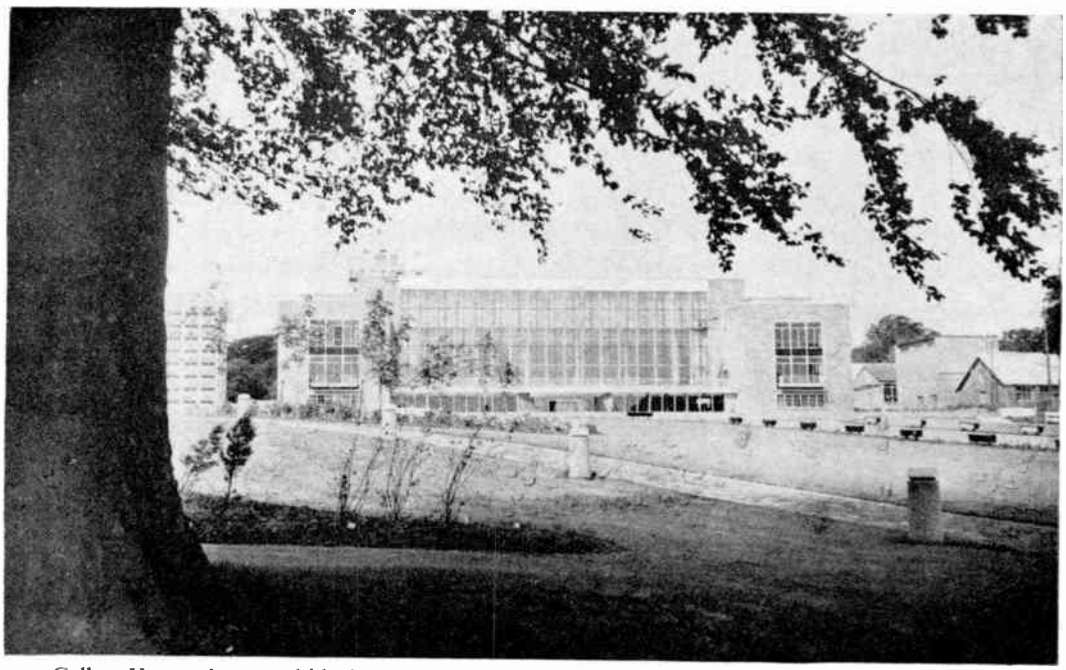
Institution Conference in Swansea in September

For its second Conference on the applications of electronic engineering at sea, the Institution is again going to a University with marine associations. Indeed, whereas the University of Southampton, where the 1966 Conference on 'Electronic Engineering in Oceanography' was held, is situated several miles from the open sea, the University College of Swansea is literally on the edge of Swansea Bay, looking out over the Bristol Channel. The College has for a number of years been active in various fields of oceanography, the emphasis in post-graduate research work being on marine geology and geophysics, sedimentology, marine biology and marine chemistry, and it is one of the few Universities in Great Britain at which oceanography can be offered as a Part 2 subject for a B.Sc. degree.

The Conference programme takes the theme of 'the gathering, transmission, processing and display of information from the sea' and it has been grouped under sessions covering oceanographic sensors (dealing with measurement of water quality and water movement respectively), sonar systems and applications, communication and telemetering, shipboard computers, and submersibles and underwater navigation. Full details of the papers to be presented will be given in the final programme, meanwhile the provisional programme given below outlines the subjects to be included.

Many of the subjects to be discussed in the sessions of the Conference will be supported by demonstrations and while it is usual at most conferences to have an adjoining room devoted to working and static displays of equipment relevant to the papers, at Swansea it is hoped that a good proportion of the demonstrations will be given aboard ship either in Swansea Docks or in some cases at sea. The R.R.S. *John Murray*, which is operated by the Natural Environment Research Council, is visiting Swansea and will be on show throughout the second day of the Conference. One of the most modern of the survey ships working with the Hydrographic Department of the Royal Navy, H.M.S. *Hecate*, will also be at Swansea and her Automatic Data Logging System and associated oceanographic and navigational equipment will be open for inspection. The third vessel to be present during the Conference will be M.Y. *Navigator*, owned by the Decca Navigator Company, which is equipped with marine navigational and position fixing equipment. Of particular interest is the possibility of demonstrations of equipment at sea being given aboard R.S. *Ocean Crest* which is the College's own oceanographic teaching and research vessel.

The great advantage of an electronic engineering conference held at a University away from the centres of research and development in South East England is that the majority of those taking part will wish to



College House, the central block on the campus which contains the refectory, dining rooms, common rooms, and facilities such as post office, bank, bookshop and barber's shop.

stay on the college campus and the University College of Swansea is well provided with modern halls of residence only a few minutes' walk from the conference lecture theatres. Meals will be taken in College House where the Conference Dinner will be held, the President of the I.E.R.E., Mr. H. F. Schwarz, will give a reception on the first evening of the Conference and there will be other social functions on subsequent evenings. All these arrangements will foster informal discussion.

Papers to be presented at the Conference will be pre-printed and bound in a volume which will be sent a few days before the Conference opens to all who have registered. Copies of the volume will be available for sale afterwards, while a selection of the papers will be reprinted in *The Radio and Electronic Engineer*. Registration forms for the Conference will be available shortly: the registration fee to members of the I.E.R.E. and I.E.E. is £15; the non-member registration fee is £18, and research students may attend at the special reduced rate of £10. The charge for accommodation in the College's hall of residence for three nights (Monday–Wednesday) is £8 15s. and this includes a ticket for the Conference Dinner. It will be possible to stay in the College on the Sunday night preceding the Conference and there will be a limited number of rooms available on the Thursday night following the close of the Conference.

PROVISIONAL PROGRAMME OF THE CONFERENCE

MONDAY, 21ST SEPTEMBER, Afternoon

Session 1: OCEANOGRAPHIC SENSORS:

I. Water Quality (Temperature, Depth or Pressure, Content)

The Undulating Oceanographic Recorder Survey Using Ships of Opportunity.
Thermistor Thermometer for Measurement of Oceanic Temperature Microstructure.
Submersible Water Quality Monitoring Equipment.
Counting and Recording Equipment for Coastal and Estuarine Pollution Studies.
Acquisition of Data with a T.S.D. Probe.
Pressure-Balanced Electronics for Underwater Instrumentation.

Evening **President's Reception**

TUESDAY, 22ND SEPTEMBER, Morning

Session 2: SONAR SYSTEMS AND APPLICATIONS

Automatic Estimation of Catch in Sea-Bed Trawling.
A Quantitative Fish Counting Echo Sounder.
Signal Processing in a Large Side-scan Sonar
A Digital Correlator using a M.O.S.T. Polarity Coincidence Detector.
Heave Correction in Echo Sounding.
Hydrographic Surveying Using Mirror Sonars.
Cathode-Ray Tube Display and Correction of Side-scan Sonar Signals.
High Resolution Sonars and Sonar Displays
An Electronic Sector Scanning Sonar

Afternoon

Session 3: SHIPBOARD COMPUTERS

Shipboard Computer Systems for Research and Survey Applications.
An Automatic Data-Logging System at Sea.
The Computer in Fishing Gear Technology and other Marine Applications.
A Portable Computer System for Oceanographic and Acoustic Research.
The Use of Shipborne Computers for Navigation.

Evening

Invited Lecture: 'The Cruise of the *Ben Franklin*' by K. R. Haigh.

WEDNESDAY, 23RD SEPTEMBER, Morning

Session 4: OCEANOGRAPHIC SENSORS:

II. Water Movements (Waves, Tides and Currents)

Environmental Monitoring in the North Sea.
Wavebuoy Techniques.
A Directional Wavebuoy Operating up to 15 Hz.
Recording of Internal Waves by Means of an Isotherm Follower Device.
A Digitally Recording Off-Shore Tide Gauge.
Measurement of Vertical Currents in the Ocean.
A Free-fall Electromagnetic Current Meter for Ocean Current Measurement.
Static Electromagnetic Current Meter.

Afternoon

Session 5: SUBMERSIBLES AND UNDERWATER NAVIGATION

Electronic Instrumentation Requirements for Submersibles, Habitats and Divers.
Underwater Navigation Systems for Manned Submersibles.
Acoustic Navigation System for Exploration Diving.
Fuel Batteries for Marine Use.
Performance of Optical Viewing System and Sonar Under Turbid Conditions.

Evening **Conference Dinner**

THURSDAY, 24TH SEPTEMBER, Morning

Session 6: COMMUNICATIONS AND TELEMETERING

A Telemetering System for Radio Transmission of Oceanographic Data.
A Data Recording and Transmitting System.
An Experimental Oceanographic Automatic Station.
Underwater Telemetry Using Steel Towing Cable.
16-Channel Real-Time Trawl Mounted Telemeter for Fishing Gear Research.
Acoustic Telemeter for Trawler Applications.
An Adaptive Tracking Array for Underwater Telemetry.
A Helium Speech Processor Operating in the Time Domain.
A Practical Processor for Helium Speech.

U.D.C. 549.621.5: 621.385.82

Colour Centres in Sodalites and their Use in Storage Displays

By

M. J. TAYLOR, D.Phil., A.Inst.P.,†

D. J. MARSHALL, Ph.D.,†

P. A. FORRESTER, Ph.D.,†

and

S. D. McLAUGHLAN, B.Sc., M.Sc.†

Based on a paper presented at the I.E.R.E. Conference on 'Lasers and Opto-Electronics' held at the University of Southampton on 25th to 28th March 1969 and subsequently repeated at the XVIIth AGARD-NATO Symposium on 'Opto-Electronics Processing Techniques' held at Tonsberg, Norway, in September 1969.

Members of the sodalite group of minerals show considerable promise as the screen material in dark-trace storage tubes. The dark-trace is due to the creation of colour centres in the material by the electron beam, and the colour gradually fades at a rate which is governed by the level of light incident on the tube. In this paper recent work on the coloration process is presented and device performance discussed.

1. Introduction

At the present time there is considerable interest in new methods of displaying information obtained from electronic systems, such as radars, computers and aircraft instrumentation, and particularly in devices with a built-in storage capability. Such requirements are usually met at the moment by rather complex and expensive devices such as the scan converter, the direct-vision storage tube, or fixed format displays using a c.r.t. fed from an internal character generator and memory. However, there is a real need for a simple, cheap and reliable device which has a built-in memory, and is capable of giving a bright, continuous and readily erasible display.

An approach to this problem is to use the reversible optical absorption bands produced by creating colour centres in crystalline solids. Colour centres can, in general, be generated both by optical beams, which is described as photochromism, and by electron beams, which we can call cathodochromism by analogy with cathodoluminescence. We have been mostly interested in cathodochromic processes since electrons can be generated, modulated and deflected more easily than light beams. By replacing the phosphor of an ordinary c.r.t. with a cathodochromic material one can make a 'dark-trace' tube in which the electron beam produces a dark spot on a light background. Such a tube may be viewed by external illumination in either reflexion or transmission.

The suggestion that a display might be based upon colour centres was first made by Rosenthal[1] in 1940, and a tube utilizing the generation of F centres in KCl was developed for use in wartime radar systems both in the U.K., where it was known as the skiatron,[2] and in the United States, where it was called the scotophor tube.[3] Although such tubes are still available commercially today, they are not widely used since their properties are far from ideal. Their writing speed is insufficient for many purposes, and the information, once written on, is not easy to erase. Also, persistent images gradually build up due to the aggregation of F centres which can only be removed by heating the screen material. In this paper we report the preliminary results of our work, the aim of which is to try to produce a dark-trace material which will lead to a device with much improved performance; the device should have a writing speed of at least 10^6 spot diameters per second, produce an acceptable contrast, preferably in excess of 50%, and be capable of rapid erasure in under one second or gradual erasure over a period of many seconds, preferably both. The material should be able to withstand many reversals to and from the coloured state without fatigue, and give an active tube life of at least 1000 hours.

Using a demountable c.r.t. apparatus we have examined a large number of inorganic materials, particularly those which were known to be photochromic, and those which were known to colour readily by X-rays or other ionizing radiations. We found many materials which coloured under electron beam excitation to varying degrees, but the greatest writing speeds were obtained from members of the sodalite group of minerals. The purpose of this paper is to report the results of some of our measurements on hydrothermally synthesized members of the sodalite family. These materials are also being studied by other laboratories, and Kiss and Phillips of RCA have recently reported measurements on a sodalite tube which they claimed to be the world's first.[4] However, Ferranti Ltd. of Manchester were making experimental devices at least as early as 1966, and Ivey has recently pointed out that he holds a patent

† Ministry of Technology, Royal Radar Establishment, Malvern, Worcestershire.

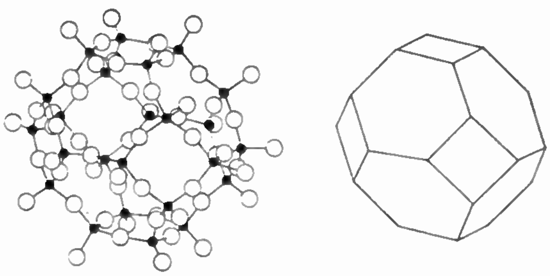


Fig. 1. Part of the alumino-silicate framework in the structure of sodalite (after Bragg, 'Atomic Structure of Minerals', Cornell Univ. Press, 1937). The cubo-octahedral cage structure formed by the Al and Si lattice points is illustrated on the right.

dated 1956 on the use of sodalite as a dark trace tube material.[5] A group at the American Cyanamid Co. have been studying the fundamental properties of sodalites for several years, and recent reports[6] show that their investigations have been closely paralleled by our own.

2. Crystal Structure and Growth

The crystal structure of the sodalite family is built up from cubo-octahedral cages of (AlO_4) and (SiO_4) groups. Each Al and Si atom is at the centre of a tetrahedron of four oxygen atoms, and the cages are made up of rings of four and six (SiO_4) and (AlO_4) tetrahedra, which form the cubo-octahedron as shown in Fig. 1. The cages are stacked together to form a crystal with overall cubic symmetry. Since each oxygen is shared between an Al and an Si atom, the total excess negative charge is equal to the number of Al atoms in the framework and is neutralized by the charge of the ions occupying the sites within the cage. The mineral from which the sodalite group takes its name has the formula $3(Na_2O.Al_2O_3.2SiO_2).2NaCl$, and is usually referred to simply as sodalite. The charge on the framework is balanced by the $(Na_4Cl)^{3+}$ groups located inside each cage. The cages can be sufficiently varied to accommodate different anions, such as S_x^{2-}, SO_4^{2-}, OH^-, etc. Many variations can be synthesized by changing the contents of the cage; for example, by replacing the NaCl in sodalite by $NaOH.xH_2O$, basic sodalite is produced. The mineral nosean contains Na_2SO_4, and the pigment ultramarine contains Na_2S_x. Other changes can be produced by cation diffusion, and such exchange-reactions have been extensively studied by Barrer[7] and his students. The detailed coloration behaviour of the sodalites depends upon their chemical composition, and we have synthesized compounds in which the Cl ions have been replaced by OH, Br and I, and the Si by Ge, and the Al by Ga. In addition, the early work of Kirk[8], Medved[9] and Radler[10], and the more recent work of Williams *et al.*,[6] has indicated that sulphur should be present for maximum photochromic sensitivity, and this has been confirmed by our own observations. We have also found similar sensitization with Se and Te.

The sodalites used in our experiments were grown by the hydrothermal method. Samples of up to 0·5 g of crystalline powder were produced in platinum capsules of 1·5 ml internal capacity heated under pressure in Tuttle-type autoclaves using a commercial apparatus made by Tem-Pres Inc. The starting materials, which were fine powders of Analar quality or better, were weighed into the capsule, the appropriate alkaline solution was added from a hypodermic syringe and capsule was sealed with a small arc-welder. The run lasted from 3 to 5 days at around 420°C and 550 bars. The autoclaves were then allowed to cool, the capsules opened and the product thoroughly washed and dried. The sodalites were produced as crystalline white powders consisting of euhedral rhombic dodecahedra usually about 25 μm across. Crystals up to about 0·5 mm in size could be produced if desired. To produce larger crystals, or to synthesize larger quantities of sodalite powder, platinum-lined Morey-type autoclaves[11] with an internal capacity of 65 ml were used. According to the temperature gradient and concentrations employed these could be used to grow larger crystals or to produce about 15 g at a time of uniform crystalline powder.

The products were identified by X-ray powder diffraction using Debye–Scherrer cameras; all the lines observed were indexed to a body-centred cubic cell and there was no trace of any second phase. Analysis for halide and sulphur content was carried out by X-ray fluorescence or mass spectroscopy, and, as a check, a number of samples were fully analysed by classical wet chemical methods. The analyses showed the stoichiometric amounts of Na, Al and Si, with variable quantities of halide, sulphur and hydroxide filling the cavities; the Al/Si ratio was $1·00 \pm 0·01$ and there were $8·0 \pm 0·15$ Na atoms per unit cell. Typical levels of impurities were of the order of 50 parts in 10^6 each of Fe and Cr.

3. Experimental Results

The measurements which we have made on these materials can be divided into two categories; we will describe first those which are relevant to the assessment of the material's potential performance in a device, and secondly those experiments which are aimed at elucidating the electronic processes which occur during coloration and bleaching.

In general, the materials, as grown, are neither photochromic nor cathodochromic. They require to be sensitized by a thermal treatment, which usually consists of firing in a hydrogen atmosphere, although some sensitization can be achieved by heating under vacuum or in an inert gas. We have found empirically that there is a wide range of temperatures and times which will produce the activation, but there is an optimum treatment which produces the greatest sensitization and this appears to be different for the various types of sodalite. A typical treatment which has been used to sensitize many of our samples is to heat to 800°C in an atmosphere of flowing hydrogen for about thirty minutes.

3.1. *Device Studies*

Probably the most important parameter which has to be determined initially is the sensitivity of the material to coloration by an electron beam, since this determines the writing speed of the display device. This was measured using a demountable c.r.t. which, when optimally adjusted, was capable of producing a 50 μm diameter spot with a current density of 0·5 A/cm^2. Powdered specimens were deposited on a pyrex glass slide to form a uniform layer, typically 50 to 100 μm thick, using the standard settling technique for depositing c.r.t. phosphor screens. The pyrex slide was coated with a layer of transparent conducting tin oxide to give an earth return for the bombarding electrons which thereby prevented the sodalite from becoming charged. Lines of coloration were drawn on the specimen by magnetically deflecting the beam at various rates, and the optical density of the lines was measured by imaging each line via a mirror and a good quality *f*/2·8 lens on the entrance slit of a grating monochromator, which fed a photo-multiplier with an S20 photocathode. The profile of the coloration could be determined by mechanically scanning the image of the coloured line across the entrance slit of the monochromator. Measurements could be made with both reflected and transmitted illumination, and the rate of decay under a known illumination intensity could be determined *in situ*.

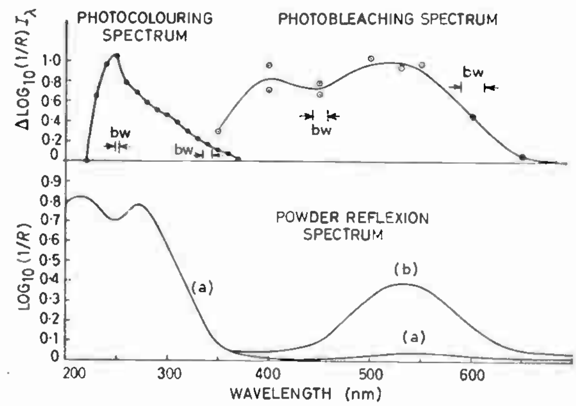


Fig. 3. Optical spectra obtained from powdered chloro-sodalites: the lower half shows the reflexion spectrum in (a) the uncoloured state, and (b) the coloured state; the upper half shows the photo-colouring spectrum on the left, and the photo-bleaching spectrum on the right—both are plotted as the change of the reflectance at the peak of the F-band at 533 nm as a function of the wavelength of the incident light, each point being normalized for a constant incident energy.

Results for a typical chloro-sodalite containing sulphur are shown in Fig. 2. The measurements were obtained with an accelerating potential of 20 kV and with the electron beam focused to give an approximately Gaussian spot of diameter 50 μm, having a current density of 0·47 A/cm^2 at its centre. The plot relates the contrast,† measured in transmission at the centre of the drawn line and at the peak of the spectral absorption band, to the writing speed, expressed in spot diameters per second, or cm/s for a 50 μm spot diameter. The shaded regions represent the spread of results obtained. The relation between contrast and logarithmic writing speed is approximately linear over several decades, and acceptable contrasts can be achieved at a writing speed in excess of 10^5 spot diameters per second. Phillips and Kiss found slightly greater sensitivity using sealed-off tubes and observing the reflected light.[4] We were able to obtain a maximum contrast of between 70 and 80%, after which further bombardment produced a more permanent coloration which could not be erased by light, but which could be removed by heating.

The shape of the reflexion spectrum for a typical chloro-sodalite is shown in the lower half of Fig. 3. Curve (a) corresponds to the uncoloured state, and curve (b) to the coloured state. The reflexion spectrum peaks at 530 nm and is about 120 nm wide at room temperature. Note the spectrum in the region below

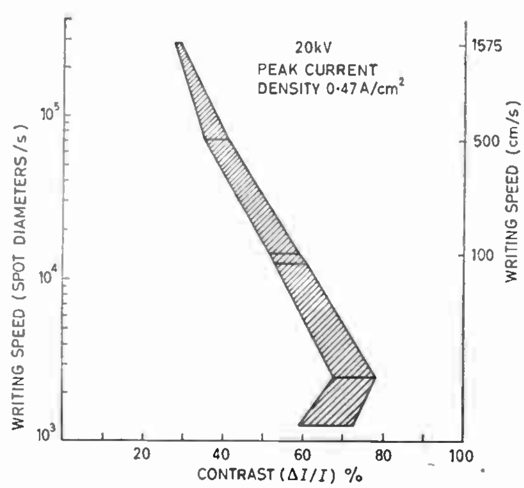


Fig. 2. The coloration rate of a sample of chloro-sodalite by a 20 kV electron beam.

† Contrast is defined as the change in light intensity from the screen expressed as a fraction of the intensity in the uncoloured state.

350 nm, which is unchanged during the coloration process; the peak of the photocolouring spectrum, which is shown in the upper half of Fig. 3, occurs in this region. The photobleaching spectrum, also shown in the upper half of Fig. 3, coincides largely with the main coloration band, although there is also a subsidiary peak at 400 μm. Detailed quantitative measurements of the rate of bleaching have not yet been made, but a preliminary experiment on a chloro-sodalite indicates that with an intensity of 5000 ft candles the coloration can be reduced to about 10% of its initial value within 4 seconds, and with 500 ft candles, the time taken is about 45 seconds. These times are somewhat shorter than those found by Phillips and Kiss. Application of heat can also be used to remove the coloration; a temperature of 150 to 200°C is sufficient to cause erasure within a few seconds. This can be done simply by passing a current through the tin oxide coating. However, this is a rather unsatisfactory solution since, although the specimen can be heated rapidly, the substrate then takes many seconds to return to the ambient temperature so that rapid write–erase sequences cannot be performed in this way unless some technique for extracting the heat is also introduced. Bromo- and iodo-sodalites can be bleached optically more readily than the chloro-sodalites.

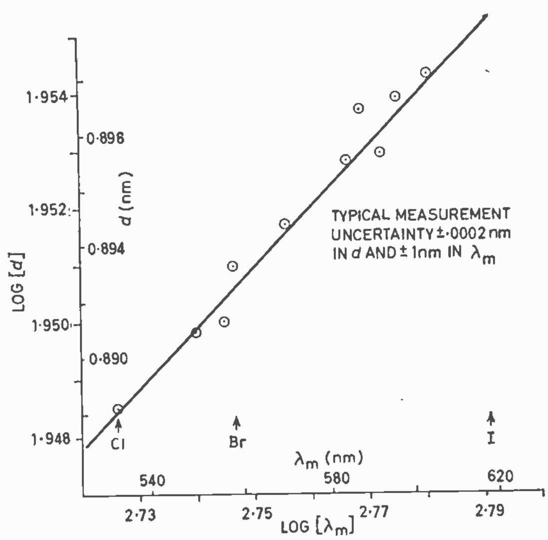


Fig. 4. The dependence of the peak wavelength of the F-band, λ_m, on the lattice parameter, d.

The wavelength of the absorption peak can be moved towards longer wavelengths by replacing the chlorine by bromine or iodine; Fig. 4 shows a plot of the peak wavelength against lattice parameter. Crystals containing mixed halogens each have a unique value for the lattice parameter and the peak can thus be shifted continuously from about 530 nm to 600 nm. Our attempts so far to introduce fluorine have not been successful, but this would be expected to move the peak to shorter wavelengths. There is thus the possibility of being able to make displays with different colours, or by mixing a suitable combination of powders, to produce a black on white display.

Since the coloration arises from entities on an atomic scale, one would expect the resolution to be limited only by the particle size or by the electron beam. We have not yet produced single crystals of sodalite which are large enough to check this, but our earlier work with LaF_3 and CaF_2 showed that it is possible to draw resolvable lines which are only 0·5 μm apart using a scanning electron microscope.

An undesirable property of the present materials is that they fatigue after repeated cycles of coloration and bleaching, in common with many other photochromic materials. However, the rate at which the fatigue occurs seems to be much less than in many photochromic systems, and is less severe than in the KCl previously used in dark trace tubes. Nevertheless it is sufficiently great to reduce the life of working devices to well below the figure of 1000 hours which is often taken to be a desirable target. There appear two forms of fatigue; one is the gradual build-up of a background coloration which bleaches optically much less readily than the normal coloration, but which decays thermally over a period of a day or so, and the other is a decrease in the rate of coloration for a given dose of light or electrons. The decrease in sensitivity is typically 10 to 20% in our powders but in some specimens a decrease of over 50% has been observed; the desensitization is accompanied by changes in the u.v. absorption and visible fluorescence spectra, which will be described in more detail later. The build-up of background coloration is more generally observed and is the most serious type of fatigue. It is not possible to quantify the fatigue process precisely since it seems to depend upon the exact chemical composition of the material, its heat treatment, the depth of coloration, and the number of reversals. Therefore the rate of fatigue of an actual device will depend upon its operating conditions, but the life of the tube at present lies in the region between tens and hundreds of hours.

3.2. *Basic Physics Studies*

The coloration produced by ultra-violet light, by X-rays and by electron beam bombardment seems to be due to the formation of F centres, and the evidence for this comes most strongly from electron spin resonance studies. If the Cl atom at the centre of the cage is missing, the vacancy acts as a trap for electrons in a similar manner to the widely studied F centre in

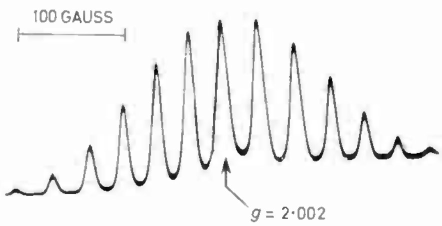


Fig. 5. The X-band e.s.r. spectrum of the F centre resonance.

the alkali halides. An electron trapped in the potential well interacts most strongly with its immediate neighbours, which are the four Na^+ ions. These each have a nuclear spin of 3/2, so that the total effective spin $I = 6$, and thus the electron resonance line is split into $2I+1 = 13$ components. Unlike the F centre resonances in the alkali halides these components are clearly resolved, and provide an unambiguous identification of the F centre. Figure 5 shows a typical e.s.r. spectrum taken at 1·4°K using an X-band superheterodyne spectrometer. The intensity distribution approximates to a Gaussian envelope and reflects the number of different ways in which the four nuclear spins can be arranged to give the resultant magnetic field corresponding to each component. Since the spectrum is isotropic, at least to first order, the spectrum can be readily observed in powdered samples. Hodgson *et al.*[6] first described this spectrum and found a linear correlation between the relative strength of the e.s.r. signal and the strength of the visible absorption peak, and it is thus a reasonable deduction to ascribe both the e.s.r. absorption and the optical absorption to the F centre. We find that such an e.s.r. resonance occurs in all samples of photochromic sodalites; the major difference between different formulations is the value of the hyperfine spacing. This can be correlated with the lattice constant when different ions are substituted into a given lattice site. For example, by substituting into the Cl site, the hyperfine constant A, expressed in gauss, can be related to the lattice parameter d, in nm, by the empirical equation $A = 464{\cdot}8 - 488{\cdot}4\,d$, where A varies from 31·0 gauss in pure chloro-sodalite to 24·7 gauss for a fully substituted bromo-sodalite. Substitution into other lattice sites produces similar relations, but with different numerical constants. A more detailed discussion on this topic will be published elsewhere.

Although the model of the F centre is almost certainly correct, it would be desirable to have additional evidence from the substitution of the Na^+ ions by other similar ions, preferably with a different nuclear spin. We have attempted to exchange the sodium ions by Li ($I = 3/2$) and Ag ($I = \frac{1}{2}$, two isotopes) using the technique described by Barrer and Raitt[12] for ultramarine. Although the chemical evidence indicated that some exchange had occurred, the resultant products were neither photochromic, nor were any e.s.r. resonances observable from F centres. Some structureless e.s.r. lines were observed in all these samples, but these have not yet been identified. We conclude that ionic exchange is not as straightforward a process as the chemical evidence might suggest.

The mechanism by which the electron is transferred to the F centre during coloration is much less well understood. Whereas a photochromic response is obtained only with sulphur-doped material we find, however, that sulphur is not essential to the cathodochromic response, although the latter is improved when sulphur is present. F centres are generated in undoped sodalite by exposure to high energy electrons or X-rays but not to u.v. radiation of wavelengths above about 200 nm. One might expect a response to higher energy u.v. radiation in the fundamental band edge region, as is found in the alkali halides,[13] but this has not yet been investigated. It is thus fairly certain that sulphur in some form is an electron donor, and by introducing absorption bands in the near u.v. region it acts to generate the photochromic response. Exactly where the sulphur is located in the lattice with respect to the F centre, and in what form, is not at present known. The optical spectra which provide evidence that sulphur is responsible for the photochromic sensitivity are shown in Figs. 3 and 6. Figure 3 shows that the photoexcitation spectrum lies in the wavelength region below 350 nm. Figure 6 shows how the absorption bands in this region depend upon the presence of sulphur; curve (1) was obtained from a chloro-sodalite sample which nominally contained no sulphur. Curve (2) was obtained from an as-grown sample of a chloro-sodalite containing 0·5 at.% of sulphur, and curve (3) was produced by the same sample after heating in a hydrogen atmosphere. The precise distribution of the absorption

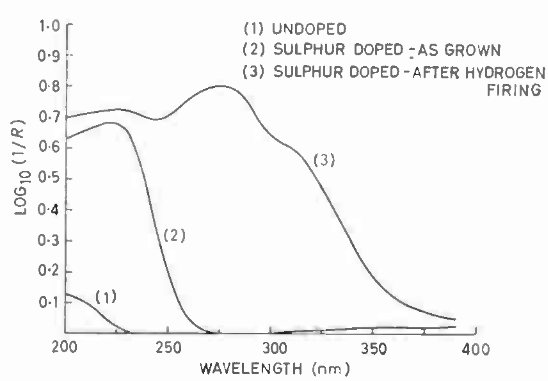


Fig. 6. Variations of the ultra-violet reflexion spectra of chloro-sodalites.

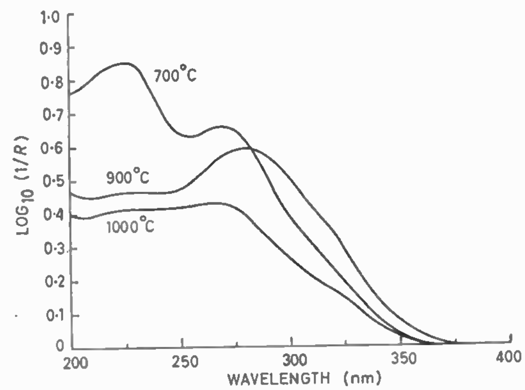


Fig. 7. The effect of hydrogen firing at various temperatures on the u.v. reflexion spectra of a chloro-sodalite.

bands depends upon the heat treatment, and Fig. 7 shows the effect of heating the samples from the same batch of material for one hour at various temperatures. We believe that the heat treatment effects the distribution of sulphur radicals in the material; the evidence that the sulphur exists in radical form comes partly from the e.s.r. spectra and partly from fluorescence studies. McLaughlan and Marshall[14] have made an e.s.r. study of a number of chloro-sodalites, containing high concentrations of sulphur, in an attempt to identify which radicals can be formed. The sulphur radicals observed in other host lattices, in general show anisotropic e.s.r. spectra, and thus, when working with powders, much of the orientational dependent information is lost, particularly as the spectra tend to occur in the same region. However, it proved possible to derive g-values for a number of species from the smeared-out spectra; by comparing these with the values obtained for the species in other host lattices where their assignment could be accurately made by isotopic enrichment and e.n.d.o.r. studies, they were able to infer the existence of several species. The existence of S_3^- radicals was reasonably well established, and spectra corresponding to SSO^- and either SO_3^- or $S_2O_2^-$ radicals were found. However, it will be necessary to make isotopic enrichment studies on sodalite single crystals before these assignments can be properly established.

The existence of the radicals S_2^- and possibly S^{--} can be inferred from optical fluorescence measurements. When sulphur doped chloro-, bromo-, and iodo-sodalites are stimulated by u.v. light or an electron beam a reddish emission is observed under certain conditions. In general, hydrogen-fired material does not show this fluorescence until it has been fatigued by an excessive exposure to u.v., X-ray or electron irradiation; however, sodalites which have been fired in argon instead of hydrogen show the effect immediately. The intensity of the emission depends to some extent upon composition; for example, sodalites doped with tetrasulphide show much more emission than those doped with monosulphide. The argon-fired powders which show the strongest emission have a pale yellow colour, which can be correlated with a reflexion peak in the u.v., and have a very poor photochromic response.

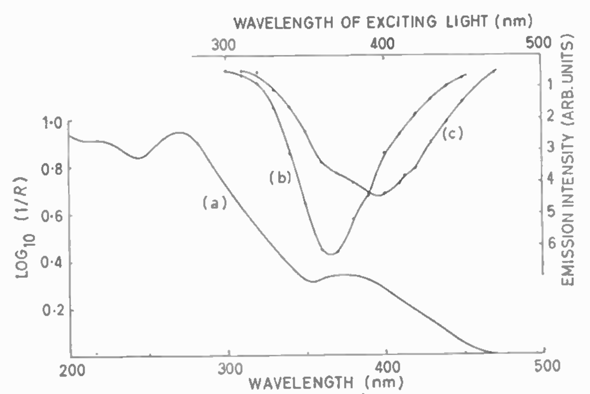


Fig. 9. Ultra-violet optical spectra obtained from the specimen used for Fig. 8, showing the relation of the fluorescence excitation spectra, (b) and (c), to a peak in the reflexion spectrum, (a). Curve (b) is the excitation spectrum of the structureless band (Fig. 8(b)), obtained by monitoring the fluorescence emission above 900 nm, while (c) is the excitation spectrum of the emission in the region of 600 nm, which comes mainly from the S_2^- radical.

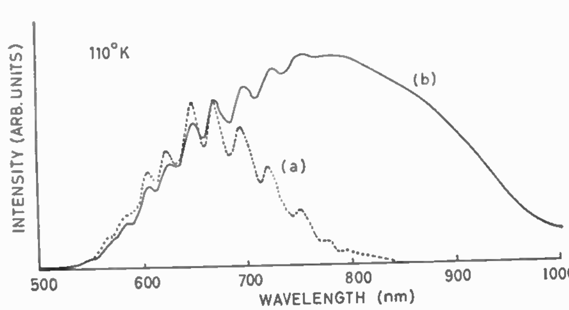


Fig. 8. Photo-emission spectra of an argon fired chloro-sodalite; (a) the emission identified as S_2^- obtained by selective-excitation, (b) the emission obtained by broad-band excitation in the region of 370 nm which shows the S_2^- emission superimposed upon a structureless band.

In general, emission can be distinguished from two distinct species, and these are shown in Fig. 8. One emission, shown in curve (a), shows a banded structure with a separation of 556 cm^{-1} when cooled to about 100°K, whereas the other emission, shown in curve (b), remains structureless and peaks at a longer wavelength. The band structure is typical of a vibrating diatomic molecule and has been assigned to S_2^- in a number of host lattices,[15] and has been observed

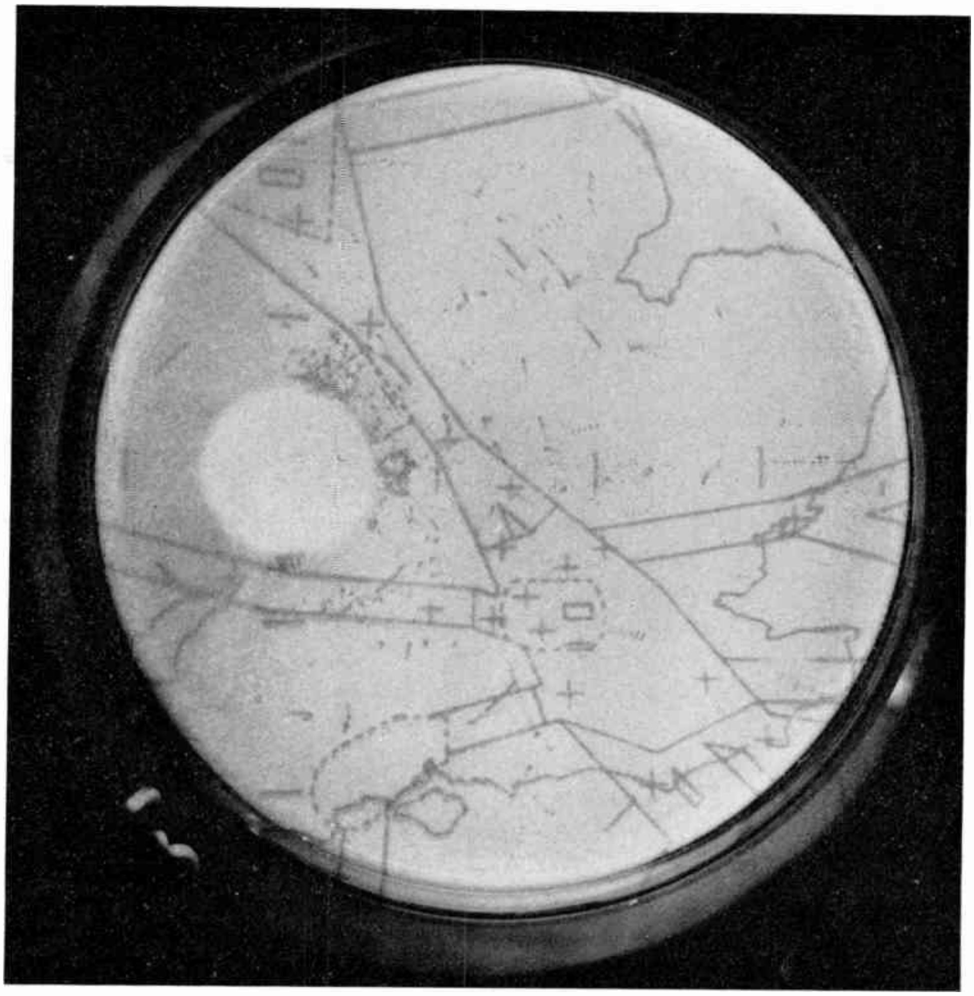


Fig. 10. Radar p.p.i. display on a sodalite tube.

previously in sodalite by Kirk,[8] and by Hodgson *et al.*[6] The structureless emission is similar to that observed by Fischer[16] in KCl which he attributed to S^{--}. These emissions are excited by different wavelengths in the near ultra-violet but their excitation spectra overlap sufficiently to make it difficult to excite them selectively. Figure 9 shows the u.v. reflexion spectrum of a typical argon-fired powder in curve (a), and the excitation spectra of the two fluorescent emissions in curves (b) and (c). The excitation spectra lie near a reflexion peak at 370 nm, which can therefore be attributed to the presence of S_2^- and another radical, possibly S^{--}.

The appearance of strong luminescence seems to be detrimental to the achievement of optimum photochromic sensitivity, and indeed the strength of this emission can be used as an indication of the degree to which hydrogen-fired powders have been fatigued. Although much further work will be required to solve the fatigue problem, it seems likely to us that it is associated with the formation under irradiation of di- and tri-atomic radicals. The question of where the sulphur is located in the lattice still remains to be solved.

4. Device Performance and Conclusions

The simplest type of display which one can envisage consists of a conventional c.r.t. in which the normal phosphor screen is replaced by sodalite, and we have made a number of such tubes and have also evaluated experimental tubes made by Ferranti Ltd. These tubes are particularly suitable for the presentation of radar p.p.i. data, since the fade period under quite bright ambient lighting conditions can be made to be a minute or so. Figure 10 shows a picture of a radar p.p.i. display presented on such a tube and one can readily see the directional trails of aircraft tracks superimposed upon an electronically generated map showing the airways and the coastline of eastern and

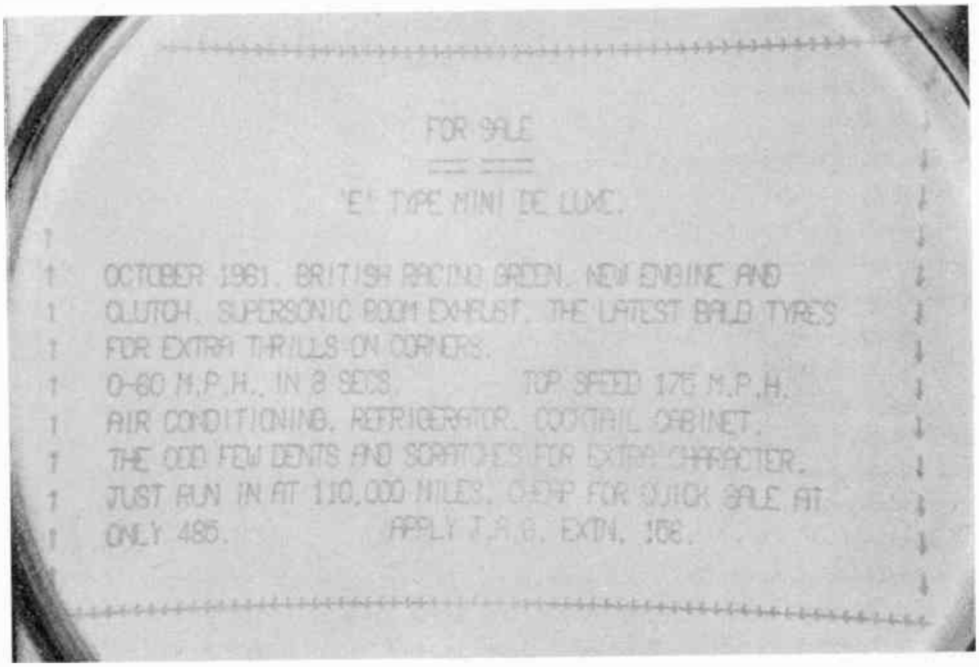


Fig. 11. Alpha-numeric display on a sodalite tube.

southern England. The level of illumination used was great enough to render the display on an ordinary magnesium fluoride long-persistence tube almost invisible. The radar makes four paints per minute, and some of the strong tracks showed up to a dozen spots on the tube, giving the direction and speed of each aircraft. Figure 11 shows the same tube used to display alpha-numeric information, each character being generated by individual strokes from a computer. The display of computer-generated information could be one of the most important applications of the sodalite dark trace tube, particularly if the erase time can be reduced to a fraction of a second.

To conclude, the main advantages and disadvantages of display devices using sodalites may be summarized as follows:

The displayed information can be viewed directly, in ambient lighting, like writing on paper. Since energy is supplied externally, the display can be bright, in contrast to long persistence phosphors where all the emitted energy has to be stored initially in the phosphor.

The material has a memory which is controllable—the decay can be varied by the level of illumination, and by the screen temperature. At present, there is no elegant way of producing a rapid erasure; there is an extra peak in the photobleaching spectrum around 400 nm, which is probably caused by transitions to higher excited states of the F centre, and it may be possible to utilize this, with suitable filtering, to control the erasure without changing the illumination level at the peak of the F band.

Writing rates are already possible which are sufficiently high for device use, although some increase would be desirable.

Resolution is extremely good and no problems are anticipated here.

At present, the occurrence of fatigue would seem to be the main deterrent to the widespread use of these materials. There is a need for more precise knowledge of the conditions under which fatigue occurs, and our present work is aimed at trying to understand the mechanism of this process. If this problem can be solved, we feel confident that sodalite dark trace tubes will make an important contribution in the storage display field.

5. Acknowledgments

The work described above has benefited greatly from the assistance of C. R. Johnson, H. Evans and Mrs. J. Constant and we would like to express our thanks for their endeavours. The photographs shown in Figs 10 and 11 were kindly obtained for us by C. F. Clutterbuck.

The paper is published with the permission of The Controller, H.M. Stationery Office.

6. References

1. Rosenthal, A. H., 'A system of large-screen television reception based on certain electron phenomena in crystals', *Proc. I.R.E.*, **28**, pp. 203–212, 1940.
2. King, P. G. R., 'The skiatron or dark-trace tube', *J. Instn Elect. Engrs*, **93**, Pt. 3A, pp. 171–172, 1946; King, P. G. R. and Gittings, J. E., 'The skiatron or daık-trace tube', *ibid.*, pp. 822–831, 1946.
3. Leverenz, H. H. W., 'Luminescence of Solids, pp. 310–312. (Wiley, New York, 1950); Soller, T., Starr, M. A. and Valley, G. E., 'Cathode Ray Tube Displays', pp. 664–705, MIT Radiation Laboratory Series Vol. 22. (McGraw-Hill, New York, 1948).
4. Phillips, W. and Kiss, Z. J., 'Photo-erasable dark-trace cathode-ray storage tube', *Proc. I.E.E.E.*, **56**, pp. 2072–2073, 1968.
5. Ivey, H. F., 'Tenebrescent or cathodochromic sodalite in cathode-ray storage tubes', *Proc. I.E.E.E.*, **57**, pp. 853–854, 1969.
6. Hodgson, W. G., Brinen, J. S. and Williams, E. F., 'Electron spin resonance investigation of photochromic sodalites', *J. Chem. Phys.*, **47**, pp. 3719–3723, 1967; Williams, E. F., Hodgson, W. G. and Brinen, J. S., 'Synthetic photochromic sodalite', *J. Amer. Ceram. Soc.*, **52**, pp. 139–145, 1969.
7. Barrer, R. M., 'Crystalline Ion-Exchangers' (Tilden Lecture), *Proc. Chem. Soc.*, pp. 99–112, 1958.
8. Kirk, R. D., 'The luminescence and tenebrescence of natural and synthetic sodalite', *Amer. Mineralogist*, **40**, pp. 22–31, 1955; 'Role of sulphur in the luminescence and colouration of some aluminosilicates', *J. Electrochem. Soc.*, **101**, p. 461, 1954.
9. Medved, D. B., 'Optical properties of natural and synthetic hackmanite', *J. Chem. Phys.*, **21**, pp. 1309–1310, 1953; 'Hackmanite and its tenebrescent properties', *Amer. Mineralogist*, **39**, pp. 615–629, 1954.
10. Radler, R. W., 'Investigation of Inorganic Phototropic Materials as a Bi-optic Element Applicable in High Density Storage Computer Memories', AD277793, Polarcoat Inc., Blue Ash, Ohio, April 1962; Radler, R. W. and Chenot, O., 'Synthesis of Inorganic Phototropic Materials for High Density Computer Memory Applications', AD608426, Polarcoat Inc., October 1964.
11. Ballman, A. A. and Laudise, R. A., Ed. J. J. Gilman, p. 232. 'Art and Science of Growing Crystals'. (Wiley, New York, 1965).
12. Barrer, R. M. and Raitt, J. S., 'Ion exchange in ultramarine', *J. Chem. Soc.*, pp. 4641–4651, 1954.
13. Goldstein, F. T., 'F centre production by ultraviolet irradiation of KCl and NaCl', *Solid State Commun.*, **6**, pp. 565–567, 1968, and 'F centre formation by fundamental absorption in KI', *Phys. Stat. Sol.*, **20**, pp. 379–393, 1967.
14. McLaughlan, S. D. and Marshall, D. J., 'Paramagnetic resonance of sulphur radicals in synthetic sodalites', *J. Phys. Chem.*, **74**, p. 1359, 1970.
15. Schulman, J. H. and Kirk, R. D., 'Luminescent sulphur centres in alkali halides and other inorganic solids', *Solid State Commun.*, **2**, pp. 105–108, 1964.
16. Fischer, F., 'Farbzentren der chalkogene in alkalihalogenidkristallen', *Z. Naturwiss.*, **12**, pp. 313–318, 1967.

Manuscript first received by the Institution on 11th June 1969 and in final form on 19th August 1969. (Paper No. 1329/CC83.)

STANDARD FREQUENCY TRANSMISSIONS—June 1970

(Communication from the National Physical Laboratory)

June 1970	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)		June 1970	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	− 300·0	− 0·1	+ 0·1	621	586·1	**17**	− 300·0	− 0·1	+ 0 1	622	590·6
2	− 299·9	+ 0·1	+ 0·1	620	585·3	**18**	− 300·0	− 0·1	+ 0·1	622	592·0
3	− 300·0	− 0·1	+ 0·1	620	585·8	**19**	− 299·9	0	+ 0·1	615	591·8
4	− 300·1	0	+ 0·1	621	585·9	**20**	− 300·0	0	+ 0·1	615	591·9
5	− 300·1	0	+ 0·1	622	585·9	**21**	− 300·0	0	+ 0·1	615	592·0
6	− 300·1	− 0·1	+ 0·1	623	586·9	**22**	− 300·0	0	+ 0·1	615	592·4
7	− 300·1	− 0·1	+ 0·1	624	588·1	**23**	− 300·0	0	+ 0·1	615	592·6
8	− 300·1	− 0·1	+ 0·1	625	589·1	**24**	− 300·0	0	0	615	592·8
9	− 300·0	+ 0·1	+ 0·1	625	588·6	**25**	− 300·0	0	+ 0.1	615	593·0
10	− 300·1	− 0·2	+ 0·1	626	590·8	**26**	− 300·0	− 0·1	+ 0·1	615	593·6
11	− 300·0	0	+ 0·1	626	591·2	**27**	− 300·0	0	+ 0·1	615	593·8
12	− 299·9	0	+ 0·1	625	591·4	**28**	− 300·0	+ 0·1	0	615	592·6
13	− 299·8	0	+ 0·1	623	590·9	**29**	− 299·9	0	+ 0·1	614	592·4
14	− 300·0	+ 0·1	+ 0·1	623	589·8	**30**	− 300·0	− 0.1	+ 0·1	614	593·4
15	− 299·9	0	+ 0·1	622	589·7						
16	− 300·0	0	+ 0·1	622	590·0						

All measurements in terms of H.P. Caesium Standard No. 334, which agrees with the N.P.L. Caesium Standard to 1 part in 10^{11}.
*** Relative to UTC Scale; (UTC_{NPL} − Station) = + 500 at 1500 UT 31st December 1968.**
†Relative to AT Scale; (AT_{NPL} − Station) = + 468·6 at 1500 UT 31st December 1968.

Contributors to this Issue from Overseas

Dr. Bernhard Koch received his doctorate in 1938 in natural sciences. He then carried out research work in the far infra-red to short millimetre waveband at the Deutsche Versuchsanstalt fur Luftfahrt. After a short period of optical instrumentation work in industry (Askania), he joined in 1940 the Technical Academy of the German Luftwaffe at Berlin-Gatow, where he did spectroscopic work related to atmospheric optical propagation problems, combustion and detonation plasmaphysics. In 1945 he became project leader in what is now called the French-German Research Laboratory of Saint-Louis. He at present leads research on applications of microwave and laser physics to ballistics.

Philippe Rateau obtained the Licencie és Science Appliquées in electronics in 1967 at the University of Strasbourg. Since 1967 he has worked as a scientist at the French-German Research Institute, his specialization being telemetry in connexion with vehicles having high acceleration.

Gûnter Schultze obtained his Dipl.-Ing. in 1939 at the 'Gauss' School of Engineering in Berlin. From 1942 to 1945 he was at the Air College in Berlin where he carried out investigations into short-wave propagation over long distances. He went to St. Louis after the war and since 1962 has held the rank of scientist at the French-German Research Institute. His field of activities has included the development of Doppler radars for trajectory measurements and investigation of projectile-borne telemetering devices submitted to high accelerations.

Dr. Heinz-Dieter vom Stein received the Dipl.-Ing. degree in electrical engineering at the Technische Hochschule, Aachen, in 1961 and in 1966 gained the Dr.-Ing. degree. Since 1961 he has been with the French-German Research Laboratory, being active in the field of microwaves and coherent optics.

Professor M. N. Srikanta Swamy received his bachelor's degree in mathematics in 1954 from Mysore University, and his M.Sc. and Ph.D. degrees in electrical engineering from the University of Saskatchewan, Canada, in 1960 and 1963, respectively. He worked as a senior research assistant at the Indian Institute of Science until 1959, when he moved to Canada for graduate study at the University of Saskatchewan. In 1963 he returned to India to work at the Indian Institute of Technology, Madras. During 1964–1965, he was an assistant professor of mathematics at the University of Saskatchewan. He has also taught, as professor of electrical engineering, at Nova Scotia Technical College, Halifax and the University of Calgary. Professor Swamy is now chairman of the Electrical Engineering Department at Sir George Williams University, Montreal. He has published a number of papers on number theory, semiconductor circuits, control systems, and network theory.

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

Dr. J. C. Giguere received his B.Eng. degree in electrical engineering from McGill University, Montreal, and, following three years in the Royal Canadian Air Force, went to the Nova Scotia Technical College, Halifax, obtaining M.Eng. and Ph.D. degrees. He was appointed a lecturer in the Department of Electrical Engineering at the College in 1965 and became an assistant professor in January, 1969. He moved to Sir George Williams University, Montreal, in July 1969 where he is an assistant professor of electrical engineering.

U.D.C. 621.317.725

Digital Carry Applied to Successive Approximation Digital Voltmeters

By

J. R. PEARCE, B.Sc., C.Eng., M.I.E.E.†

J. BLOOMFIELD†

and

U. QURESHI†

Reprinted from the Proceedings of the I.E.R.E Conference on 'Digital Methods of Measurement' held in the University of Kent at Canterbury on 23rd to 25th July 1969.

Successive approximation encounters difficulties, when used in high resolution voltmeters, due to the settling time of the decision loop. Solutions so far have either slowed the reading rate, or have significantly increased complication and cost. Digital carry solves the problem both simply and inexpensively. It is applied to a voltmeter which, at 50 readings per second, has a scale length of 100,000.

1. Introduction

Digital voltmeters and analogue-to-digital converters which employ successive approximation have two major advantages over instruments based on alternative techniques, namely speed and precision. Although other kinds of voltmeter give better performance in some respects, it appears to be acknowledged that successive approximation is capable of giving the most accurate results at the highest speed. An a.-to-d. converter with an accuracy of 0·1% can complete a reading in a few microseconds, while accuracies of 0·002% take a little longer—perhaps several milliseconds.

However, all techniques have their limitations, and successive approximation has, in the past, given problems when pushed to very high accuracy and resolution. Digital carry provides a modified approach which is inexpensive and simple. It makes possible a d.v.m. which, while retaining high speed operation, has an accuracy which can be verified only in the best equipped standards laboratories.

2. The Basic System

The principle of successive approximation is illustrated by the schematic diagram of Fig. 1. The unknown d.c. signal is applied to one input of a comparator amplifier. The other input is a known voltage derived from a reference supply via a digital divider. The divider is stepped through a fixed sequence so that precise increments of reference voltage are applied to the comparator.

The increments are offered in a binary coded decimal (b.c.d.) sequence, starting with the largest. As each increment is tried the comparator signals 'too much' or 'too little' to the decision circuits, which

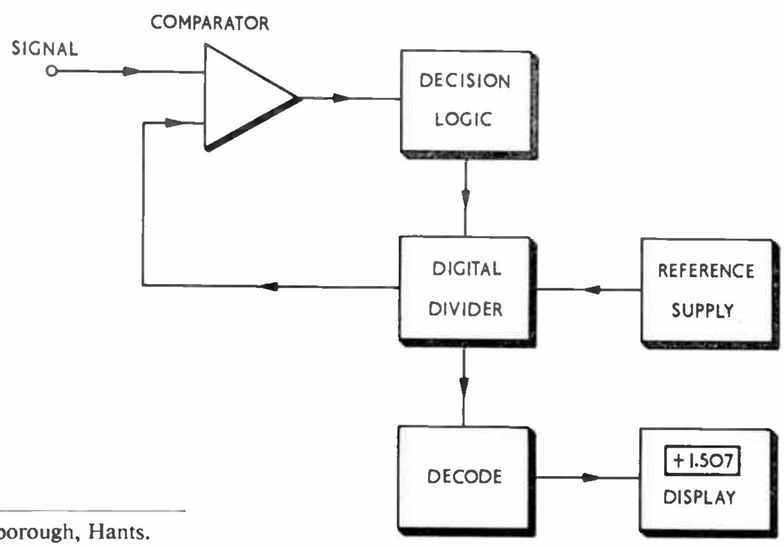


Fig. 1. The basic system.

† The Solartron Electronic Group, Farnborough, Hants.

then cause that increment to be either rejected or retained. At the end of the sequence the divider has built up a potential which balances the input signal. The state of the divider is then decoded and is fed to the display to give a decimal reading.

Consider a voltmeter having a maximum display of 999. On a range which is nominally 1 V maximum, the instrument will read up to 0·999 V. Each of the three decades can be built up in b.c.d., a code which is frequently used being 8, 4, 2, 1. In this case the voltages are:

800 mV	80 mV	8 mV
400 mV	40 mV	4 mV
200 mV	20 mV	2 mV
100 mV	10 mV	1 mV

and suitable combinations of these values can be made to build up any voltage from zero to 999 mV with a resolution of 1 mV. Whatever the value of the unknown input the divider starts by trying the largest increment and works down to the smallest. For example, with a signal of 0·798 V the first trial of 800 mV is found to be too big so it is switched out. The next is 400 mV, the decision is 'too small' so this is left in. Similarly the addition of 200 mV and 100 mV give a total which is still too small and they are left in. The other decades are tried in turn, a decision being taken on every event. Table 1 shows the result of each decision, and a graph of the divider output building up is shown in Fig. 2.

Table 1

Signal	7	9	8
Decade decisions	~~8~~	8	8
	4	~~4~~	~~4~~
	2	~~2~~	~~2~~
	1	1	~~1~~
Reading	7	9	8

3. Decision Making

The decision to accept or reject a given increment is irrevocable, hence it must be correct. For fast operation it is desirable to take each decision (each step in Fig. 2) in, say, 1 μs. Bearing this in mind, consider the decision required on the first (largest) increment, which, in the example being pursued, is 800 mV. For input signals of 0·799 V or less it must be rejected, while for inputs of 0·800 V or more it must be accepted. Indeed, for a good voltmeter, even greater discrimination is desirable. Because the input is an analogue voltage, and the reading is a digital approximation to the input, it is desirable to base decisions on a decade which is one order more sensitive than the display. Thus, for the 800 mV increment the decisions should be:

Signal	Decision	Display
0·799(4) V	Reject 800 mV	0·799
0·799(5) V	Accept 800 mV	0·800

It is seen that, at the time of decision, the reference increment and comparator loop must have settled to 1 part in 8000 in something less than 1 μs. In an a.-to-d. converter or a voltmeter having a sensitivity of 1 mV this speed can be achieved. However, in a much more sensitive voltmeter with a scale length of 99999 and a sensitivity of 10 μV, the comparator should have a discrimination of 1 μV. The comparator gain must then be at least 10^6 and the major increments must be defined by wirewound resistors. As each increment is applied the switches (usually transistors) must settle and there are time-constants associated with the divider networks and the wiring to them. The comparator will initially be in overload in one direction so it should be given time to come out of overload, perhaps overshoot, and pass valid information to the decision circuits. Hence the decision time to accept or reject 80000(0) when the unknown is 79999(4) extends to several hundred microseconds.

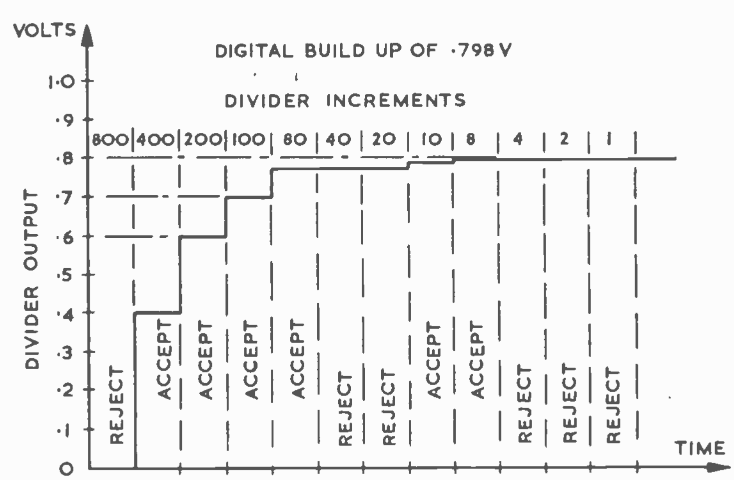


Fig. 2. Divider output build-up.

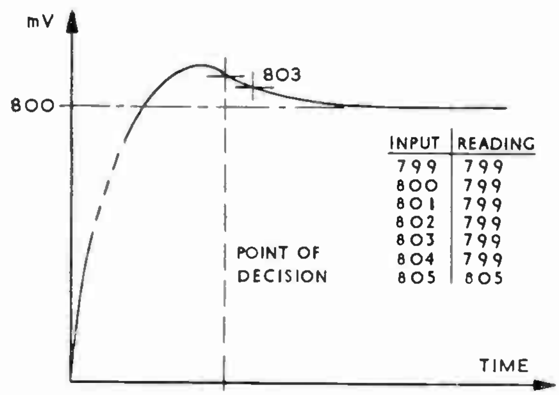


Fig. 3. Mechanism of miscounting.

4. Miscounting

If the loop has not settled at the time a decision is made, reading errors may occur. The effect is known as miscounting and takes place near the major decision points. That is to say, miscounting will tend to occur, with an 8, 4, 2, 1 code, around readings of 80,000, or 40,000. The mechanism is illustrated by Fig. 3, in which a full scale of 999 is again used for clarity.

The input signal is 803 mV, so acceptance of the 800 mV increment would be expected. However, because the comparator loop has overshot at the point of decision the answer is 'too much' and the increment is switched out. A short while later, when things have settled, the answer is the correct reply of 'too little,' but by then it is also 'too late' because the decision has been taken! As a result when the subsequent increments are offered, 400, 200 and 100, they are all accepted, as are those in the lower decades, and the instrument finishes the sequence by displaying 799.

A similar result occurs due to undershoot, as shown in Fig. 4. From the moment the increment is switched in the comparator is signalling 'too little', and is still doing so at the decision point. A short while later the 800 increment has reached full value, and thereafter all choices are rejected. Hence, with a slowly-increasing input signal, the voltmeter will read correctly until an input such as 797 is reached. The reading then jumps to 800, where it remains until the input reaches 801, the intervening digits being missed.

5. Some Solutions

The obvious way to overcome the problem is to allow enough time for settling before taking a decision. The total time for digitizing is then considerably extended, particularly if allowance is made for variations in circuit parameters. The speed of reading then becomes unacceptable for fast data acquisition systems.

Another approach uses the fact that only the major increments need to settle with great precision. In the less significant decades the decisions may be taken rapidly, while longer times are allowed for the big decisions. This complicates the control operation since, instead of stepping the reference divider at equal time intervals, a variable arrangement must be used. Even then a doubt may exist whether enough time has been allowed for worst case parameters.

A further arrangement employs successive approximation to give a rough balance against the input signal, and a different technique, such as voltage-to-frequency conversion, is brought in to act as a fine adjustment. While capable of great precision, the method is very complex and relatively slow.

6. Digital Carry

An approach which avoids these disadvantages is digital carry, the essence of which is deliberately to allow errors in decision at the major choices and to correct for these errors in the less significant decades. If the input signal is close to a major increment the instrument is caused to make a deliberate mistake! Then in the lower decades, additional increments are permitted which are carried forward to the upper decades in the final reading.

As explained earlier, the technique is only necessary for voltmeters with very high resolution, since the miscount problem is not exhibited with low-resolution instruments. It has been employed in a voltmeter having a maximum reading of 101999, but the explanation is simplified if a 999 scale is considered.

When offering the increments in the first (most significant) decade the digital divider is biased to give outputs that are slightly in excess of nominal. This bias is switched out before offering the increments in the second and third decades. Taking again the signal of 803 mV the decisions are set out in Table 2.

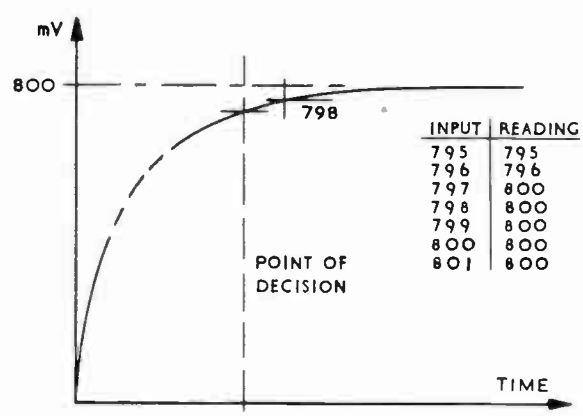


Fig. 4. Mechanism of undershoot.

Table 2

Signal	8	0	3
Decade decisions	~~8~~	8	~~8~~
	4	~~4~~	~~4~~
	2	2	2
	1	~~1~~	1
Divider output	7	10	3
Reading	8	0	3

The 800 mV choice is rejected because, with bias, it is 'too much'. The offers of 400, 200 and 100 are then all accepted. Before bringing in the second decade the small bias is removed, so the contribution so far is 700, giving a remainder of 103. In the second decade the 80 and 20 increments are accepted, leaving 3 (2+1) to be selected in the third decade. This is then the state of the digital divider at the end of the sequence. In decoding the result from b.c.d. to decimal it is only necessary to carry the 10 in the second decade to the first decade.

In this example, the bias applied to the first decade is only 7 or 8 mV, so when it is removed the divider output settles to the requisite accuracy very rapidly. However, it may be that, even with bias, the 800 mV choice undershoots so much that it is accepted. This makes no difference to the result because the choices will now follow the sequence shown in Table 3.

Table 3

Signal	8	0	3
Decade decisions	8	~~8~~	~~8~~
	~~4~~	~~4~~	~~4~~
	~~2~~	~~2~~	2
	~~1~~	~~1~~	1
Divider output	8	0	3
Reading	8	0	3

Hence, with digital carry, it is not possible to predict exactly which choices will be made when the signal is close to a major decision point. On the other hand, conventional successive approximation must make a fixed array of choices for a given input, and if this goes wrong then the answer is wrong.

It will have been realized that an 8, 4, 2, 1 code can provide a maximum of 15 in any decade. Thus, in normal successive approximation, each decade is limited to 9, either by inhibiting the 4 and the 2 if the 8 is selected, or by using another code such as 4, 2, 2, 1. With digital carry, the carry decade must be capable of exceeding 9, and it is convenient to permit a maximum of 11.

Table 4

	8	8	8
	4*	4*	4*
	2*	2	2*
	1	1	1
	9	11	9
1	0	1	9

* Inhibited if the 8 is selected in that decade.

Table 4 shows that total increments of 1019 are available for a full scale of 999. Thus there is an excess of 20 'bits' contributed by the second decade; consequently the major increments could overshoot by up to 20 bits at the point of decision and the error would be corrected. In order to allow for undershoot as well as overshoot the increments would be biased by +10 bits, thus allowing errors of ± 10. For a discrimination one order better than the display, the original requirement was 1 part in 8000 in 1 μs. But the 800 increment may now be 810, or only 790, at the point of decision. Hence the permissible error is now 100 parts in 8000.

Of course in the long run, the major choices must settle to their nominal, unbiased values. The first decision in the second decade (80 in Table 5) must be correct or there will be errors in reading other signals: 879 for example. Hence increments accepted in the first decade must settle before the 5th decision. Therefore, by using digital carry, the available settling time for the 800 increment is increased by a factor of five.

In the example used, the technique has reduced the precision required of the most significant choice from 1 in 8000 to 1 in 80. However, the second decade must still achieve accuracy and settling time

Table 5

BIAS IN				BIAS OUT							
	*	*			*				*	*	
800	400	200	100	80	40	20	10	8	4	2	1

DECISIONS 1 2 3 4 5 6 7 8 9 10 11 12

* Inhibited if the 8 is selected in that decade.

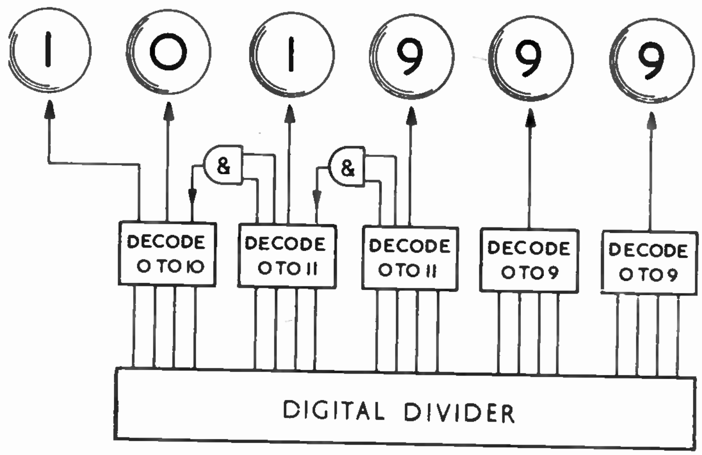


Fig. 5. Arrangement of divider for decoding.

which is unchanged—namely 1 part in 800 in 1 μs. In effect a 3-window digital voltmeter is produced in which the performance of the decision loop is only that required of a 2-window instrument.

7. Practical Application

It was decided to extend an existing voltmeter design to give a scale of 99999. Using digital carry the decision performance necessary would be only that required by a 999 voltmeter, and the operating speed of 50 readings per second would be retained. The arrangement of the decades is shown in Table 6.

Table 6

		Bias in		Bias out		
		4	8	8	8	8
		2	4*	4*	4*	4*
		2	2	2	2*	2*
		1	1	1	1	1
Bits per decade		9	11	11	9	9
Total available	1	0	2	1	9	9

* Inhibited if the 8 is selected.

When the voltmeter reading is to be transmitted to other equipment in b.c.d. an 8, 4, 2, 1 code is usually required. Thus the use of this code for the steps in each decade avoids further encoding, as well as providing the surplus digits required for digital carry. However, there are some advantages in choosing a 4, 2, 2, 1 code for the most significant decade:

(a) the major choice is half that in 8, 4, 2, 1 so this increment will settle more rapidly;

(b) the voltage of the reference supply, which must at least equal the total 'bits', can be lower with 4, 2, 2, 1;

(c) the divider resistors in this decade must provide the greatest precision, and they can be matched more closely in a 4, 2, 2, 1 code because there is a smaller spread of values.

The total available gives basic ranges (without input attenuation) of, nominally, 1 V and 10 V with sensitivities of 10 μV and 100 μV. Furthermore, the ability to read just over 1 V means that the instrument can be calibrated directly against a standard cell with a resolution of 10 parts in 10^6.

In operation the choices in the top two decades are biased to be very slightly greater than nominal. (On the 1 V range the bias is approximately 80 μV.) Hence, for signals close to the major decision points, the totals built up in these two decades will tend to be too small. For the three lower decades the bias is removed. At the end of the sequence, if the second or third decades exceed 9, the decoding from b.c.d. to decimal carries the 10 to the next higher decade.

Table 7 shows that a correct answer can be obtained in three possible ways. The instrument selects appropriate increments dependent on whether the early decisions are found to be correct or erroneous.

Table 7

Input	4	0	0	1	3	4	0	0	1	3	4	0	0	1	3
	4	8	8	8	8	4	8	8	8	8	4	8	8	8	8
	2	4	4	4	4	2	4	4	4	4	2	4	4	4	4
	2	2	2	2	2	2	2	2	2	2	2	2	2	2	2
	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1
	4	0	0	1	3	3	10	0	1	3	3	9	10	1	3
Reading	4	0	0	1	3	4	0	0	1	3	4	0	0	1	3

8. Decoding

The carry involves only a slight modification to conventional decoding. Figure 5 shows the schematic arrangement. The display tube decodes now have to cover 0 to 11, 0 to 10 and 0 to 9, while a carry into the 1st and 2nd decades is produced by AND gates which detect the presence of 8 AND 2 in the lower decade.

9. Conclusions

The usual method of 'stretching' an existing design to higher orders of accuracy is either to carry out expensive selection of components, or to re-engineer the equipment into a more complex, and probably less reliable, product. The problems described in this paper are analogue, but they have been overcome by a simple digital solution. In this case a digital voltmeter which had a scale of 30000 has been extended to a scale length of 100000. The new product is easier to manufacture than the old one while the increase in component cost is minimal.

The example of the digital voltmeter shows how digital carry may be used to improve accuracy. On the other hand, it can also be applied in a device of given accuracy to provide faster operation.

10. Acknowledgment

The authors thank the Directors of The Solartron Electronic Group for giving permission to present this paper.

Manuscript first received by the Institution on 5th May 1969 and in final form on 13th March 1970. (Paper No. 1330/IC26.)

J. R. Pearce took a London engineering degree with honours after completing Army service as a radar instructor. For seven years he was engaged on instrumentation for engine research with Bristol-Siddeley, and he then joined Solartron where he has been for the past thirteen years. He has worked on the development of power supplies, X-ray spectrometers and digital voltmeters, and for four years he was production manager for digital voltmeters. Mr. Pearce is now responsible for engineering standards in the R & D Division.

J. Bloomfield joined Solartron in 1955 as a Technical Assistant and subsequently obtained an H.N.D. at Kingston Technical College. He was responsible for some of Solartron's early oscilloscopes, but soon transferred to digital work. He built one of the first digital voltmeters in the world, and as team leader, developed many of his company's designs.

U. Qureshi was educated at King's College, London, and Twickenham Technical College where he gained an H.N.C. and joined Solartron in 1959. After a period in production and quality control he moved to R & D where he has been concerned with the development of digital voltmeters for seven years.

U.D.C. 621.376.3(091)

The Invention of Frequency Modulation in 1902

By

Professor D. G. TUCKER,
D.Sc., C.Eng., F.I.E.E., F.I.E.R.E.†

It is shown that the accounts in the literature of the early history of frequency modulation are very confused and contradictory. After demonstrating that Fessenden in his 1901 patent application almost certainly did not intend to have or to utilize f.m., but was concerned only with a.m., it is shown that the real first inventor of f.m. was C. D. Ehret with his very clearly stated patent application of 1902.

1. Introduction

Frequency modulation (f.m.) is now a very important process in radio communication, and is widely used in preference to amplitude modulation (a.m.) at the higher radio frequencies (e.g. v.h.f.) where the availability of adequate bandwidth enables better signal/noise discrimination to be obtained by the use of f.m. Its introduction as an important process in the communication of speech and music is generally, and correctly, credited to Armstrong,[1] who published his comprehensive paper in 1936 following extensive pioneering experiments from 1933 onwards. In this paper he described technical means for achieving f.m. communication, he reviewed the state of knowledge on noise interference in radio and on thermal agitation (or 'Johnson') noise[2] as a fundamental limitation on communication, and he described for the first time the special noise-discriminating properties of f.m.

It is also well known that Carson[3] in 1922 had analysed for the first time the sideband structure of f.m., and had shown that, unlike a.m., f.m. had an infinite set of sidebands. Above all he showed that:

'...the transmission of the signal by frequency modulation requires the transmission of a band of frequencies *at least* $2p/2\pi$ in width; *that is* a band of width equal to twice that of the signal itself.'

Here p is the signal frequency in rad/s.

He alleged, and subsequent writers, e.g. Heising,[4] have repeated, that earlier workers had believed that f.m. would permit speech to be transmitted in a *narrower* bandwidth than that required by a.m. His work certainly should have disposed of this fallacy, but Heising states that there was subsequent work done by others in apparent ignorance of Carson's conclusions.

When, however, we try to discover the earlier history of f.m., that is, before Carson, we find a very confused picture in the literature. Contradictions abound, as will be demonstrated below. The purpose of the present paper is to clarify the origins of f.m. It will be shown that f.m. was invented in 1902 by C. D. Ehret.

2. The Confusion

Various writers who have discussed the history of f.m. give various accounts of its early development, and attribute the basic invention differently. To illustrate this, it will be best to let the various writers speak for themselves. Heising, who was himself an early worker in the field of communications,[5] has recently written:[6]

'The earliest discussion on f.m. in telephony is a short item in a book by Dr. A. N. Goldsmith in 1918.'

Yet in the text-book by Ruhmer[7] (probably the first ever published on wireless telephony, appearing in German in 1907 and in English translation by Erskine-Murray in 1908) we read:

'The conversion of the sound waves may of course take place in any of the ways already described, and may depend on the variation of either the intensity or frequency of the oscillations. In both groups of methods the transmission of speech is by means of varying electrical waves, in the first case with constant, and in the second with varying frequency, which act with varying intensity on the detector in the receiving circuit and thus reproduce the corresponding sounds in the telephone attached to it.

'When a constant wave-length is used the variations of the intensity produce similar variations in the receiver, while if transmission depends on variation of the wavelength, the receiver must be a persistent oscillator which only responds when acted upon by waves which are of its own natural frequency, and falls out of resonance when the frequency of the impressed waves varies.'

This looks very much like a description of f.m., although the second paragraph has a suggestion that Ruhmer may have had his thoughts rather more on frequency-shift keying in telegraphy than on telephony. In this connexion it is worth stating that the keying of telegraph signals by frequency-shift was an established practice in those days,[8] and may be regarded

† Department of Electronic and Electrical Engineering, University of Birmingham.

as a very crude kind of f.m.

In his classic paper, already referred to, Armstrong[1] says:

'The subject of frequency modulation is a very old one. While there are some vague suggestions of an earlier date, it appears to have had its origin shortly after the invention of the Poulsen arc, when the inability to key the arc in accordance with the practice of the spark transmitter forced a new method of modulation into existence.'

This clearly is equating f.m. with frequency-shift telegraphy and erroneously placing the date of origin after 1903, which was the year in which the Poulsen arc was introduced.[9]

Some uncertainty in the exact meaning of the wording in the patent specification filed in 1901 by the famous early wireless engineer Fessenden,[10] who is usually and probably incorrectly credited with the invention of a.m., has led to some suggestion that he also had f.m. in mind. Heising[11] says:

'Fessenden, in 1902, proposed using a condenser type of microphone in circuits that appear to produce f.m., but in his published remarks about tests he seemed interested in using the frequency variation to produce amplitude variation by throwing the frequency in and out of resonance with the antenna or other tuned circuit to modulate the amplitude.'

The same idea that Fessenden was proposing frequency modulation, although merely to obtain amplitude modulation, appears in Ruhmer's book[12] of 1907, already referred to. He says:

'In another of Fessenden's senders the control of the radiation is effected by alteration of the frequency through variation of the inductance or capacity of the aerial system.'

The present author has long held the view[13] that Fessenden did not actually intend the frequency of the radiated signal to vary significantly as the speech amplitude varied, and that the inductance (or capacitance) which varied in value as the speech amplitude varied was intended as what we would now call a magnetic (or a dielectric) amplitude-modulator. This was also the view of Pungs[14] in 1923. Referring to Fessenden's patent[10] he says:

'Für die Steuerung von Hochfrequenzströmen in das Verfahren zuerst von Fessenden (1902) vorgeschlagen worden, jedoch in einer praktisch nicht ausführbaren Form (Ringspule mit kurzgeschlossener Steuerwicklung).'†

† 'For the control of high-frequency currents the arrangement proposed by Fessenden in 1902 was the first, although in a form not workable in practice (ring-coil with short-circuited control winding).'

As it is rather important to settle this matter, the relevant drawings from Fessenden's patent and the relevant part of his description are reproduced in the Appendix.

The important words here are those put into italics: 'sustained oscillations of practically constant frequency'. It seems clear that Fessenden intended to produce a.m. (not f.m.) by causing the speech amplitude to control the division of current between the low-loss tuned circuit (of constant parameters) and the more lossy tuned aerial (of variable parameters). Of course, there would have been some frequency variation, but this was merely a stray undesired effect. It was not an efficient arrangement, quite apart from the fault mentioned in Pungs' critical comment, quoted above.

If this interpretation, that Fessenden had no idea of using f.m., is accepted, then the field is left clear for the true inventor, Cornelius D. Ehret of the U.S.A., who filed his patents in 1902. The clue to Ehret's status in this regard was given by Guy,[15] who, however, gave no reference to the documents concerned.‡

3. Ehret's Patents on F. M., 1902

Ehret's patent on f.m. was divided into two almost identical specifications,[16,17] one dealing with the method and the other with the systems. There can be no doubt whatever that he describes f.m. as now understood:

'It comprises, further, a method of modifying and varying the frequency of the electroradiant energy in a manner corresponding and in accordance with the signal to be transmitted.

'It resides also in an additional step of modifying the energy to be transmitted and received by and in accordance with sound-waves, such as speech.

'It comprises, further, a method of receiving the modified transmitted energy and causing the reproduction of speech and other signals by the effects of variations or changes in the frequency of the received energy.'

His claim is a very concise definition of f.m.:

'The method of transmitting intelligence, which

‡ The author is grateful to one of the referees for pointing out that Ehret's patents on f.m. are also cited, albeit very casually and without discussion, by Hammond and Purington,[30] and by Miessner[31] who incorrectly states that Ehret was concerned only with frequency-shift telegraphy. Unfortunately these additional references, while interesting in their own right as personal statements by authors who themselves contributed greatly to the development of radio, nevertheless tend to increase the confusion already commented on by the present author.

consists in generating electroradiant energy, modifying the frequency of said energy in accordance with the signal to be sent and receiving the energy in a device responsive to changes in the frequency of the transmitted energy.'

The process of modulation was to modify the frequency of the second stage of a two-stage spark transmitter, e.g. by connecting the microphone across a tuning inductance. The receiver used the variation of voltage across an inductance connected in series with the aerial or across a shunt-tuned circuit coupled to it. It is perhaps unlikely that the system could have worked well with Ehret's proposed circuits, but the idea of the system is perfectly clear and correct. The patent also very clearly covers frequency-shift telegraphy. It is, moreover, so worded as to suggest that Ehret thought he was the first to discover any way of achieving the transmission of speech by wireless, although we have seen above that Fessenden had filed the patents for his method the year before. Ehret's words are:

'...it is apparent that I have disclosed a method of transmitting speech electrically without the employment of conductors joining the transmitting and receiving stations; that speech is transmitted by this method by the agency of electroradiant energy, as employed heretofore for telegraphy only; ...'

4. Some Additional Notes on Ehret's Work

As far as can be ascertained from the U.S. Patent files, Cornelius D. Ehret was awarded 35 U.S. patents, from 1902 to 1933, with eight preceding his f.m. patents. Of the 35, four were for inventions not connected with electricity or communications: centrifugal machine, 1908; making wire-glass, 1914; optical pyrometry, 1920; and method and apparatus for compressing fluids, 1928. None of the patents preceding the f.m. disclosure were particularly related to it, except that one[21] which was filed around the same time shows a system in which the transmitted frequency is changed in synchronism with the tuning of the receiver, at quite high speed, to make a secrecy system. In the same year as the f.m. patents were issued, Ehret also had patents[22,23] granted covering telephone line repeaters and frequency-division multiplex.

It is interesting that at intervals other inventors with the name Ehret occur in the U.S. Patent Indexes, and as some of these give Philadelphia as their address (as does C. D. Ehret in many of his patents), one cannot help wondering if there was a whole family of inventors. C. D. Ehret himself took the assignment of several non-electrical patents, so he must have varied business interests. Guy[15] describes him as a Philadelphia patent attorney.

It is interesting that Ehret has almost completely escaped the attention of those writing on the history of radio. Apart from books and papers already cited, there are several others which give a substantial historical review of the early days of radio and communications generally (e.g., Refs. 24–28), and as far as the present author can detect, no reference to Ehret or his ideas is given in any of them.

It is probable that Ehret did not publish his ideas in any form other than patent specifications, as the only contribution he made to the American Institute of Electrical Engineers during the years 1902–1910 inclusive was a group of three short questions[29] on the audion at a meeting in Philadelphia. The author has found no evidence that Ehret carried out any experimental work on his ideas. It was to be thirty years before practical significance was to be attached to f.m., and it is perhaps understandable that during this time Ehret's patent should be forgotten. Even at the time of issue it may well have had no impact, as it was clearly premature; the technology of 1902–5 did not give it a real chance of application then.

5. The Word 'Modulation'

Guy[15] seems to suggest that Ehret actually used the word 'modulation' in his patents. Guy's words are:

'So far as is known these were the first disclosures to describe any system of modulation by name.'

But the word 'modulation' does *not* appear in Ehret's patents, except once (in the plural) in the traditional meaning of the fluctuations of the voice itself: never in connexion with the electrical system.

Heising[18] attributes the introduction of the word 'modulation' to Fessenden in an article of 1907.[19] The relevant sentence by Fessenden is:

'The difficult problem in wireless telephony is, of course, the modulation of the large amount of energy used for transmission.'

No explanation of the word is given, and one cannot help feeling that the author was under the impression that the word was generally understood in this sort of context.

6. References

1. Armstrong, E. H., 'A method of reducing disturbances in radio signalling by a system of frequency modulation', *Proc. I.R.E.*, **24**, pp. 689–740, May 1936.
2. Johnson, J. B., 'Thermal agitation of electricity in conductors', *Phys. Rev.* **32**, p. 97, July 1928.
3. Carson, J. R., 'Notes on the theory of modulation', *Proc. I.R.E.*, **10**, pp. 57–64, February 1922.
4. Heising, R. A., 'Modulation methods', *Proc. I.R.E.*, **50**, pp. 896–901, May 1962 (see p. 899).
5. For short biography see *Proc. I.R.E.*, **50**, p. 1430, May 1962.
6. Heising, R. A., *loc. cit.*, p. 897.

7. Ruhmer, Ernst, 'Wireless Telephony in Theory and Practice', translated from the German by J. Erskine-Murray. (Crosby Lockwood & Son, London, 1908). (Published in German 1907). See pp. 136–7.

8. Weldon, J. O., 'Transmitters', *Proc. I.R.E.*, **50**, pp. 901–906, May 1962 (see p. 903).

9. Poulsen, V., 'Method of producing alternating currents with a high number of vibrations', U.S. Patent No. 789,449, issued 9th May 1905, filed 19th June 1903.

10. Fessenden, R. A., 'Wireless signaling', U.S. Patent No. 753,863, filed 28th September 1901, patented 8th March 1904.

11. Heising, R. A., *loc. cit.*, pp. 896–7.

12. Ruhmer, E., *loc. cit.*, p. 140.

13. Tucker, D. G., 'Modulators and Frequency Changers', p. 93. (Macdonald, London, 1953).

14. Pungs, L., 'Die Steuerung von Hochfrequenzströmen durch Eisendrosseln mit überlagerter Magnetisierung', *Elektrotechnische Zeitschrift*, **4**, pp. 78–81, 1923 (see p. 78).

15. Guy, R. F., 'A.M. and F.M. Broadcasting', *Proc. I.R.E.*, **50**, pp. 811–817, May 1962 (see p. 816).

16. Ehret, C. D., 'Art of transmitting intelligence', U.S. Patent No. 785,803, filed 10th February 1902, patented 28th March 1905.

17. Ehret, C. D., 'System of transmitting intelligence', U.S. Patent No. 785,804, filed 10th February 1902, patented 28th March 1905.

18. Heising, R. A., *loc. cit.*, p. 896.

19. Fessenden, R. A., 'Wireless Telephony', *Electrical Review*, **60**, pp. 251–253, 15th February 1907. (N.B. This was Part 1 of a 3-part article; the other parts were pp. 327–329 and 368–370 of the volume).

20. Palmer, L. S., 'Wireless Principles and Practice', p. 400. (Longmans, Green, London, 1928).

21. Ehret, C. D., 'Wireless selective signalling system', U.S. Patent No. 717,978, issued 6th January 1903, filed 3rd July 1902.

22. Ehret, C. D., 'Art of electrical wave transmission', U.S. Patent No. 789,087, issued 2nd May 1905, filed 3rd December 1901.

23. Ehret, C. D., 'System of electrical wave transmission', U.S. Patent No. 789,124, issued 2nd May 1905, filed 3rd December 1901.

24. Armstrong, E. H., 'Evolution of frequency modulation', *Electrical Engineering*, **59**, p. 485, December 1940.

25. Eccles, W. H., 'Wireless'. (Thornton Butterworth, London, 1933).

26. Blake, G. G., 'History of Radio Telegraphy & Telephony', (Radio Press, London, 1926).

27. Colpitts, E. H. and Blackwell, O. B., 'Carrier current telephony and telegraphy', *Trans. Amer. Inst. Elect. Engrs.* **40**, pp. 205–300 (1921).

28. Dunlap, O. E., 'Radio's 100 Men of Science'. (Harper, New York and London, 1944).

29. Ehret, C. D., *Trans. Amer. Inst. Elect. Engrs.* **25**, pp. 775, 776, 778 (1906).

30. Hammond, J. H. and Purington, E. S., 'A history of some foundations of modern radio-electronic technology', *Proc. I.R.E.*, **45**, pp. 1191–1208, 1957 (see p. 1205).

31. Miessner, B. F., 'On the Early History of Radio Guidance'. (San Francisco Press, 1964) (see p. 60).

7. Appendix

Extract from Fessenden's Patent Specification

'In the practice of my invention I provide at the sending-station a conductor 1 of suitable construction and arrangement and connect the same to one terminal of a coil 2, surrounding a core 3, preferably annular in shape and preferably formed of fine iron wire. The other terminal of the coil is connected to one of the knobs or terminals 4 of an induction-coil or other suitable generator 6 capable of producing practically continuous and rapid oscillations in the conductor. The systems described in connection with Fig. 10 of Patent No. 706,742 or in Patent No. 730,753 are well adapted for this purpose. The other terminal, 5, of the generator is connected to ground. A second coil 7, forming a part of the circuit for the battery 8, is placed on the core 3, and a transmitter 9, preferably microphonic in construction, or other mechanism capable of modifying the current in the circuit is included in the circuit of the battery and coil 7. A capacity 18 and inductance 19 are connected in shunt to the spark-gap for the purpose of maintaining *sustained oscillations of practically constant frequency*. [Italics due to present author.] The capacity 18 and inductance 19 should be arranged to have the same period of oscillation as the receiving-conductor 10 and the sending-conductor 1. It will be seen that the circuit containing capacity 18 and inductance 19 being connected across the spark-gap, forms a parallel circuit in the sending conductor 1, whose aerial and grounded sections are also connected across the spark-gap. On account of the fact that the circuit 18, 19, and the sending-conductor 1 are in parallel and not in series the difference of potential across these two circuits is the same, while the currents in the two circuits are different, this construction being thus differentiated from a series connection, in which the circuit 18, 19, would be connected electrically between the aerial portion of the sending-conductor 1 and the ground.

'At the receiving-station I employ a conductor 10, connected to one terminal of the mechanism capable of responding to oscillations in the conductor 10. A form of mechanism adapted to the purpose consists of a coil 11, having one terminal connected to the conductor and the other terminal grounded. A telephone-diaphragm 12, formed of metal or consisting of insulating material having a metal plate or coil of wire secured thereto, or any other suitable construction adapted to vibrate in unison with changes of current or voltages produced by the waves radiated from the sending-station is suitably supported in operative relation to the coil 11. The apparatus at the receiving-station is tuned or made resonant by any suitable means known in the art to the sending-conductor 1. The terms "tuned" and "resonant" are used herein, one to include the other. When an alternating current is set up in the conductor 10, as by waves or impulses from the sending-station, such current acts to repel or attract the diaphragm, according to the time constant of the metal part of the diaphragm, through induced currents set up in the diaphragm. When the generator is operated, the diaphragm 12 will take up a mean position relative to coil 11, the distance of such position from the coil varying with the intensity of the oscillations in the sending-conductor; but

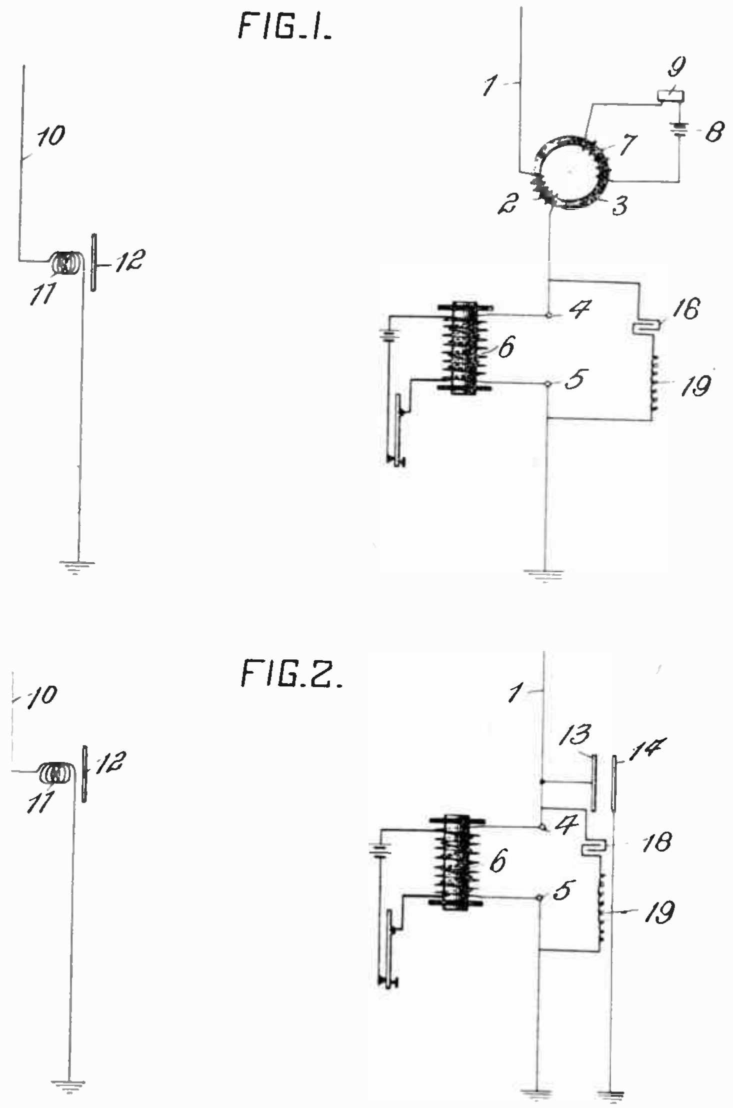


Fig. A. From R. A. Fessenden's U.S. Patent of 1904.[10]

when the current in the circuit of the coil 7 is modified or changed by speaking into the transmitter the permeability of the core 3 is correspondingly changed or modified, thereby producing a corresponding change or modification in the self-inductance and a change in the frequency of the natural period of vibration of the sending-conductor 1, which is thereby thrown out of resonance with a resonating-circuit 18, 19, connected in parallel to said sending-conductor 1, and due to this failure in resonance producing a corresponding change or modification in the intensity of the waves or impulses given off by the conductor 1 and in the intensity of the oscillations produced in the receiving-conductor. The changes in the intensity of the oscillations will produce corresponding changes in the mean position of the diaphragm, such changes corresponding to the vibrations of the diaphragm of the transmitter, exactly reproducing any of the waves or impulses which affected the transmitter. The same result may be effected by changing the capacity of the conductor 1, as shown in Fig. 2. To this end the conductor 1 is connected to a fixed plate 13 of a condenser, while the other plate, 14, is formed by or connected to a diaphragm capable of responding to waves or impulses. As the plate 14 in vibrating moves toward or from the other plate the capacity of the conductor 1 is changed, correspondingly altering the intensity of the waves or impulses generated by the conductor.'

Manuscript received by the Institution on 6th February 1970. (Paper No.1331/Com. 31.)

U.D.C.631.372.5

Synthesis using Symmetric Distributed RC-Structures

By

Professor M. N. S. SWAMY,
B.Sc.(Hons.), D.I.I.Sc., M.Sc., Ph.D.,†

J. C. GIGUERE,
B.E., M.E., Ph.D.‡

and

B. B. BHATTACHARYYA,
B.Tech.(Hons.), M.Tech., Ph.D.†

This paper deals with the synthesis of transfer functions using symmetric structures, a symmetric structure being a three-terminal network composed of distributed and/or lumped RC elements, and having $z_{11} = z_{22}$. Necessary and sufficient conditions are presented for the realization of a class of transfer functions by a three-terminal structure, which is a cascade of a number of networks of two types, one type being a ladder network whose different elements are cascades of symmetric structures, and the other type being a cascade of n symmetric structures. Realization procedures for this class of transfer functions are given. An extension is considered for the realization of a transfer function as a parallel interconnexion of such structures. Existing transfer function synthesis procedures using uniform lines are extended to include symmetric structures.

1. Introduction

In the realization of network functions using RC-lines, one approach which has received some attention in the literature is to transform the distributed network problem in the s-plane into an equivalent lumped one in some transformed plane.[1–8] The present paper adopts the same approach and considers the realization of transfer functions using symmetric structures (s.s.).

The authors have, in a previous paper, used this approach to examine the synthesis of driving point functions[7], and have established that the necessary and sufficient conditions that a driving point function be realized as an interconnexion of one-port impedances (o.p.i.) proportional to $p(s)$ and $q(s)$ are that it be expressible in the form

$$Z(s) = p(s)\,[\text{RC-impedance function in } u = p/q], \qquad \ldots\ldots(1)$$

where $p(s)$ and $q(s)$ correspond either to any two o.p.i.'s of a single RC-line, or to those of any two lines. It was also noted in that paper that the transfer function of a two-port consisting of an interconnexion of one-ports with impedances proportional to $p(s)$ and $q(s)$ must necessarily be of the form

$$T[u(s)] = [\text{'RC-transfer function' in } u = p/q]. \qquad \ldots(2)$$

That is, $T(u)$ has simple poles on the negative real axis of the u-plane, but having arbitrary zeros of transmission and satisfying the Fialkow–Gerst conditions. Hence, if a transfer function of the form (2) is given, it can always be realized first in the appropriate lumped plane by any known technique, after which the different resistors R_i and capacitors C_i are replaced by RC-lines having one-port impedances given by $R_i p$ and $(1/C_i)q$ respectively.

Since this paper deals with synthesis using s.s.'s, we shall recall the definition of a s.s. as given elsewhere.[7] A s.s. is a three-terminal network consisting of distributed and/or lumped RC elements with a $[z]$ of the form

$$[z]_i = \frac{r_i}{F_2}\begin{bmatrix} F_1 & 1 \\ 1 & F_1 \end{bmatrix} \qquad \ldots\ldots(3)$$

where F_1 and F_2 are functions of s, and r_i is an impedance scaling factor. There may be a number of ways of obtaining a s.s. using tapered lines. One way, attractive from the point of view of fabrication, is by a cascade of the given line in a 'back to back' or 'front to front' connexion.[7] Such s.s.'s are called 'symmetric lines' (s.l.).

The synthesis of transfer functions to be discussed later depends heavily on the realization of driving point functions (d.p.f.) as cascade of commensurate s.s.'s, and hence we shall briefly discuss the synthesis of a d.p.f. The following theorem gives the conditions for realizing a d.p.f.

Theorem 1:

(a) The necessary and sufficient conditions that an impedance $Z(s)$ be cascade-synthesizable using commensurate symmetric structures with a $[z]_i$ given by (3) is that it be expressible in the form

$$Z(s) = K\frac{F_1}{F_2}\frac{(F_1^2 - z_1^2)\ldots(F_1^2 - z_n^2)}{(F_1^2 - p_1^2)\ldots(F_1^2 - p_n^2)}$$

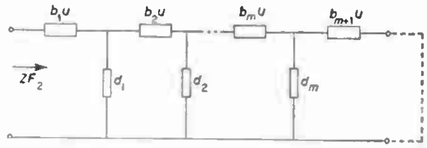
where

$$0 \le p_n^2 < z_n^2 \ldots < p_1^2 < z_1^2 \le 1, \text{ and } K > 0.$$

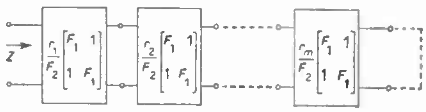
(b) The number of symmetric structures required to cascade-synthesize $Z(s)$ is one less than or equal to the degree of the numerator polynomial of ZF_2, the last

† Department of Electrical Engineering, The University of Calgary, Canada; now at Sir George Williams University, Montreal, Canada.

‡ Department of Electrical Engineering, Sir George Williams University, Montreal, Canada.



(a)



(b)

Fig. 1. Realization of a driving point function $Z(s)$ satisfying the conditions of Theorem 1.

structure being short- or open-circuited depending on whether (F_1^2-1) is or is not a factor of ZF_2.

The proof of this theorem may be found elsewhere.[7] We shall just give here a method of cascade-synthesizing a given $Z(s)$ satisfying Theorem 1. For this purpose, we first let $u = (F_1-1)$. Then,

$$ZF_2 = \frac{\text{polynomial in } u \text{ of degree } (m+1)}{\text{polynomial in } u \text{ of degree } m},$$

where m may be odd or even. We now expand ZF_2 as a continued fraction to obtain a Cauer Ladder, as shown in Fig. 1(a). As $Z(s)$ satisfies the necessary and sufficient conditions of Theorem 1, the different b and d automatically satisfy the conditions,[7]

$$b_1 = d_1 = r_1; \quad b_{m+1} = d_m = r_m; \quad b_i = d_{i-1}+d_i, \quad (i = 2, 3, \ldots m).$$

We now extract the different T-sections as corresponding to those of symmetric structures. The last structure is recognized as being open- or short-circuited, depending on whether the last term of the continued fraction expansion contains u or not. The synthesis is now complete and is shown in Fig. 1(b).

We shall now consider the realization of a class of transfer functions using s.s.'s.

2. Class of Transfer Functions Realizable by a Three-terminal Network Composed of Symmetric Structures

The class of networks N to be studied is shown in Fig. 2(a). Each of the three-terminal networks, $N_1, N_2, \ldots, N_r$ may be either a ladder network, N_L, as shown in Fig. 2(b), or a network N_C which is a cascade of n s.s.'s, as shown in Fig. 2(c). Each element of a given N_L may itself be an interconnexion of the short- and open-circuit impedances $K_ip(s)$ and $K_jq(s)$ of a s.s., K_i and K_j being impedance scaling factors. It should be remembered that each element may be realized as a cascade of s.s.'s.[7] In order to establish the main theorem of this Section, we need the following lemma.

Lemma:

The chain matrix of a cascade of n symmetric structures is of the form

$$[a]_n = \frac{1}{(1-u)^{n/2}} \begin{bmatrix} K_\alpha \prod_1^k (u+\alpha_i) & K_\beta p \prod_1^l (u+\beta_i) \\ K_\gamma \frac{u}{p} \prod_1^l (u+\gamma_i) & K_\delta \prod_1^k (u+\delta_i) \end{bmatrix} = \begin{bmatrix} A_n & B_n \\ C_n & D_n \end{bmatrix} \qquad \ldots\ldots(4)$$

where $u = p/q$, p and q being proportional to the short- and open-circuit impedances of the s.s., and where the following conditions hold:

(a) If n is even, $k = n/2$ and $l = (n/2)-1$, while if n is odd, $k = l = (n-1)/2$,

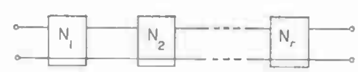
(b) α_i and γ_i, δ_i and γ_i, δ_i and β_i, α_i and β_i all interlace on the negative real axis of the u-plane. Further, $0 < \alpha_i < \gamma_i$, $0 < \delta_i < \gamma_i$, $0 < \delta_i < \beta_i$, and $0 < \alpha_i < \beta_i$.

(c) In addition, if the s.s. is a symmetric line, then

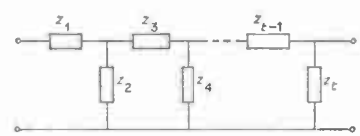
$$K_\alpha = \left[1/\prod_1^k \alpha_i\right], \qquad K_\beta = \left[\sum_1^n r_i\right] \Big/ \left[\prod_1^l \beta_i\right],$$

$$K_\gamma = \left[\sum_1^n (1/r_i)\right] \Big/ \left[\prod_1^l \gamma_i\right], \quad K_\delta = \left[1/\prod_1^k \delta_i\right],$$

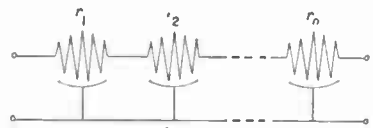
r_i being the impedance scaling factor for the $[z]$ of ith symmetric line.



(a) The structure N composed of the networks N_L and N_C.



(b) The ladder network N_L where each of the impedances $z_1, z_2, \ldots, z_r$ are cascades of symmetric structures.



(c) The network N_C where the symmetric structure is drawn as a symmetric line.

Fig. 2. The three-terminal structure realizing the class of transfer functions given by Theorem 2.

Proof: Using (3), the chain matrix of the ith symmetric structure may be written as

$$[a] = F_1 \begin{bmatrix} 1 & r_i(F_1^2-1)/F_1F_2 \\ (1/r_i)(F_2/F_1) & 1 \end{bmatrix}. \quad \ldots\ldots(5)$$

Letting

$$p(s) = (F_1^2-1)/F_1F_2 \quad \text{and} \quad q(s) = (F_1/F_2) \quad \ldots\ldots(6)$$

(5) may be rewritten as

$$[a] = \frac{1}{(1-u)^{1/2}} \begin{bmatrix} 1 & r_i p \\ \dfrac{u}{r_i p} & 1 \end{bmatrix} \quad \ldots\ldots(7a)$$

where

$$u = p/q = (F_1^2-1)/F_1^2, \quad \ldots\ldots(7b)$$

or in the form

$$[a] = \frac{1}{(1-u^2)^{1/2}} \begin{bmatrix} 1 & r_i\sqrt{pq}\,u \\ \dfrac{u}{r_i\sqrt{pq}} & 1 \end{bmatrix} \quad \ldots\ldots(8a)$$

where now

$$u = \sqrt{p/q} = \frac{\sqrt{F_1^2-1}}{F_1}. \quad \ldots\ldots(8b)$$

We may now establish part (a) of the lemma by induction, while part (b) follows from the fact that (A_n/uC_n), (D_n/uC_n), (B_n/A_n) and (B_n/D_n) are all RC impedance functions in u.

From a consideration of the low-frequency asymptote of the A-parameter of an RC-line[9], we conclude that $F_1 = (1-u)^{-1/2}$ tends to unity as s tends to zero, where u is given by (7b). Hence at low frequencies, for a symmetric line, $u(s)$ tends to zero as s tends to zero. Now taking the low-frequency limit of $[a]_n$ and using the results of Protonatorios and Wing[9], we conclude that in the case of symmetric lines

$$K_\alpha \prod_1^k \alpha_i = 1$$

$$K_\beta p(0) \prod_1^l \beta_i = R_T = p(0) \sum_1^n r_i \quad \ldots\ldots(9)$$

$$K_\gamma [1/q(0)] \prod_1^l \gamma_i = C_T = [1/q(0)] \sum_1^n (1/r_i)$$

and

$$K_\delta \prod_1^k \delta_i = 1,$$

where R_T and C_T are the total resistance and capacitance of the cascade of the n symmetric lines. From (9), part (c) of the lemma follows.

With this result, we now present the main theorem of this section.

Theorem 2:

A necessary and sufficient condition that a transfer function $T(s)$ be realizable by the three-terminal structure N, (N is a cascade of a number of networks of two different types, one type being a ladder network, N_L, whose different elements are cascades of symmetric structures, and the other type, N_C, being a cascade of n symmetric structures), is that $T(s)$ be expressible as

$$T[u(s)] = (1-u)^{n/2} \cdot \frac{P(u)}{Q(u)} \quad \ldots\ldots(10)$$

where $u = p(s)/q(s)$, $p(s)$ and $q(s)$ being respectively proportional to the short- and open-circuit impedances of the symmetric structure, and where the following conditions are met:

(a) $P(u)$ and $Q(u)$ are polynomials in u,
(b) zeros of $P(u)$ are on the negative real axis of the u-plane,
(c) zeros of $Q(u)$ are simple and are on the negative real axis of the u-plane, and
(d) (the degree of Q) $\geq$ (the degree of P)+(n/2), if n is even,
(the degree of Q) $\geq$ (the degree of P)+($n-1$)/2, if n is odd.

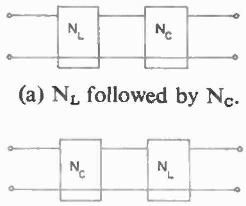
Proof:

Necessity: We may assume that the chain matrix of a given N_L (Fig. 2(b)) is of the form

$$[a]_{N_L} = \frac{1}{E} \begin{bmatrix} A & pB \\ (1/p)C & D \end{bmatrix} \quad \ldots\ldots(11)$$

where A, B, C, D and E are polynomials in u with zeros on the negative real axis of the u-plane such that (A/C), (B/A), (B/D) and (D/C) are RC impedance functions of u, and $0 \leq$ (the degree of E) $\leq$ (the degrees of A, B, C, D). The chain matrix of a given N_C (Fig. 2(c)) has already been obtained in the form (7) in the lemma. It follows that the open-circuit voltage transfer function of the structure of Fig. 2(a) will be of the form (11), where the conditions (a) and (b) of the theorem are obviously satisfied. Condition (c) follows from the fact $Q(u)$ must correspond to the numerator of an RC driving point function.

Now let (the degree of Q) = m_q, and (the degree of P) = m_p. To prove (d), we short all the series impedances and open all the shunt impedances of the sub-networks N_L. If n is even, we have that $m_p = 0$ and $m_q = (n/2)$. We now return to the network those series or shunt elements which consist of single short-circuited symmetric structures, and those shunt elements which consist of single open-circuited symmetric structures. In the u-plane, these elements correspond to resistors and capacitors respectively. Hence, $m_p = 0$, but now $m_q \geq (n/2)$, since the addition of the elements did not create any zeros of transmission, but did increase the degree of the driving point function associated with Q. Finally, we replace the remaining series and shunt impedances. Each



(a) N_L followed by N_C.

(b) N_C followed by N_L.

Fig. 3. The structure used to prove the sufficiency part of Theorem 2.

element contributing a zero of transmission will increase both m_p and m_q by one. A similar argument holds when n is odd. Hence we have the result (d), and the necessity part of the theorem.

Sufficiency: We shall prove the sufficiency part by showing that (11) can always be realized by the structure of Fig. 3(a). For this purpose, we divide both the numerator and denominator of (11) by an arbitrary polynomial $d(u)$ such that (Q/d) is an RC impedance function in u. We may now identify z_{11} and z_{21} as

$$\frac{z_{11}}{p} = \frac{Q}{d} \quad \text{and} \quad \frac{z_{21}}{p} = (1-u)^{n/2} \cdot \frac{P}{d}.$$

The function $p(s)$ has been introduced so that z_{11} is of the form (1), and as such may be realized using the impedances $p(s)$ and $q(s)$.

We will demonstrate the sufficiency when n is even. (A similar argument will apply when n is odd.) Let $n = 2r$, and the degrees of P, Q, and d be m_p, m_q and m_q respectively. Then, from the given necessary conditions,

$$m_q \geq (m_p + r), \quad \text{or} \quad m_q = (m_p + r + t).$$

The finite transmission zeros which are m_p in number may be realized in the u-plane by synthesizing (z_{11}/p) as an RC ladder network. Each one-port may then be replaced by a cascade of symmetric structures.[7] Once these zeros are realized, the remaining driving point function, say (Z_1/p), will have $(r+t)$ as the degrees of its numerator and denominator. Next the t transmission zeros at $u = \infty$ are removed by a ladder synthesis of (Z_1/p). The resistors and capacitors correspond to the short- and open-circuited symmetric structures respectively. The remaining impedance, say (Z_2/p), has r poles and zeros, one of the poles being at the origin. We see from the lemma that the impedance Z_2 can be realized as an open-circuited cascade of $2r$ symmetric structures, and that the required transmission zeros at $u = 1$ are automatically realized. (When the symmetric structure is a symmetric line, this corresponds to realizing transmission zeros at $s = \infty$.) Thus the sufficiency part of the theorem is proved.

Sufficiency could also have been demonstrated by constructing suitable y_{21} and y_{22}, and realizing the network as the structure of Fig. 3(b). Hence, the theorem is proved.

We shall now illustrate our synthesis procedure by a suitable example.

Example:

Let the transfer function to be realized be

$$T[u(s)] = k_m \frac{(1-u)^{3/2} u(u+1)}{(u+0{\cdot}04)(u+0{\cdot}5)(u+1{\cdot}3)(u+2{\cdot}5)}$$
$$= k_m (1-u)^{3/2} \cdot \frac{P}{Q}. \qquad \ldots\ldots(12)$$

It is obvious that the realization of N_C will contain three symmetric structures in cascade. Choosing the arbitrary polynomial to be

$$d(u) = u(u+0{\cdot}06)(u+1)(u+1{\cdot}7)(u+3{\cdot}5),$$

we have

$$\frac{z_{21}}{p} = k_m \frac{(1-u)^{3/2} u(u+1)}{u(u+0{\cdot}06)(u+1)(u+1{\cdot}7)(u+3{\cdot}5)}$$

and

$$\frac{z_{11}}{p} = \frac{(u+0{\cdot}04)(u+0{\cdot}5)(u+1{\cdot}3)(u+2.5)}{u(u+0{\cdot}06)(u+1)(u+1{\cdot}7)(u+3{\cdot}5)}.$$

The zeros of transmission at $u = 0$ and $u = -1$ are realized by the usual ladder synthesis of (z_{11}/p). This is shown in Fig. 4(a). The remaining impedance is

$$\frac{Z_1}{p} = \frac{0{\cdot}6866u^2 + 1{\cdot}823u + 5.790}{0{\cdot}5766u^3 + 4{\cdot}954u^2 + 7{\cdot}309u + 6{\cdot}896}.$$

Next, the zero of transmission at $u = \infty$ is realized in such a way that the remaining driving point function may be realized as a cascade of three symmetric structures. For this purpose a ladder network is removed as shown in Fig. 4(b). The remaining impedance (Z_2/p) is

$$\frac{Z_2}{p} = \frac{0{\cdot}5969u + 0{\cdot}5790}{u(2{\cdot}605u + 4{\cdot}652)}.$$

Since

$$p = (F_1^2 - 1)/F_1F_2, \quad q = F_1/F_2 \quad \text{and} \quad u = (F_1^2 - 1)/F_1^2,$$

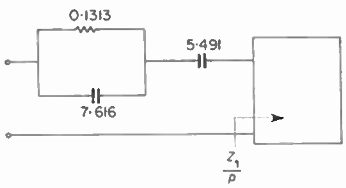
the impedance Z_2 may be written as

$$Z_2 = 0{\cdot}1620 \frac{F_1}{F_2} \frac{(F_1^2 - 0{\cdot}5076)}{(F_1^2 - 0{\cdot}3590)}$$

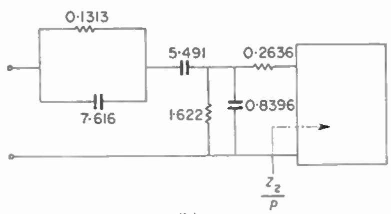
and can be realized as a cascade of three symmetric structures.[7]

It should be emphasized that until now we have made no assumption about the nature of the symmetric structure to be used. If we assume that uniform lines of unity RC product are to be used, then

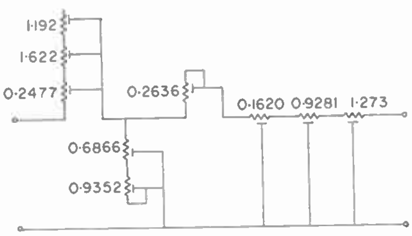
$$F_1 = \cosh(\sqrt{s}) \quad \text{and} \quad F_2 = (\sqrt{s}) \sinh(\sqrt{s}).$$



(a)



(b)



(c)

Fig. 4. Realization of Example given by (12). (All values given are in ohms and are the total resistances of the different uniform lines.)

The realization of the given transfer function using uniform lines is given in Fig. 4(c).

3. Extension to Include Complex Transmission Zeros

If a transfer function given by (11) has complex transmission zeros such that the coefficients of $P(u)$ are all positive,† then a procedure similar to that of Guillemin for lumped RC synthesis may be used to realize (11) as a parallel interconnexion of structures similar to that of Fig. 3(b). This point will be illustrated by the following example.

Example:

Suppose we are required to realize the following transfer function using commensurate uniform lines of unity RC product:

$$T[u(s)] = \frac{(1-u)^{1/2}(u^2+u+1)}{(u+1)(u+3)} = (1-u)^{1/2}\cdot\frac{P}{Q}. \qquad \ldots\ldots(13)$$

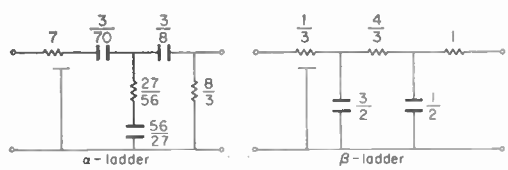
As there are complex zeros of transmission to be realized in the u-plane, we will use Guillemin's procedure. Choosing the arbitrary polynomial $d(u)$ to be $d(u) = (u+2)(u+4)$, and decomposing P so that each ladder will have its zeros of transmission on the negative real axis of the u-plane, we get

$$-py_{21\alpha} = \frac{(1-u)^{1/2}u(u+1)}{(u+2)(u+4)}; \quad -py_{21\beta} = \frac{(1-u)^{1/2}}{(u+2)(u+4)};$$

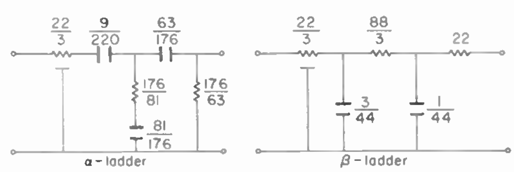
$$py_{22} = \frac{(u+1)(u+3)}{(u+2)(u+4)}.$$

The two ladder networks shown in Fig. 5(a) are realized by the method described in the previous Section. We note that in each case, the last impedance is that of a single short-circuited line. These ladders are then impedance-scaled according to Guillemin's procedure (Fig. 5(b)), and then the different one-ports are realized using uniform lines. The final realization of the given $T(s)$ is shown in Fig. 5(c).

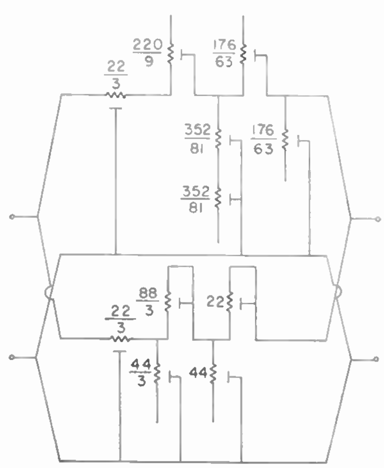
† This is equivalent[10] to stating that, if $P(u) = 0$, then $|\arg u_i| \leq (\pi/n)$, where n is the degree of $P(u)$.



(a) The ladder networks in the u-plane.



(b) The scaled ladder networks in the u-plane.



(c) Complete realization.

Fig. 5. Realization of $T(s)$ given by (13). (All values given are in ohms and are the total resistances of the different uniform lines.)

4. Extension of Some Known Transfer Function Synthesis Procedures to Include Symmetric Structures

We will now consider some procedures devised by different authors for the synthesis of transfer functions using uniform lines, and extend their results to synthesis using symmetric structures. We will show how each realization applies only to a limited version of the structure N of Fig. 2.

(a) The configuration used by Scanlan and Rhodes[3] is that of Fig. 3, namely, it consists of a single ladder network N_L followed by a single cascade network N_C. However, the series (shunt) impedances of N_L must be realizable as short (open) circuited cascades of uniform lines. This is a special case of our own procedure.

(b) A procedure developed by Wyndrum[1] does not permit the use of series elements, and demands that each shunt element be followed by a single cascade section of the uniform line. The main advantage of this realization is that it can be fabricated as a monolithic structure.

(c) Transfer functions realizable by the procedure of Stein, Mulligan and Shamis[8] satisfy conditions (a), (b) and (c) of our theorem. In their method, each subnetwork N_L is followed by a single cascade section of the uniform line. The total number of cascade sections must be greater than or equal to the number of zeros of transmission. Thus, a further restriction is imposed on condition (d) of our theorem. However, it is claimed that this configuration has certain advantages from the point of view of fabrication. By using a procedure similar to that of Guillemin, complex zeros of transmission in the u-plane may also be realized.

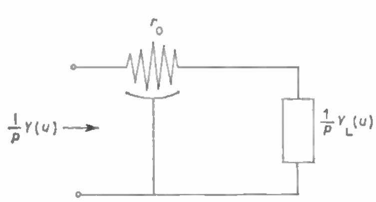


Fig. 6. A loaded symmetric structure.

This procedure may be extended to include synthesis using symmetric structures. Given a transfer function of the form (11), subject to the additional condition noted above, a suitable polynomial $d(u)$ is chosen such that (Q/d) is an RC admittance function in u. The quantities y_{22} and $-y_{21}$ are then identified as

$$\frac{y_{22}}{p} = \frac{Q}{d} \quad \text{and} \quad -\frac{y_{21}}{p} = (1-u)^{n/2} \cdot \frac{P}{d}.$$

Shunt and series elements arise from the normal ladder development of (y_{21}/p). These may then be synthesized as cascades of symmetric structures.

Now, only the realization of a cascade section remains to be considered. Expressing the chain matrix of a single symmetric structure in the form of (7), the admittance $(1/p)\,Y(u)$ of Fig. 6 is found to be

$$\frac{1}{p}\,Y(u) = \frac{g_0}{p}\left[\frac{g_0 u + Y_L(u)}{g_0 + Y_L(u)}\right] \qquad \ldots\ldots(14)$$

or

$$Y_L(u) = g_0\left[\frac{Y(u) - u g_0}{g_0 - Y(u)}\right] \qquad \ldots\ldots(15)$$

where $g_0 = (1/r_0)$. From (14) we see that

$$\lim_{u \to 1} Y(u) = g_0. \qquad \ldots\ldots(16)$$

In passing, we note again that if the symmetric structure is a symmetric line,

$$\lim_{s \to \infty} u(s) = 1.$$

Now, using (16), eqn. (15) becomes†

$$Y_L(u) = Y(1)\left[\frac{Y(u) - u\,Y(1)}{Y(1) - Y(u)}\right]. \qquad \ldots\ldots(17)$$

This form is identical to that obtained by Stein, Mulligan and Shamis for uniform lines. Thus, their procedure may be extended to include symmetric structures.

Wyndrum's procedure may be considered as a special case of that of Stein, Mulligan and Shamis, or it may be established directly by considering the reactance form (8) for the chain matrix of a symmetric structure. In this case,‡

$$Y_L(u) = Y(1)\left[\frac{Y(u) - u\,Y(1)}{Y(1) - u\,Y(u)}\right], \qquad \ldots\ldots(18a)$$

where now

$$u = \sqrt{(p/q)}. \qquad \ldots\ldots(18b)$$

5. Conclusions

This paper has considered, in detail, the realization of transfer functions using commensurate symmetric structures. Necessary and sufficient conditions were presented for the synthesis of a class of transfer functions using a three-terminal structure, which is a cascade of a number of networks of two different types N_L and N_C, N_L being a ladder network each of whose elements are cascades of symmetric structures, while N_C is itself a cascade of n symmetric structures.

† Equation (17) provides an alternative method of realizing a given driving point function as a cascade of symmetric structures.

‡ Using O'Shea's notation[2], Bassett[11] has similarly generalized Wyndrum's procedure. However, he did not consider the more general procedure of Stein, Mulligan and Shamis.

This was further extended to realize transfer functions as a parallel interconnexion of such structures. Also, the different existing synthesis procedures for transfer functions using uniform lines were extended to include symmetric structures. When symmetric lines are used, the structures discussed in this paper are capable of realizing transmission zeros at s equal to infinity, and hence may find applications in band limited networks.

6. Acknowledgments

This work was supported by the National Research Council of Canada, Grant Nos. A-2875, A-7315 and A-4399, and is based on a portion of the Ph.D. thesis submitted by J. C. Giguere under the supervision of Dr. M. N. S. Swamy, to the Nova Scotia Technical College, Canada.

7. References

1. Wyndrum, R. W., Jr., 'The Exact Synthesis of Distributed RC Networks', Technical Report No. 400–76, Department of Electrical Engineering, New York University, May 1963.
2. O'Shea, R. P., 'Synthesis using distributed RC networks', *I.E.E.E. Internat. Conv. Rec.*, **13**, Part 7, pp. 18–29, 1965.
3. Scanlan, J. O. and Rhodes, J. D., 'Realizability and synthesis of a restricted class of distributed RC networks', *I.E.E.E. Trans. on Circuit Theory*, **CT-12**, pp. 577–85, December 1965.
4. Lee, S. C., 'Synthesis of absolutely stable active distributed RCG networks', Proc. of the Eleventh Midwest Symposium on Circuit Theory, pp. 99–113, 1968.
5. Rhodes, J. D., 'Transfer function realizability of grounded URC networks', *I.E.E.E. Trans. on Circuit Theory*, **CT-14**, pp. 129–39, June 1967.
6. Giguere, J. C., Swamy, M. N. S. and Bhattacharyya, B. B., 'Driving point function synthesis using tapered RC-lines and their duals', Proc. of the First Asilomar Conference on Circuits and Systems, California, pp. 309–18. 1967. (Also *I.E.E.E. Trans. on Circuit Theory*, **CT-16**, pp. 93–4, February 1969).
7. Giguere, J. C., Swamy, M. N. S. and Bhattacharyya, B. B., 'Driving point function synthesis using nonuniform lines', *Proc. Instn Elect. Engrs*, **116**, pp. 65–70, January 1969.
8. Stein, J. J., Mulligan, J. H. and Shamis, S. S., 'Realization of transfer functions using uniform RC distributed networks with common ground connections', Proc. of the Symposium on Generalized Networks, Polytechnic Institute of Brooklyn, pp. 149–72, April 1966.
9. Protonotarios, E. N. and Wing, O., 'Theory of nonuniform RC-lines, Part I: Analytic properties and realizability conditions in the frequency domain', *I.E.E.E. Trans. on Circuit Theory*, **CT-14**, pp. 2–12, March 1967.
10. Seshu, S. and Reed, M. B., 'Linear Graphs and Electrical Networks', Chapter 8 (Addison–Wesley, New York, 1961).
11. Bassett, P. E., 'Cascade networks of symmetric proportional elements', *J. Franklin Inst.*, **285**, pp. 222–8, March 1968.

Manuscript received by the Institution on 4th December 1969. (Paper No. 1332/CC 84.)

U.D.C. 535.417:623.451

New Laser Interferometry Methods of Measuring the Velocity of High-Speed Model Missiles

By

H. D. VOM STEIN, Dr.Ing.,†

P. RATEAU, L. ès Sci. Appl.,†

G. SCHULTZE, Ing. grad.†

and

B. KOCH, Dr. rer. nat.†

Reprinted from the Proceedings of the I.E.R.E. Conference on 'Lasers and Opto-Electronics' held at the University of Southampton on 25th to 28th March 1969.

A modified Doppler technique is discussed, whereby a moving object is simultaneously illuminated by two intersecting laser beams which are coherent with one another. By comparing the different Doppler-shifted signals scattered, refracted, or diffracted from two overlapping light beams at the object, the velocity of high-speed missiles has been measured with good accuracy.

1. Introduction

Advances in laser design have made it possible to apply the well-known Doppler methods already used with microwave techniques for determining distance-time diagrams and velocity-time diagrams, to optical wavelengths. Thus, with this reduction of wavelength, the distance between two successive test points and hence the spatial resolving power have been reduced by about four orders of magnitude.

2. Methods Used Hitherto

In the simplest case, an optical Doppler apparatus consists of a Michelson interferometer using a c.w. laser as a light source (Fig. 1). The Doppler beat signal is obtained at the output of a photomultiplier by heterodyning the reference beam with the object beam. The frequency f_D of the Doppler beat signal which is determined by the wavelength 0·6328 µm of the laser and the velocity component v' of the object in the direction of the wave vector **k** may be assessed as follows:

$$f_D = \frac{2v'}{\lambda} \qquad \ldots\ldots(1)$$

Some numerical examples are shown in Table 1.

Table 1

λ	v'	f_D
0·6328 µm	1 m/s	3·13 MHz
0·6328 µm	100 m/s	313 MHz
0·6328 µm	10 km/s	31·3 GHz

It is seen from Table 1 that even for object velocities of only 100 m/s, recording and analysis of the resulting Doppler-shifted frequencies of approximately 300 MHz is not easy. On the one hand the Michelson type interferometers described in detail in References 1 and 2 are no longer applicable and on the other hand

† French-German Research Institute, 68 Saint-Louis, France.

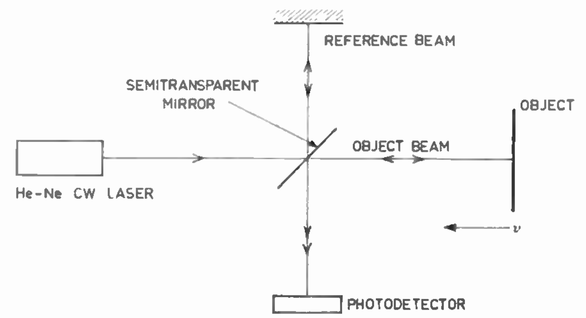


Fig. 1. Michelson interferometer.

restrictions are imposed by the cut-off frequency of the multiplier in the heterodyne stage.

Even the transient phenomena measuring techniques are subject to restrictions since difficulties arise in the adjustment procedure. These techniques are based on one of the following principles:

(a) The Doppler-shifted signal is superposed with the same signal delayed in time by a constant value T; thus it is possible to derive the acceleration of the object from the superheterodyned signal.[3]

(b) A passive Fabry-Perot resonator is used as a frequency discriminator.[4]

Reference 5 gives further explanations and details concerning these techniques as well as on the method mentioned by Dr. E. R. Pike (Royal Radar Establishment, Malvern) during the discussion of this paper.

3. Description of a Method Based on the Doppler Difference Frequency (Doppler Shift)

The following method is not only free from technical drawbacks but also particularly suitable for transient phenomena measurements.

The light from a c.w. laser is split into two partial beams which are coherent relative to each other and make the angles α_1 and α_2 with the velocity vector **v** of the object. Due to this, the Doppler frequency shifts corresponding to the two partial beams show different values at the point of the object:

$$\left.\begin{aligned} f_{D1} &= \frac{v}{\lambda}\cos\alpha_1 \\ f_{D2} &= \frac{v}{\lambda}\cos\alpha_2 \end{aligned}\right\} \qquad \ldots\ldots(2)$$

By heterodyning of the two partial beams on a photomultiplier, one obtains a beat signal whose frequency is given by the following equations:

$$\left.\begin{aligned} f_D^* &= f_{D1} - f_{D2} \\ f_D^* &= \frac{v}{\lambda}(\cos\alpha_1 - \cos\alpha_2), \end{aligned}\right\} \qquad \ldots\ldots(3)$$

On the basis of a suitable choice of the difference between $\cos\alpha_1$ and $\cos\alpha_2$, it is always possible, for given values of λ and v, to transfer the Doppler shifted frequency f_D^* into a region which is easily accessible to metrology.

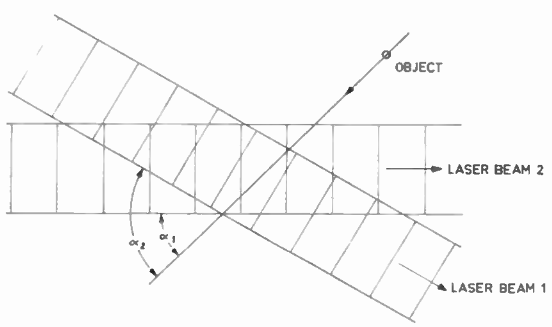
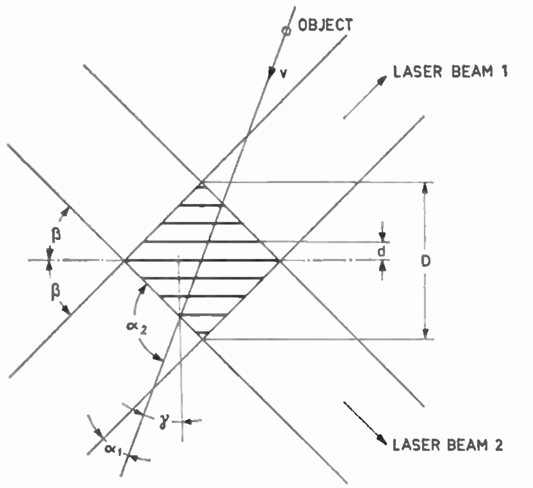


Fig. 2. Doppler difference method.

A somewhat different reasoning which is equivalent to the Doppler technique leads to the same result. Here the interference-fringes system is observed from the point of intersection of the two broad partial beams (Fig. 3). The distance between two succeeding interference-fringes is defined by the formula

$$d = \frac{\lambda}{2\sin\beta} \qquad \ldots\ldots(4)$$

If the object under investigation, whose linear dimensions are initially supposed to be small with respect to d, moves across the interference-fringes system in the indicated direction, a photomultiplier located somewhere in the area emits a signal every time the object passes through the maximum of intensity. The analysis of the output signal transmitted from the photomultiplier (the frequency of which must be identical with the Doppler frequency shift f_D), yields the absolute value of the velocity in the case of known λ, β, and γ.



$\gamma + \alpha_2 - \beta = 90°$
$\gamma + \alpha_1 + \beta = 90°$

Fig. 3. Doppler difference method.

$$\overline{f_D^*} = \frac{v}{\lambda} 2\sin\beta\cos\gamma \qquad \ldots\ldots(5)$$

It can be seen from Fig. 3 that

$$\alpha_1 + \gamma + \beta = 90°$$
$$\alpha_2 - \beta + \gamma = 90°$$

Consequently equations (3) and (5) can be regarded as equivalent. In the case of small angles of intersection β and of a constant α, a first approximation allows us to say that the number of interference-fringes is given by the width of the partial beams, i.e. by the diameter D of the field of interference-fringes (see Fig. 3). For a given performance of the laser, the choice of these values is essentially governed by the energy density which appears in the sectional area of the two beams emanating from the laser and which secures a sufficient signal/noise ratio at the output of the photomultiplier.

4. Experimental Set-up and Tests

The above method has been tested by firing model projectiles at velocities ranging from 40 to 100 m/s.

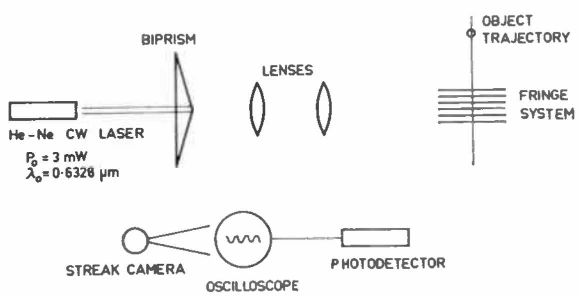


Fig. 4. Schematic diagram of experimental set-up.

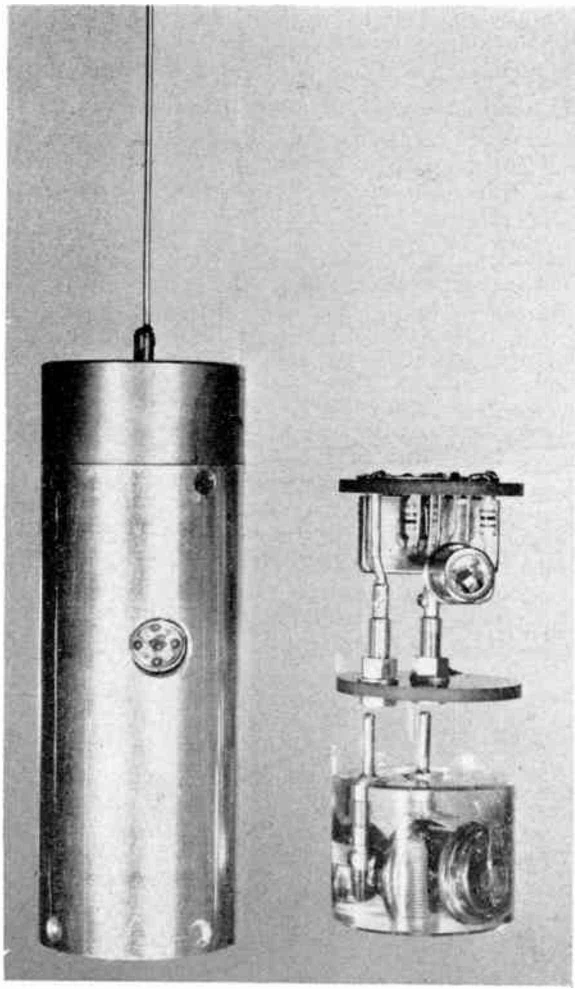


Fig. 5. The photomultiplier as mounted in a projectile.

The experimental set-up is shown in Fig. 4. By using a biprism, light of a He-Ne c.w. laser is split into two partial beams. A system of lenses causes the two partial beams to intersect at such a distance that the maximum extension of the sectional surface reaches about 22·5 cm. The number N of the interference-fringes was 96 and the distance d between the fringes was 2·52 mm ($\beta \simeq 7{\cdot}2 \times 10^{-3}$ degrees).

We distinguish between the following two methods:

(a) The first active method. Here the phototransistor and a telemetry transmitter are mounted in the projectile.

(b) The passive method. After reflexion from the projectile, the two partial beams experience heterodyning in a locally fixed photomultiplier.

The realizations of the two methods are as follows:

(a) A phototransistor cast in araldite is mounted in the wall of the projectile (Fig. 5). Its effective cathode surface is small compared to the distance d between the fringes. When the projectile passes through the field of interference-fringes, the output signal from the phototransistor modulates a small-size telemetry transmitter. After demodulation of the h.f. signal in a locally fixed station receiver the Doppler beat signal is recorded by oscilloscope and compared with a standard frequency, v is then obtained by applying equation (3).

(b) The two partial beams of the laser are reflected from the semi-spherically shaped, polished projectile ogive. Thereafter they undergo heterodyning in a locally fixed station photomultiplier. Recording and analysis of the Doppler beat signal are identical with those reported in (a).

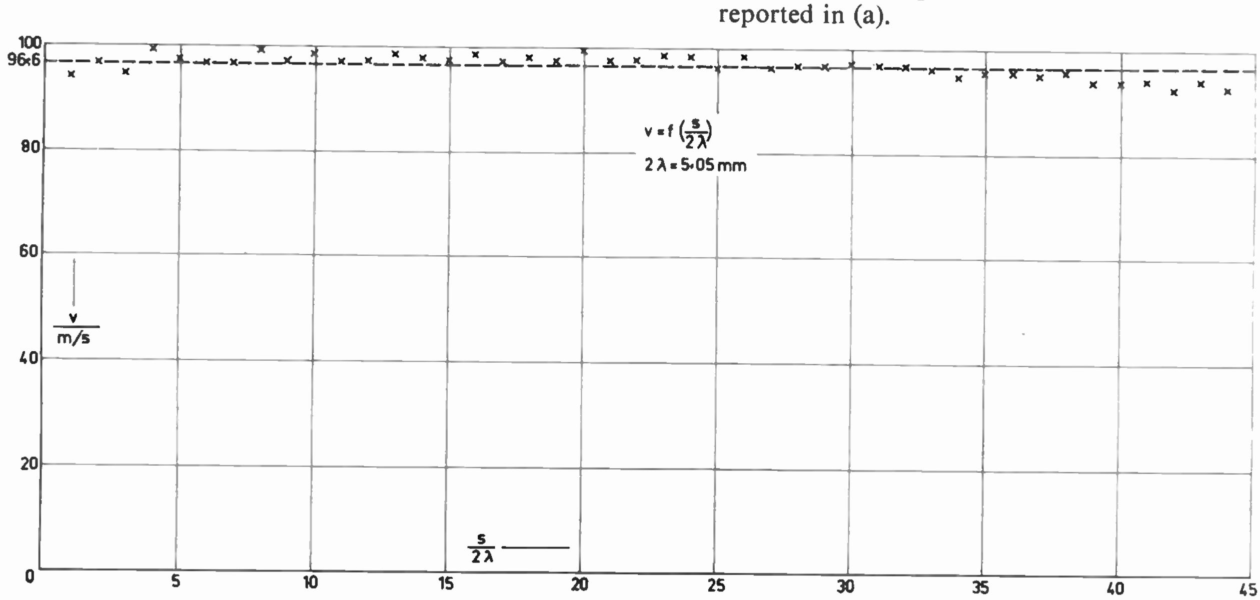


Fig. 6. Distance-velocity diagram.

In opposition to the previously made assumption, it may be shown that upon reflexion of the two partial beams from a spherical surface, the diameter $2R$ of the sphere must not be small compared to the distance between the interference-fringes. Moreover, a large distance between object and photomultiplier is to be assumed with respect to $2R$, since otherwise formula (5) would then no longer be valid.

A typical distance-velocity diagram of one shot is represented in Fig. 6. The standard deviation is 1·8%. In comparison with the value $\bar{v}_L$ determined by using light barriers, the average value $\bar{v}_{DD}$ obtained with the Doppler difference method shows an error of 5‰.

The first measuring error listed above is primarily due to the fact that in consequence of the optical arrangement used, the distance between the fringes towards the limits of the system does not remain constant and the edges of the fringes are somewhat blurred. The precision of the distance measurements limits the systematic variations between the two average values.

5. Conclusion

It may be concluded that the Doppler difference method described herein may be satisfactorily used as a measuring method for displacement-time analysis. Moreover by using simple instrumentation, this method may be easily adapted to current problem investigations. Even in the case of constant λ, the non-intermittent variation of d makes it possible to cover vast velocity ranges provided the reflexion factor for electromagnetic waves is sufficiently high and differentiated material constants allows distinction to be made between the object and the ambient area.

This method has been submitted to reduced-scale tests in our laboratory. Investigations are being carried out in order to extend this method to long-distance measurements.

6. References

1. Barker, L. M. and Hollenbach, R. E., 'Interferometer technique for measuring the dynamic mechanical properties of materials', *Rev. Sci. Instrum.*, **36**, p. 1617, November 1965.
2. James, R. N., Babcock, Wayne R. and Seifert, H. S., 'A laser-Doppler technique for the measurement of particle velocity', *Amer. Inst. Aeronautics Astronautics J.*, **6**, No. 1, p. 160, 1968.
3. Barker, L. M., 'Fine structure of compressive and release wave shapes in aluminium measured by the velocity interferometer technique', Symp. H.D.P. J.U.T.A.M., Paris, p. 483, September 1967. (Dunod, Paris, and Gordon & Breach, New York).
4. vom Stein, H. D. and Pfeifer, H. J., 'Direkte Messung von Geschwindigkeitsänderungen über kurze Wegstrecken durch Ausnutzung des Dopplereffekts', ISL-Aktennotiz N 10/68.
5. vom Stein, H. D., Pfeifer, H. J. and Koch, B., 'Geschwindigkeitsmessungen an kurzzeitigen Strömungsvorgängen mittels Laser-Strahlung', *Optics Communications*, **1**, No. 1, p. 207, 1969.

Manuscript first received by the Institution on 4th February 1969 and in final form on 10th November 1969. (Short Contribution No. 138/AMMS 31.)

U.D.C. 531.715.1:621.374.32

A Commercial Laser Interferometer for Length Measurement by Fringe Counting

By

B. W. BARRINGER, B.Sc.†

and

A. J. BONNER†

Reprinted from the Proceedings of the I.E.R.E. Conference on 'Digital Methods of Measurement' held in the University of Kent at Canterbury on 23rd to 25th July 1969.

The advantages of length measurement by fringe counting are described. The instrument employs a stabilized frequency source which maintains the frequency to ± 1 part in 10^7, for a temperature range of at least 26 deg C. The error rate due to amplifier noise in the fringe circuits is negligible, and 20% drift can be tolerated. A measuring accuracy of 3 parts in 10^6 in average atmospheric conditions is achieved with an economical wavelength to inch and millimetre conversion method.

1. Introduction

Accurate length measurement is fundamental to dimensional control of manufactured parts, both in the positioning of tools in machining operations and in the inspection of finished parts. The need for automatic position control of modern machine tools, and for rapid inspection of complex parts, has mothered the invention of numerous digital methods of measuring relatively long lengths (exceeding 5 mm), with high accuracy. All these methods employ divided scales of some form with optical, electromagnetic or electrostatic sensing transducers. Amongst them may be mentioned optical gratings, linear inductosyns, shaft encoders used in conjunction with leadscrews, and multiple plate capacitance transducers. These all rely for their accuracy on precision manufacture, and are limited in their true digital resolution by their method of manufacture and the resolving power of the sensing means. For practical reasons, such scales are generally not made in long continuous lengths (e.g. exceeding 1 m), and long scales are produced by combining shorter segments. Some of these scales have the advantage of providing unique information at each digital interval,[1] which enables dimensions to be read out from static transducers. Other methods using counting techniques require relative translation between parts of the system to measure length.

The standard of length, the metre, is defined in terms of the wavelength of light from a standardized source, made possible by Michelson's method of counting interference fringes traversing an aperture when an element of the interferometer is displaced. Electronic sensing techniques applied to Michelson's interferometer have long made attractive the idea of an automatic counting interferometer for general length measurement.[2] The coherence limitations of conventional light sources, however, have prohibited the general use of such techniques in industrial measurement. The temporal coherence or frequency bandwidth of conventional sources[3] limits the lengths over which good fringes can be obtained to about 200 mm, and the poor signal/noise ratios obtained[4] limits the traverse speeds to less than about 1 mm/s.

The invention of the laser has made available light sources of such high temporal and spatial coherence that length measurement by automatic fringe counting is now entirely practical and its industrial use is growing.[5]

2. Advantages of Length Measurement by Laser Interferometry

Laser interferometry has some obvious advantages over other methods of length measurement:

(1) The laser sources employed (helium–neon gas lasers) generate at a frequency which is reproducible[6] to 1 part in 10^7. Although the wavelength is a function of the refractive index of the medium in the light path (generally air), the effects are known to such a high order of accuracy that measurements of length to about 1 part in 10^6 are quite feasible.

(2) Interferometer instruments can have a truly digital resolution of one quarter of the wavelength of the light employed, about 0·16 µm. Finer digital resolutions can be obtained by employing two or more phase shifted fringe signals.

(3) The range of measurement is large. Good fringes have been obtained[7] at interferometer path length differences of 100 m.

(4) Traverse rates can be high. Reliable operation at rates of 3 m/s have been reported,[8] although much slower rates are generally adequate.

Other, less obvious, advantages are:

(5) The light beams may be deflected by prisms or mirrors in order to make measurements in awkward and inaccessible situations.

† Rank Research Laboratories, Brentford, Middlesex.

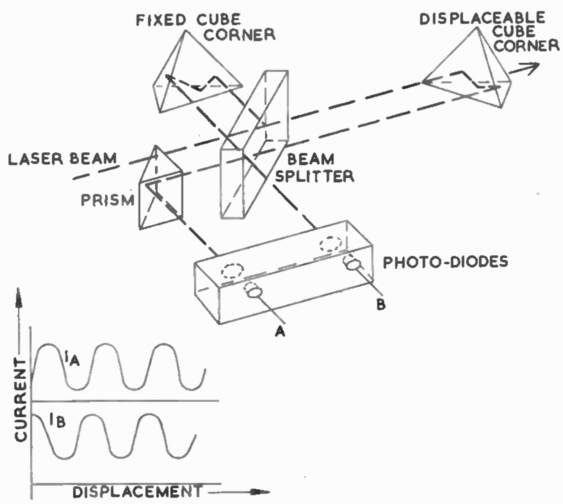


Fig. 1. The interferometer optics.

(6) Application to machines or instruments only requires alignment of a light beam along the axis of measurement. Careful supporting of scales, which may also suffer local deformation by heating from illumination sources, is avoided.

3. The Interferometer

The optical arrangements of laser interferometers are adequately described elsewhere[9] and because this paper is mainly concerned with circuit aspects, only a brief description is given here, for the sake of clarity. The optical elements used in the instrument are shown in Fig. 1.

Light from the laser source is collimated and directed onto a glass plate with a metallic partially transparent coating. The light is split into two paths by the coated surface, the transmitted beam travelling parallel to the incident light beam, and the reflected beam travelling at approximately right angles to the incident beam. The reflected or reference beam is returned to another position on the beam splitter by a cube corner prism reflector, fixed in position relative to the beam splitter. The transmitted beam is also returned to this position by a second, displaceable cube corner reflector and forms the measuring arm of the interferometer. The two returned beams interfere at the partially transparent surface and two light patterns are created which vary in intensity as the second cube corner reflector is displaced. These light variations or fringes are converted to electric current variations by the photosensitive diodes at A and B. The metallic coating employed introduces a phase difference between the two fringe signals, which enables the direction of displacement of the cube corner to be determined. Cube corner prisms have the well-known property of reflecting light parallel to the incident direction, irrespective of the angle which the entrance face makes with this direction. This property removes the need for accurate orientation of the reflectors and gives the instrument tolerance of angular variations during displacement of the reflectors. A bidirectional counter, controlled in its direction of counting by the relative phase of the two fringe signals, is used to total the wavelengths of displacement. Reversal of the direction of displacement automatically reverses the direction of counting, which is essential in situations where elements of the interferometer are subject to vibration.[2]

4. Frequency Control of the Laser Source

Long path difference interferometry requires a stable single frequency source. A source generating more than one frequency gives rise to path length differences of the interferometer where the interference effects of the component frequencies are out of phase, thus reducing fringe contrast (amplitude). Instability of the frequency results in wavelength instability and consequent imprecision of length measurement, as well as variations in source power. The laser source in this instrument is controlled to generate in a single axial mode at 633 nm (red light). The fact that this wavelength lies in the visible region is an advantage in setting up the instrument.

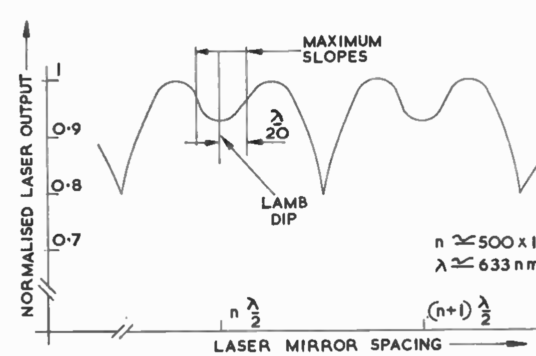


Fig. 2. The laser response to mirror spacing.

The frequency generated within the laser medium bandwidth is a function of the spacing of the resonator mirrors, an integral number of half-wavelengths fitting between them. The bandwidth of the helium neon medium is sufficient to allow oscillation at wavelengths covering a range of about 2 parts in 10^6. A curve of typical variation in laser output with mirror spacing is shown in Fig. 2. As the cavity length increases from n half-wavelengths to $(n+1)$ half wavelengths, the curve repeats, and passes through a region where a sudden jump in frequency takes place. The mirror spacing is therefore controlled to prevent temperature rise of the source from taking the laser through these length modes, and to reduce power output changes accompanying the wavelength variations.

A number of methods for controlling the laser frequency have been described.[10] Temperature stabilization of the laser cavity is commonly employed, but this suffers from the disadvantage of a time lag after switch-on before the cavity reaches its controlled temperature. The control system used in this instrument is similar to that described by Rowley and Wilson[11] and utilizes the shape of the laser response to mirror spacing to sense the laser centre frequency. The curve shows a decrease of power, known as the Lamb dip, at the centre frequency. A small amplitude sinusoidal modulation of the mirror spacing is applied by a piezoelectric transducer supporting one of the mirrors. The resultant modulation of the laser power is detected by a photo-sensor, amplified, and synchronously detected. This output is applied to the piezoelectric transducer in a sense which drives the operating point towards the Lamb dip. The maximum slopes of the laser response curve in the region of the Lamb dip occur at departures from the centre frequency of about 2 parts in 10^7, which correspond to cavity length deviations of ± 30 nm in 150 mm, the nominal mirror spacing. The frequency control error signal can only increase up to these points and therefore, if the source is under control, the frequency deviation cannot exceed ± 2 parts in 10^7. The maximum working range of the actuating transducer is about ± 2 μm so that a loop gain exceeding $2000/30 = 67$, is required for control over the full transducer range. In practice, loop gains of several hundred are obtained and the laser frequency deviations are less than 1 part in 10^7. The open-loop temperature coefficient of the laser cavity is less than 1 part in 10^6 per deg C, or 150 nm/deg C for the mirror spacing of 150 mm, so that the frequency control is effective over a temperature range of at least $\pm 2000/150 = \pm 13$ deg C. This is sufficient to control the effects of self-heating and ambient variations likely to be experienced by the instrument. If, however, temperature changes are large enough to take the control close to the end of its range, an alarm operates and the instrument read-out is extinguished until the control is manually reset to mid-range.

The modulation applied to the mirror spacing is about 5 nm peak at 5 kHz, causing a periodic wavelength change of about 3·3 parts in 10^8 peak. The interferometer is therefore subject to regular variation of the wavelength which causes the fringe patterns to execute cyclic variations. At the maximum designed difference of the interferometer reference and measuring paths of 6 m (= 3 m of reflector displacement), this effect causes an apparent path length vibration of 200 nm peak. This length 'vibration' has a peak velocity of about 6 mm/s which is only 1% of the maximum rate of path length change the counter will register. There is, therefore, an insignificant reduction in maximum traverse speed by the modulation of the laser wavelength. Much more serious effects may be caused by mechanically-induced vibration in the laser source structure and great care has been taken to make the structure very rigid, mounted in a way that minimizes these effects.

5. Fringe Signal Circuits

The fringe patterns produced in the interferometer generate currents in the photo-diodes which are sinusoidal functions of reflector displacement. A complete cycle of current change occurs for a displacement of one half-wavelength of the reflector, since the path length change is twice the reflector displacement. Measuring displacement at traverse rates of 300 mm/s ($\simeq$ 12 in/s) generates fringe signal frequencies of about 1 MHz. Detector current amplitudes of 3 μA are typical for laser power outputs of 100 μW. Commercial integrated circuit wideband operational amplifiers are used to amplify the fringe signals to suitable voltage and impedance levels for transmission to the counter circuits. Individual stages respond from d.c. to cut-off frequencies (−3 dB) exceeding 5 MHz.

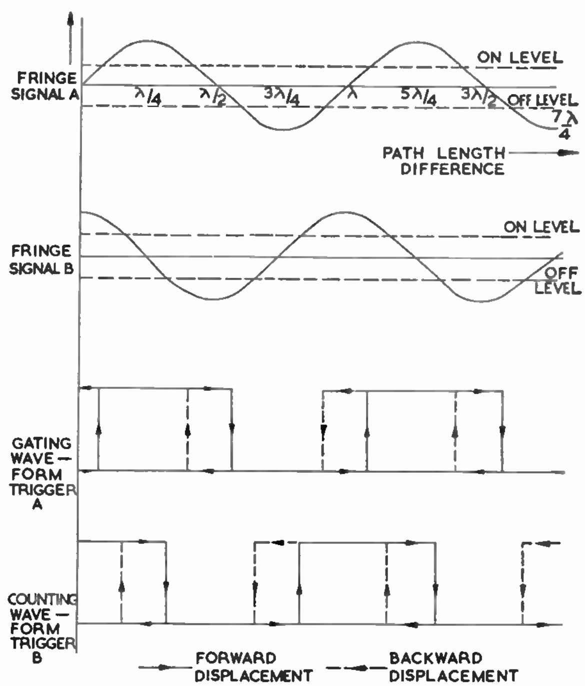


Fig. 3. Fringe and trigger waveforms.

Ideally the two fringe signals should have a phase difference of 90°, but the phase shift introduced by the beam splitter optimized for best overall performance, differs from the ideal. A 90° phase difference which is not a function of frequency is created by summing the two fringe signals, proportioned so that the sum is in quadrature with one of the components.[9] The quadrature fringe signals operate Schmitt trigger circuits which provide suitable waveforms to the bidirectional counter. Idealized waveforms are illustrated in Fig. 3. Rectangular waveform A is used as a gating signal for waveform B. A negative-going transition of waveform B during the 'on' period of A registers a forward count, but a positive-going transition of B registers a backward count. The need to register counts down to zero velocity of the reflector requires d.c. response in the signal circuits. The waveforms are consequently affected by zero drift, as well as by noise and phase errors.

5.1. *The Effect of Noise and Drift*

One factor which will affect confidence in using fringe counting instruments is the error rate caused by these effects. Noise in the signal waveforms, which arises almost entirely from the first amplifier stages, can cause rapid changes of state of the count trigger circuit if the trigger hysteresis is exceeded. A negative transition followed by a positive transition during the on period of the counter will register a forward count and a backward count, providing the interval between the transitions is long enough to allow for the total propagation delay of the forward/backward switching. Changes of state faster than this interval will cause erroneous registering of counts. It is therefore necessary to set a trigger hysteresis which is large enough to reduce the probability of counting from this cause to an acceptably low frequency. If the r.m.s. noise level, assumed to be white noise of 2 MHz bandwidth, is one-tenth of the trigger hysteresis, the average error rate will be less than one count per hundred hours. Measured noise levels are generally about one-third of this value so that a considerable safety margin exists.

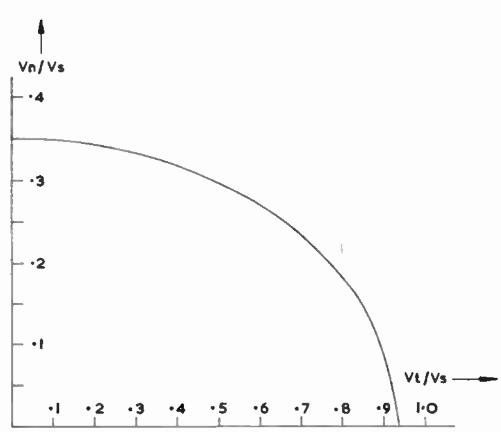


Fig. 4. Signal voltage error at which miscounting may occur against trigger threshold voltage.

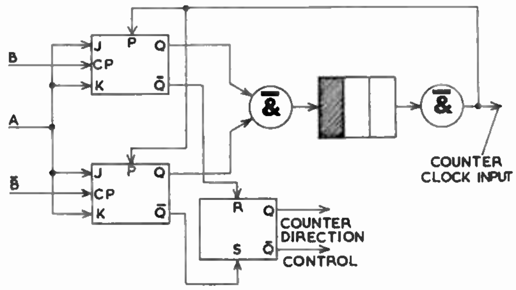


Fig. 5. Counter input circuit.

The input circuits of the counter described in the following paragraphs require a minimum time between the gate signal switching on and the arrival of a count transition. Under the ideal condition of zero drift, noise and phase error, the time between gate on and the count edge is one quarter of the counting period, i.e. 250 ns at the maximum counting rate of 1 MHz, but drift, noise and phase error can act in a way that reduces this time. The relationship of voltage and phase error to the trigger threshold (defined here as half the hysteresis or backlash) is derived in the Appendix. A curve of the allowable voltage error V_n, against the trigger threshold V_t, is shown in Fig. 4 for the worst case condition of a phase error of 20° and the minimum required gate to count time of 80 ns at a counting rate of 1 MHz. This curve shows that the amount of voltage error which can be allowed only reduces slowly as the trigger threshold is increased at low trigger levels. There is little advantage gained in tolerance to drift and noise by setting the threshold level to less than 20% of the signal amplitude, which is the level chosen in practice. At this value of trigger threshold, a combined voltage drift and instantaneous noise equal to 34% of the signal amplitude can be tolerated before miscounting at the fastest traverse rate occurs. With the noise level of the previous example, and in the presence of 20% drift, the probable error rate is about 1 count in 10^8. Again, a considerable safety margin exists in practice. Worst case drift equal to 20% of the signal amplitude corresponds to a temperature range of ±22 deg C, which is adequate for all practical purposes.

6. The Counter Input Circuit

It has already been shown that the bidirectional counter derives its direction of counting control by information gained from the fringe signals. The circuit which provides this information is shown in Fig. 5. Positive logic convention is adopted in the circuit descriptions. Two integrated circuit edge-triggered JK bistables are fed with clock input signals from the count waveforms and with common J and K inputs from the gate waveform. The clock input to one bistable is the complement of the clock input to the other. When the gate signal to the JK is a logical 0, changes of state on the clock input have no effect on the bistables. When the gate signal is a logical 1, the count input which changes from 0 to 1 changes the state of that bistable to which it is connected. Both bistables are initially set with output $\overline{Q}$ at 0, and the change of state of $\overline{Q}$ to 1 sets an RS bistable to a state dependent on which JK bistable has changed. A monostable is triggered by either JK bistable transition to provide sufficient delay for the direction control signals from the RS bistable outputs to take effect in the counter, the count being registered at the end of the monostable pulse period. The JK bistables are reset to $\overline{Q} = 0$ by feedback from the monostable pulse, in readiness for a new count input. This a positive-going transition at one clock input (negative-going complement at the other) sets the counter to count forward or up, and a positive-going transition at the other clock input sets the counter to count backwards or down.

7. The Wavelength to Inch and Millimetre Conversion

One of the penalties that has to be paid for using the wavelength of light as a measure of length is that length measurement in wavelength units is not acceptable for general industrial applications. The fringe count has to be converted therefore to inch or millimetre units. An economical method[12] was developed for this instrument which is sufficiently accurate for the majority of uses. Any number of input pulses, p, may be made to register a number, $q = p/n$, where n is a non-integer, by inhibiting the registration of pulses in predetermined ratios of p. It is only necessary to consider values of p lying between q and $2q$ since larger values can be reduced by a divider stage. Figure 6(a) shows one way of performing this conversion. Synchronous counter operation is represented by a common clock pulse line to all counter stages. An inhibit line prevents the registration of clock pulses in any stage to which it is connected when the inhibit signal is operative. Any stage which has all its carry inputs operative and the inhibit signal inoperative will register a clock pulse, i.e. the carry is the inverse of an inhibit.

Considering first the action of the inhibit counter of scale a, an inhibit signal appears at every ath input pulse and $(p-p/a)$ pulses are entered into counter 'q'. From the action of 'a' alone, a first approximation, q_1, to the desired value is given by

$$q_1 = p\left(1 - \frac{1}{a}\right)$$

The action of the second inhibit counter of scale b is similar, but there are values of p at which the inhibit signals of 'a' and 'b' coincide. The next approximation to q is given by q_2

$$q_2 = p\left(1 - \frac{1}{a} - \frac{1}{b} + \frac{1}{f}\right)$$

where f is the lowest common multiple of a and b. The factor $1/f$ makes this expression difficult to manipulate. An alternative method of conversion is shown in Fig. 6(b). Counter 'a' inhibits p/a pulses in both 'q' and 'b', and counter 'b' inhibits $1/b$ of the remaining pulses into 'q', i.e. counter 'b' inhibits $(p/b)(1-1/a)$ in 'q'. The approximation, q_2, for two inhibit counters is given by

$$q_2 = p\left(1 - \frac{1}{a} - \frac{1}{b} + \frac{1}{ab}\right)$$

Closer approximations may be obtained by adding further inhibiting counters.

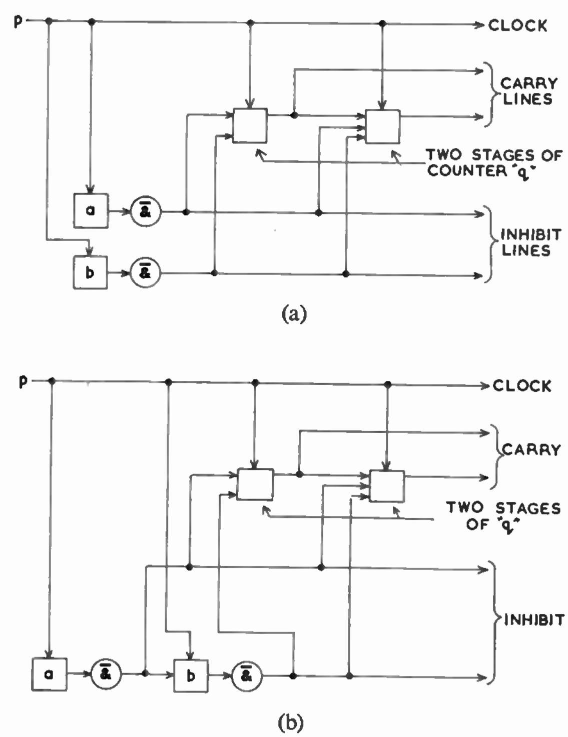


Fig. 6. Count conversion stages.

The method of Fig. 6(b) is used in the instrument to convert the fringe count to either inches or millimetres. Because it operates on a stream of pulses generated by the fringe signals, this method cannot convert a count registered in one length unit to a count in another length unit without translation of the reflector. The half wavelength of the helium neon laser at the Lamb dip in 'standard' air conditions of 20°C, 1013·25 mb and 50% relative humidity is 316·40985 nm. A metre corresponds to 3 160 457·9 half wavelenths or count trigger transitions. Inhibit counters of scales 21 and 302 allow 3 160 457·9 $(1-1/21-1/302+1/21\times 302)$ counts to register for displacement of 1 metre. Thus 2 999 993 counts register. The least significant stage is a scale of 3 counter followed by seven decades which indicate the displacement in micrometre units. The counter indication is low by only 2·3 parts per million in the atmospheric conditions quoted. The inch conversion is effected with one inhibit counter of scale 291 and registered in a least significant stage of scale 8 followed by six decades. The inch indication is low by 2·9 parts per million. The counter is constructed of conventional bidirectional stages using integrated circuit modules.[13] A maximum counting rate of 1 MHz allows length to be measured at traverse rates up to 300 mm/s and the direction of counting can be changed in about 1 μs.

8. Conclusion

The application of laser sources to interferometry has made it possible to construct instruments for measuring long lengths by automatic fringe counting. Frequency stabilization of the laser source by feedback from the output allows the instrument described to be used a few minutes after switch-on, and the servo has adequate range to control the effects of source warm-up.

The measured noise levels of the fringe signal circuits are well below those which give an acceptably low error rate. Drift equal to about 20% of the fringe amplitude can be tolerated before miscounting occurs at the maximum traverse rate of 300 mm/s. High accuracy of measurement in inch and millimetre units is achieved in specified atmospheric conditions by the use of an economical conversion method, and the counter circuits have adequate speed for practical traverse rates.

The method of measurement is not subject to factors which influence the accuracy of material scales, such as supporting forces, ageing and wear, coarse digital resolution and limited range. These factors give the instrument characteristics which make it suitable for industrial measurements of high precision.

9. Acknowledgments

The instrument described was developed for the Metrology Division of Rank Precision Industries Limited, and the authors thank the Directors for permission to publish this paper. The authors are grateful to W. T. Moore for assistance in preparation of the theoretical sections of the paper and to all members of the team involved in the development of the instrument.

10. References

1. Coates, L. J., 'A new measuring system for machine tools', *Machinery and Production Engineering*, **111**, pp. 832–5, 18th October 1967.
2. Branin, F. H., 'A bidirectional electronic counter for use in optical interferometry', *J. Opt. Soc. Amer.*, **43**, No. 10, pp. 839–48, October 1953.
3. Stroke, G. W., 'Photoelectric fringe signal information and range in interferometers with moving mirrors', *J. Opt. Soc. Amer.*, **47**, No. 12, pp. 1097–1103, December 1967.
4. Goldberg, J. L., 'Interferometry in length measurement', *Electronic Technology*, **39**, No. 5, pp. 186–91, May 1962.
5. Budzilovich, P. N., 'Lasers boost machine tool accuracy', *Control Engineering*, **15**, No. 12, pp. 62–6, December 1968.
6. Mielenz, K. D., Nefflen, K. F., Rowley, W. R. C., Wilson, D. C. and Engelhard, E., 'Reproducibility of helium-neon laser wavelengths at 633 nm', *Applied Optics*, **7**, No. 2, pp. 289–93, February 1968.
7. Morokuma, T., Nefflen, K. F., Lawrence, T. R. and Klucher, T. M., 'Interference fringes with long path difference using a helium neon laser', *J. Opt. Soc. Amer.*, **53**, No. 3, pp. 394–5, 1963.
8. Rowley, W. R. C. and Wilson, D. C., Design tolerances in laser measuring systems', *Proc. Instn Mech. Engrs*, **183**, Pt. 3D, 1968–69. (To be published.)
9. Rowley, W. R. C., 'Some aspects of fringe counting in laser interferometers', *Trans. I.E.E.E. on Instrumentation*, **IM-15**, No. 4, pp. 146–9, December 1966.
10. Birnbaum, G., 'Frequency stabilization of gas lasers', *Proc. I.E.E.E.*, **55**, No. 6, pp. 1015–26, June 1967.
11. Rowley, W. R. C. and Wilson, D. C., 'Wavelength stabilisation of optical masers', *Nature*, **200**, pp. 745–7, 23rd November 1963.
12. Stillwell, P. F. T. C. and Barr, J. D., British Patent Application No. 12827/66.
13. 'Micrologic Handbook', SGS-Fairchild, September 1965.

11. Appendix: The Relationship of Drift, Noise, and Phase Error to the Trigger Hysteresis

It is assumed that both fringe signals are of equal amplitude V_s, and that equal trigger levels V_t are set for both signals. V_t is equal to half the trigger hysteresis. The input to the gate trigger, V_g, in the presence of an error voltage, V_{n1}, is given by

$$V_g = V_s \sin\theta - V_{n_1}$$

where $\theta = \dfrac{2\pi s}{\lambda}$

s = path length change

λ = wavelength of light.

The gate-on level, V_t, is reached at an angle, θ_g, given by

$$V_s \sin \theta_g = V_t + V_{n_1} \qquad \ldots\ldots(1)$$

Similarly, the input to the count trigger, V_c, in the presence of error voltage V_{n2}, is given by

$$V_c = V_s \sin\left(\theta + \frac{\pi}{2} + \phi\right) - V_{n_2},$$

where ϕ = phase difference from the ideal quadrature relationship between the waveforms.

The forward count-off trigger level, $-V_t$, is reached at an angle θ_c, given by

$$V_s \sin\left(\theta_c + \frac{\pi}{2} + \phi\right) = -V_t + V_{n_2} \qquad \ldots\ldots(2)$$

The angle, θ_s, between gate-on and the count transition is given by

$$\theta_s = \theta_c - \theta_g \qquad \ldots\ldots(3)$$

θ_s is a minimum when θ_c is minimum and θ_g is maximum.

These conditions are given when $V_{n1} = V_{n2} = V_n$ where V_n is the worst case drift of each channel and is positive, and ϕ is positive. Then, from (1), (2) and (3)

$$2V_t = V_s\,[\sin \theta_g - \cos(\theta_g + \theta_s + \phi)] \qquad \ldots\ldots(4)$$

$$2V_n = V_s[\sin \theta_g + \cos(\theta_g + \theta_s + \phi)] \qquad \ldots\ldots(5)$$

θ_s has a minimum value fixed by the characteristics of the circuit of Fig. 5, and the maximum counting rate. ϕ is the maximum phase error between the fringe signals at the highest signal frequency. A curve of V_n/V_s against V_t/V_s is obtained by substituting values for θ_g. The error voltage V_n may be considered to be the sum of drift voltage and instantaneous noise.

Manuscript first received by the Institution on 5th May, 1969 and in final form on 11th March 1970 (Paper No. 1333/IC27.)

The Authors

A. J. Bonner joined Rank Research Laboratories after a period with Avo Ltd. He has been closely associated with all aspects of the laser interferometer design, particularly with the frequency control and digital circuits. He has also been responsible for the design of other instruments using optical transducers to sense mechanical displacements.

B. W. Barrington qualified in electrical engineering (communications) at the City and Guilds College (Imperial College), London. After completing post-graduate training with Standard Telephones & Cables Ltd., he joined the Company's Telephone Apparatus Laboratory where he was concerned with telephone transmitter and hearing aid receiver design. He is at present a senior engineer in Rank Research Laboratories where he has been associated with many metrology and control R & D projects.

Recent Contributors to the Journal

K. G. Nichols (M.1960, F.1967) obtained London University degrees of B.Sc. in physics in 1949, B.Sc. in mathematics in 1958 and M.Sc. in mathematical physics in 1959. He is at present a senior lecturer in the Department of Electronics, University of Southampton, and he takes up an appointment as reader in the Department in October next. Before going to Southampton in 1961 he was a senior lecturer at Letchworth Technical College. After graduating he held appointments as development engineer with E.M.I. Research Ltd., and S.T. & C. Ltd. Mr Nichols served for several years on the Institution's Examination Committee and he has recently been appointed to the Education and Training Committee; he has also been an active member of the Southern Section Committee and was associated with the planning Institution and joint conferences. He represented the Institution on the Technical Committee of B.S.I. concerned with integrated circuits and on working parties of the Council of Engineering Institutions. In addition to numerous papers in the Journals of the Institution and other bodies he is joint author of a book on transistor physics.

T. S. Plews (M. 1970) graduated from Kings College, Newcastle University (then part of Durham University) in 1959 with a B.Sc. degree in physics and mathematics. He joined the Electronics Department of the Guided Weapons Division of what is now the British Aircraft Corporation in the same year. While employed by B.A.C. he worked for an M.Sc. in electronics at Southampton University which he was awarded in 1967.

C. R. Barber (G. 1959) who died shortly after the completion of the final manuscript of the paper of which he was the joint author, was an experimental officer in the Department of Electronics at the University of Southampton. Before taking up this appointment in 1967 he was a member of the experimental staff at the G.E.C. Applied Electronics Laboratories, Stanmore. Mr. Barber obtained his Higher National Certificate with endorsements following studies at the R.A.E. Technical College, Farnborough and at Hendon Technical College.

The paper 'A Micrologic Vector Generator' by Messrs. Plews, Barber and Nichols was published in the June issue of *The Radio and Electronic Engineer*.

International Broadcasting Convention 7th-11th September 1970

The 1970 International Broadcasting Convention will be held at Grosvenor House, Park Lane, London, W.1, and will comprise an exhibition of the latest equipment in the field of broadcast engineering (both sound and vision) and a concurrent programme of technical sessions.

The Sponsors of the Convention are: Electronic Engineering Association, Institution of Electrical Engineers, Institute of Electrical and Electronics Engineers, Institution of Electronic and Radio Engineers, Royal Television Society, Society of Motion Picture and Television Engineers.

PROVISIONAL PROGRAMME

Monday, 7th September
Introductory Session
Session 1—Studio Design
Session 2—Origination Equipment

Tuesday, 8th September
Session 2 (continued)—Origination Equipment
Session 3—Recording and Films
Session 4—Digital Techniques

Wednesday, 9th September
Session 5—Automation
Session 6—Signal Distribution

Thursday, 10th September
Session 7—Receivers
Session 8—Transmitters and Transposers

Friday, 11th September
Session 9—Aerials
Session 10—Propagation, Service Planning and Satellites
Session 11—Operational Experience
Session 12—Closing Discussion

The Convention fee of £12 10s. 0d. entitles registrants to attend the technical sessions and exhibition, to receive a copy of the publication containing the texts of the contributions included in the technical programme, to attend the Cocktail Party, and to receive mid-morning and afternoon refreshments.

All enquiries concerning the Convention should be made to:

The Secretariat,
International Broadcasting Convention,
c/o I.E.E., Savoy Place, London, WC2R 0BL.
(Telephone: 01-240 1871 Ext. 261. Telex: 261176).