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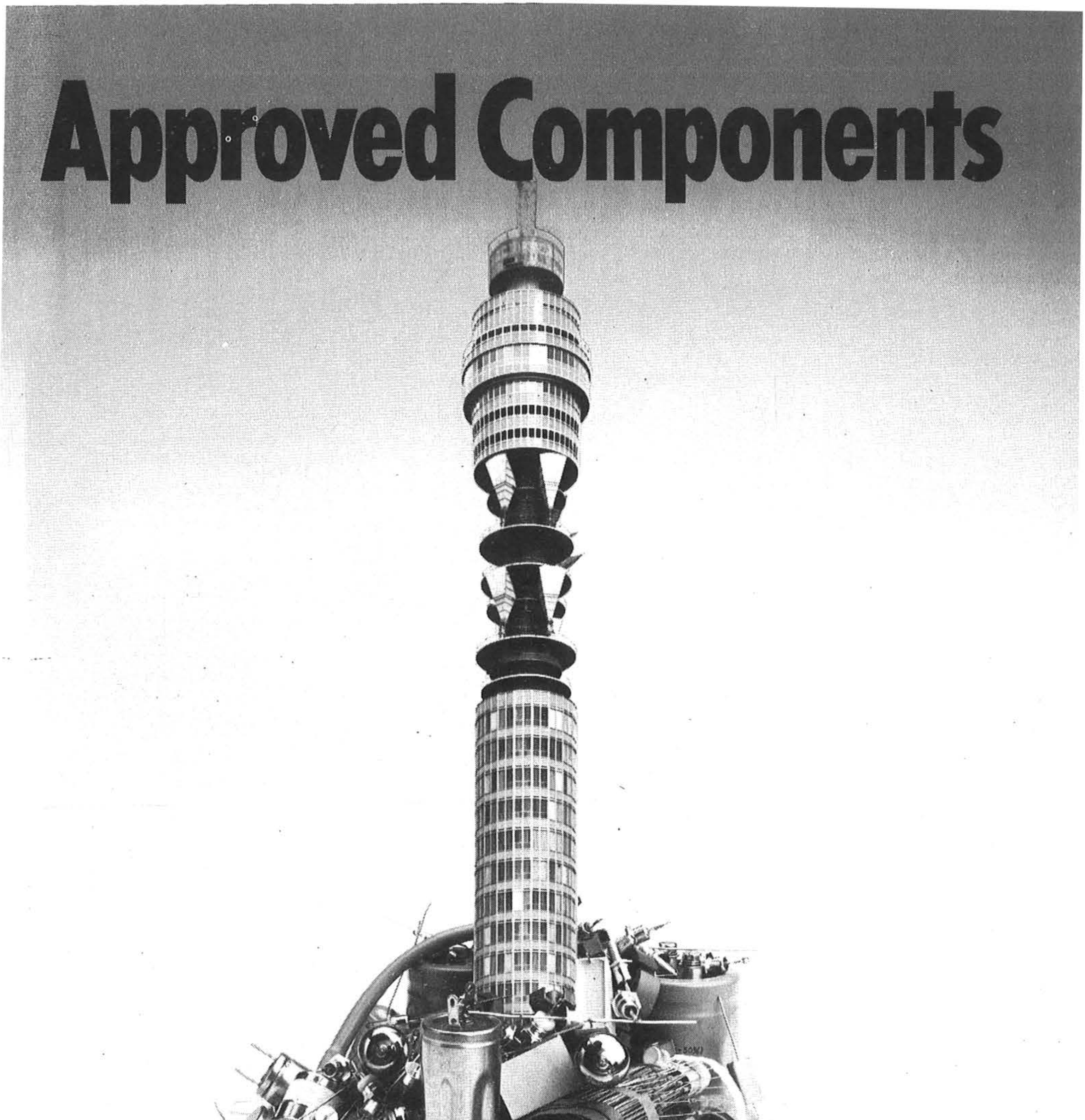
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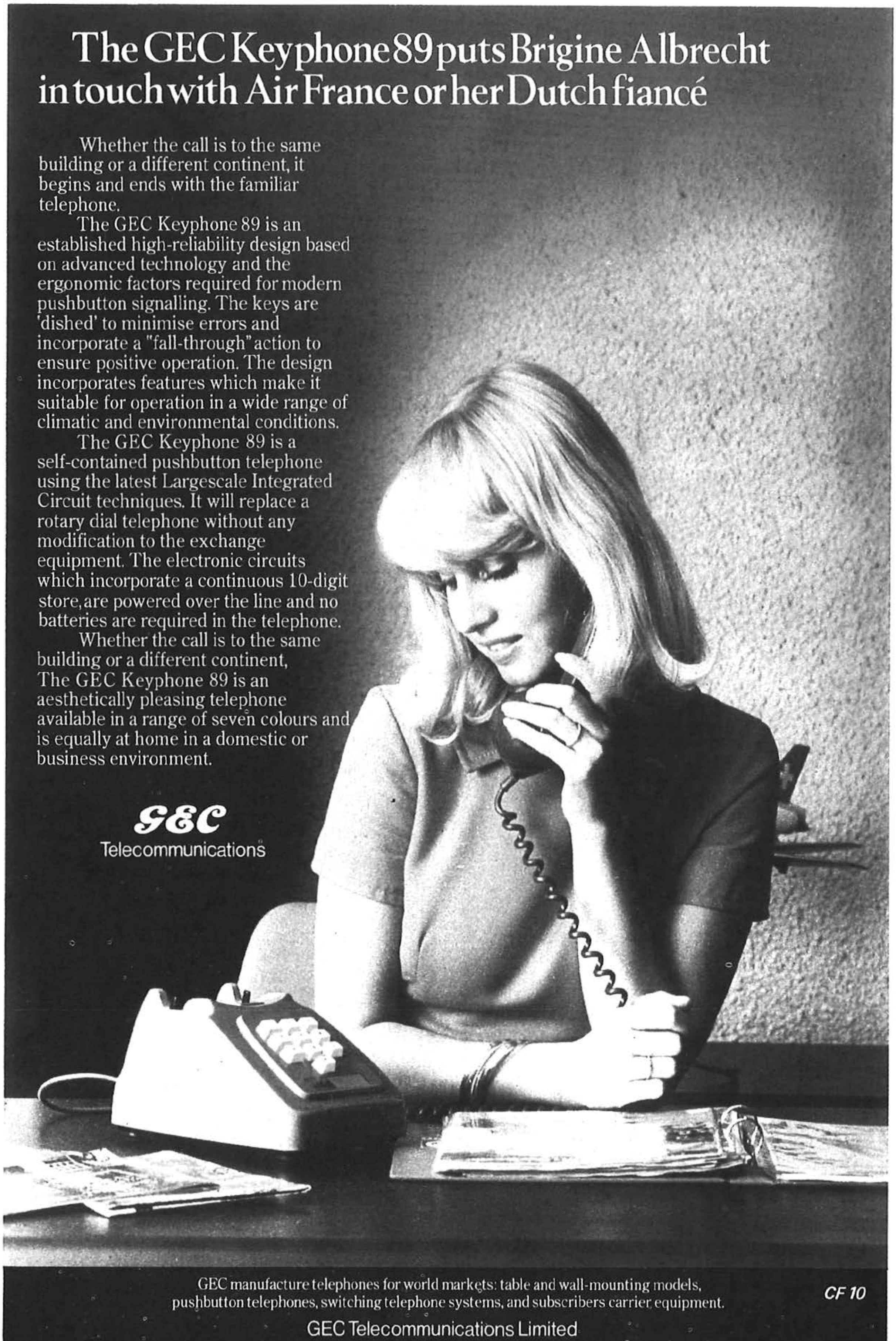
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The Post Office Electrical Engineers' Journal

EDITORIAL

The development of postal mechanization is a continuous process. From the introduction of the first stamp-cancelling machine (*circa* 1859) to the present-day mechanized letter-offices and parcel-concentration offices, the motivation of postal engineers has been the development of machines and facilities to ease the physical effort required of postal workers and improve the efficiency of the postal services. In latter years, the influence of technology has extended beyond the direct substitution of manual processes by their automatic equivalent to affect the traditional way of handling mail.

Under the British Post Office's (BPO) letter post plan, many items of mechanized plant have been introduced in recent years and are in common use: for example, automatic letter-facing machines, segregators, coding desks and automatic letter-sorting machines. In all such developments, ergonomic factors are given the utmost consideration; conditions at the man/machine interface, physiological and environmental factors, all influence the design and development of plant used in the postal business.

An example of the influence of ergonomics and other related factors on the design and realization of plant in the field of postal engineering is given in the article on page 70 of this *Journal*, which describes the 'easy-view' coding desk that is being installed by the BPO in its new mechanized letter-offices.

An Easy-View Letter-Coding Desk

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UDC 681.187: 656.851: 658.3.04

Letter-coding desks must be ergonomically sound and provide a pleasant working environment, particularly as coding duties occupy a considerable part of the desk operator's working life. Because coding desks are needed in greater numbers than any other machine in postal mechanisation, they must be of low cost. A novel type of coding desk, for which human factors and mechanical simplicity were the main design aims, is described in this article. This coding desk has now been adopted by the British Post Office as the standard for equipping its mechanized letter-offices.

INTRODUCTION

A letter-coding desk is the equipment used by a postman to print machine-language addresses on items of letter mail so that, subsequently, the letters can be sorted automatically, thus saving manpower. Letters are presented to the coding-desk operator who copies the public postcode (written by the sender) onto a typewriter-style keyboard. The coding-desk control system consults a computerized library of addresses known as a *translator*[1]; from the translator is obtained a machine-language version of the address, which is printed on the envelope in the form of a binary code of luminescent dots[2].

The first generation of coding desks used by the British Post Office (BPO) evolved from the letter presentation unit of an earlier single-position letter-sorting machine[3], and included many complex and costly mechanisms[4]. Because coding desks are required in considerable numbers, the BPO reviewed all aspects of coding-desk design; in particular, attention was given to operator comfort, mechanical simplicity, reliability, cost and compatibility with other machines in the letter-sorting system. From this review, a number of operational requirements and engineering design aims emerged.

A study of the main operational requirements established the following points:

(*a*) Unlike the previous designs of coding desks, which had a height of approximately 2 m, the new coding desk had to be designed to a height approximating to that of an office desk.

(*b*) The coding desk should be ergonomically matched to the operator.

(*c*) The capital cost should be low.

(*d*) The floor space taken up by the coding desk and its operator should not exceed that of the earlier types of coding desk.

(*e*) The supply of letters to a suite of coding desks should be replenished automatically from a single letter-feed point. (This requirement was later dropped in favour of the hand loading of individual coding desks.)

(*f*) The coding desks should fit together to enable a coding suite to be assembled, and positioned such as to be compatible with the office traffic-flow pattern, thus minimizing the transportation of mail within an office.

(*g*) After coding, letters from all the desks forming a suite should be collected by a single aggregating belt and fed automatically to letter-sorting equipment.

(*h*) The intrinsic rate (that is, the rate at which letters can be presented to a coding-desk operator) should be fast enough to prevent frustration of the fastest operator.

To the operational requirements were added the following engineering design aims:

(*a*) simple and reliable mechanics,
(*b*) low maintenance cost,
(*c*) use of established letter-handling principles, and
(*d*) minimal effect on the sorting office environment; in particular, a low noise level.

ERGONOMICS

During the examination of ergonomic factors, the method of letter presentation was considered to be the most important area in which change was necessary (when compared with the existing design of coding desks); and it was from this aspect that the new design of coding desk evolved.

One of the prime aims of the coding-desk design was to ensure that it would fit the person required to operate it. The aspects considered fell into 3 groups: letter-presentation factors; physiological factors (such as, leg space and elbow room); and environmental factors (such as noise and heat).

Letter Presentation

The first consideration for the design of the new coding desk was the method of presenting letters to the operator. It was observed that, under prescribed conditions, it was possible for a coding-desk operator to read an address while a letter was moving.

It had already been established (from experience of earlier models of coding desks) that, if more than one letter was on display at a time, the operator's output could be increased. Also, it was known that machines that presented letters to an operator's view at a fixed rate were unacceptable because of the variation in time required to code individual letters.

Human eyes are able to range more comfortably over a greater angle laterally than they can vertically. If letters are presented to an operator in the form of a continuous stream moving from right to left at a speed controlled by the operator, then natural reading ensues; this fact was confirmed by subjective tests, and the method was found by experienced operators to be superior to other modes of presentation. It was from this feature that the name *Easy-View Coding Desk* was derived. Furthermore, this simple direct-viewing system was preferred on grounds of readability and operator access

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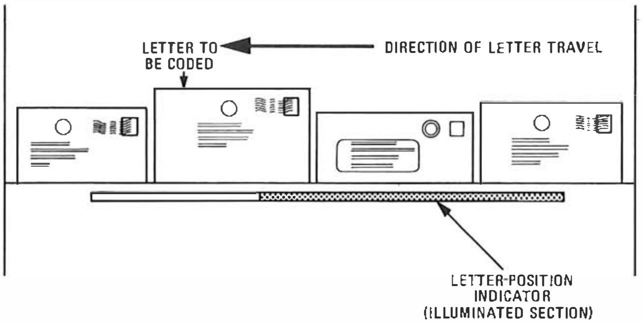


Fig. 1—Presentation of letters to coding-desk operator

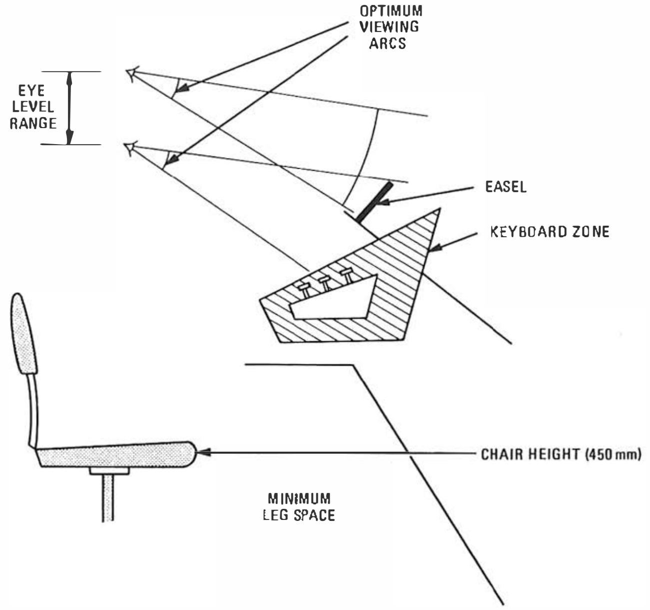
to the letters. Fig. 1 shows the method whereby a queue of letters moves from right to left across the field of view of the operator.

The moving queue of letters is totally unlike the presentation system of other coding desks, in that there is no fixed location for a letter when coding takes place. Therefore, it is imperative that the position of the letter to be coded is indicated to the operator. This is achieved by means of a *letter-position indicator*, which takes the form of a ribbon of light that extends from the letter entry point on the right to the letter to be coded; thus, all uncoded items are underlined. Naturally, all the letters forming the queue move at the same velocity across the presentation zone; the velocity is determined by the position of the letter being coded. If all letters on view have been coded, the queue velocity will be at a maximum of 20 m/min. An uncoded letter entering from the right will cause the queue speed to reduce and, if no code keying takes place, the queue speed will reduce to zero with the letter to be coded positioned on the left-hand side of the presentation zone. Code-keying can take place at any time after a letter has entered the presentation zone. When the coding of a letter has been completed, the speed control reverts automatically to the next letter to be coded. The mean velocity of the moving queue is dependent on the position of each letter when coded. By matching the velocity of conveyance to the reading task, the angular movement of an operator's eyes is minimized.

Code-keying and address reading are 2 functions that can be carried out simultaneously, but independently, by an experienced operator; that is, an operator can key the code for one letter while reading the address of the next letter following. This faculty is known as *pre-reading* and increases operator throughput.

Physiological Factors

The new letter-presentation principle would prove successful only if other ergonomic factors (for example, viewing angle and distance) could be optimized for the wide range of physical characteristics of the postmen who operate the desks. Based upon the initial assumption of the use of a standard (450 mm) high office chair, and the use of anthropometric data, the critical dimensions for the desk, in terms of arm and leg reach, range of eye levels etc., were determined. It was found that a fixed height of letter presentation, combined with a fixed-angle easel of length 60 cm, gave the vast majority of the operators an optimum viewing angle and distance (assuming of course, that the operators have normal or corrected vision). Although humans are adaptable to reading under wide ranging conditions, it has been established that the most relaxed position for reading is with the head and eyes angled slightly down. Having set the nominal height of the chair and the position of the viewing easel, it was found that an adjustable keyboard level was required to accommodate the range of personnel who operate the desk. A generous leg-room requirement was a determining factor in the size of the desk and precluded some of the early component layouts. The letter-viewing and seating arrangements are shown in Fig. 2.



(*a*) Letter-viewing arrangements

(*b*) Coding desk

Fig. 2—Coding-desk letter-viewing and seating arrangements

The basic design of the presentation zone has the added advantage that little glare or specular reflection from the letters occurs. Therefore, there is no need for individual desk lighting, provided that ambient lighting in the coding-desk area is correctly positioned and provides an illumination of about 330 lx. This factor is significant because glare can produce fatigue and frustration in operators. (Glare can be a particular problem with window-type envelopes.) An operator has direct access to the letters; this facility enables removal of any letters which are not suitable for the subsequent automatic letter-sorting process; for example, re-used envelopes that already bear code-marks.

Environmental Factors

High noise pressure levels in a working environment are not conducive to accuracy, and experience with early coding-desk equipment showed the need to minimize noise at source rather than recourse to acoustic suppression treatment later. The major potential sources of high levels of noise were motors, gearboxes, a vacuum pump, and the code-mark printer mechanism. The method of letter presentation has been achieved by use of an exceedingly simple mechanical arrangement that, in the main, uses low-speed components. The use of small motors, mounted independently from the machine framework, reduced the level of vibration transmitted to the main machine surfaces, with the result that the machine noise level is below 70 dBA. The simple mechanical design of the coding desk results in a low heat dissipation and avoidance of operator discomfort. The low height of the desk ensures a pleasant open-style office layout, and the non-vertical method of presentation ensures that no local lighting is required.

MECHANICAL DESIGN

The Easel

The presentation section of the desk is an easel, on which the sill supporting the bottom edge of the letters takes the form of a narrow conveyor belt. To ensure adequate drive to the letters, the friction of the easel material has to be as low as possible, and that of the sill belt as high as possible. The frictional requirements of this simple form of drive dictate that a letter's angle of repose is controlled within a narrow sector, which fortunately includes the optimum viewing angle. It is necessary to correctly register the bottom edge of each letter for code-mark printing. This is achieved by the sill belt which is used as the base reference. The letter-position indicator is sited immediately beneath the sill drive belt, and is well within an operator's field of vision.

Equipment Mounting Planes

An equipment and seating arrangement which provides a space to accommodate the operator's legs is illustrated in Fig. 3, and affords a natural plane upon which to mount the desk components. A second equipment-plane, mounted at 15° to the vertical, is provided to carry the replenishment components and facilitate the inevitable cross-overs between the input mail from the common replenisher feed-unit and the mail discharged onto the aggregating conveyor (see Fig. 4). The components of the letter path are mounted on 2 aluminium plates which form the equipment planes.

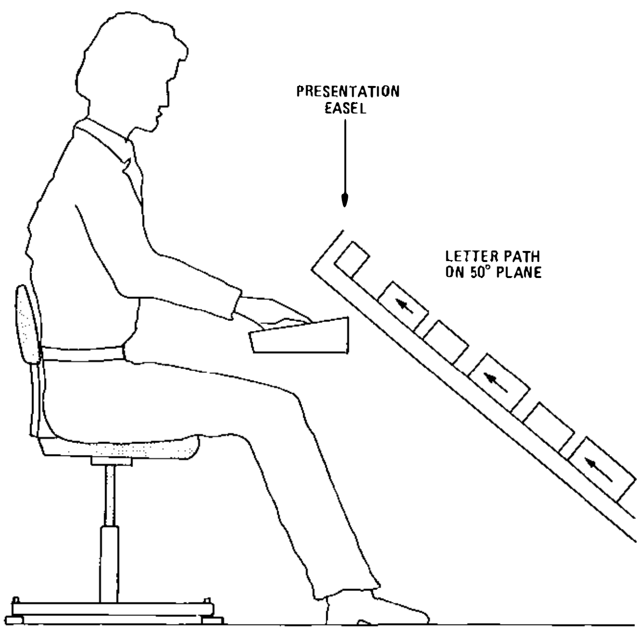


FIG. 3—Letter-feed presentation

Replenishment Systems

Automatic-Replenishment, Destacker and Transport System

The automatic-replenishment transport system consists of a common highway running the entire length of the coding-desk suite, with a distributor branch to each desk. Batches of about 30 letters are despatched from the feed unit to the coding desks in the form of short streams of overlapping letters, known as *tiles*. A call for mail from a particular coding desk may take some time to satisfy because of the transit time along the highway, and because priority is given to any other desk calling at the same time and further from the replenisher feed-unit. To avoid delaying an operator due to lack of letters, the branch to each desk was engineered as a separate belt-system, known as the *back-up store*, from which a destacker can receive an immediate supply of letters. When the mail in the destacker falls to the *re-order* level, the desk registers its call for more letters from the replenisher feed unit, and signals the back-up store to run its waiting tile of letters into the destacker. The replenishment sequence is completed when the back-up store runs briefly for a second time to receive the replacement tile.

To make the desk as compact as possible, the diverter and part of the back-up store are mounted on the adjacent up-stream desk and, in the case of the coding desk adjacent to the replenisher feed unit, a back-up store and diverter are incorporated within the feed unit.

The automatically-replenished destacker, which is similar to that used in earlier postal equipment, is shown in Fig. 5, and operates on the principle whereby the bottom letter is pulled out by a vacuum-assisted friction belt while the next letter is held back by rubber snubbing rollers. These rollers must remain virtually stationary as far as the letters are concerned, but allowed to rotate very slowly to even out wear. In the easy-view coding desk, the snubbing rollers have been replaced by spindle-mounted polyurethane spheres whose axes are inclined at a small angle to the letter path so that a small component of the letter velocity is available to cause the spheres to rotate slowly. This innovation[5], which is illustrated in Fig. 6, has proved effective and cheap. To minimize the presentation velocity for any given throughput, it is important to keep the average length of gap between items small. However, as the gaps produced by the destacker may vary, the control system has been designed to accommodate small overlaps (30 mm) that may occasionally occur.

The letters are transported from the destacker to the presentation section sandwiched between 2 belts. The transfer from the 15° to the 50° plane is achieved by the simple artifice of twisting the transport belt through 35° along the letter path axis which is common to both planes.

Hand-Load Replenishment System

To provide operational flexibility and cost savings, the design of a hand-loaded desk was undertaken. This requirement was met by replacing the automatic-replenishment system with a large letter-stack (see Fig. 7). The added cost of the individual stacks, each accommodating nominally 800 letters, is more than outweighed by the saving of the replenishment components and the feed unit. The stack rests on a conveyor belt, which is incremented towards the destacking unit under the control of a letter-position sensor. Vertical destacking has proved more efficient than the horizontal method of destacking. This is a valuable improvement in presentation of items to the operator.

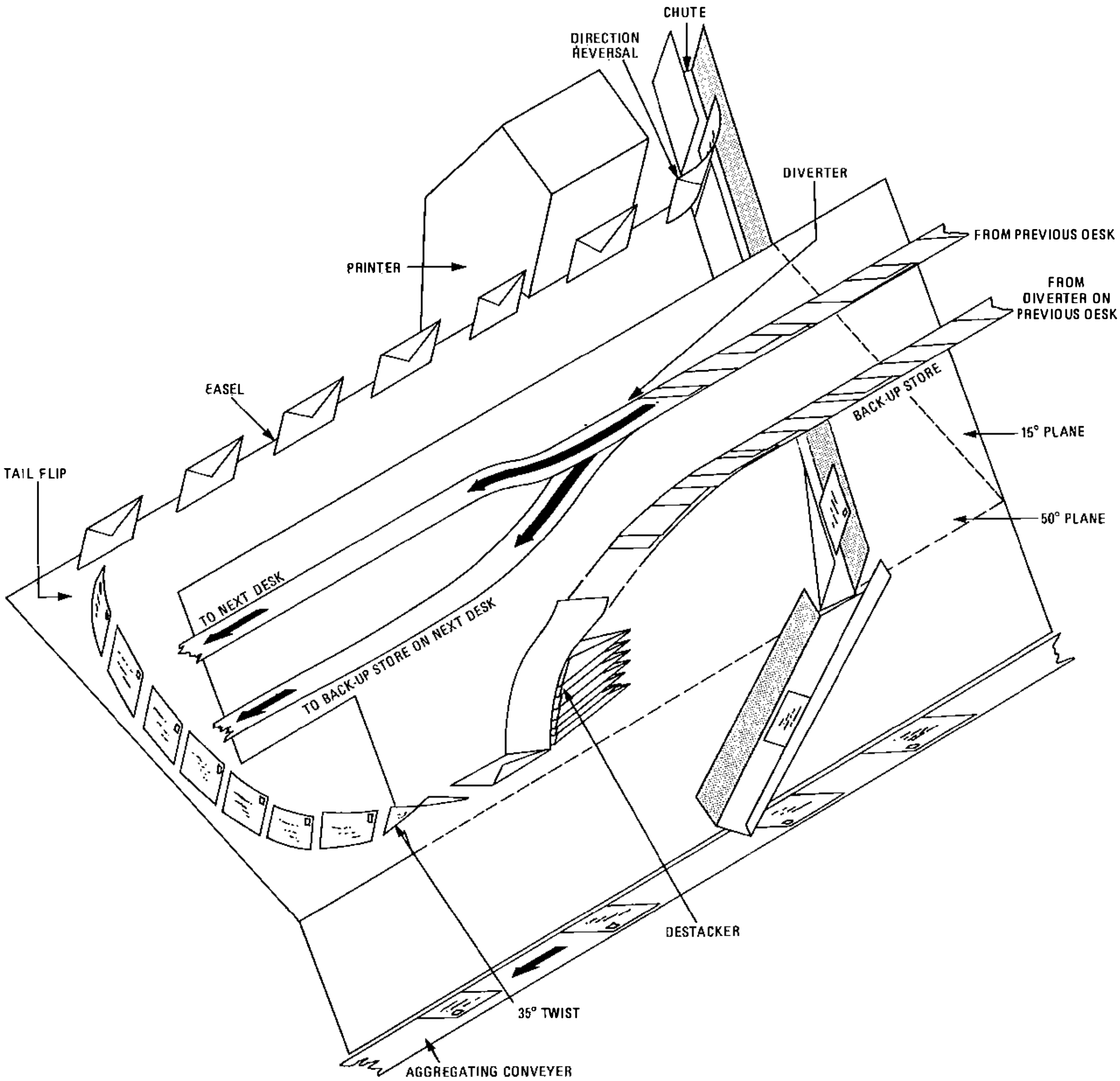


FIG. 4—Letter path of automatically-replenished coding desk

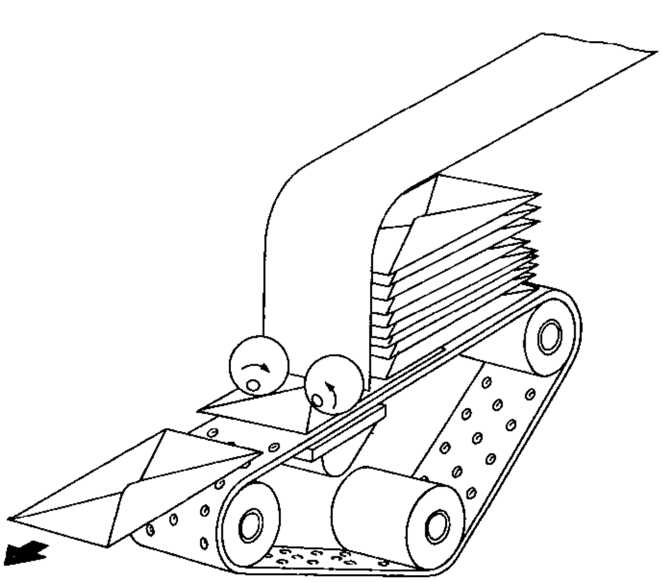


FIG. 5—An automatically-replenished destacker

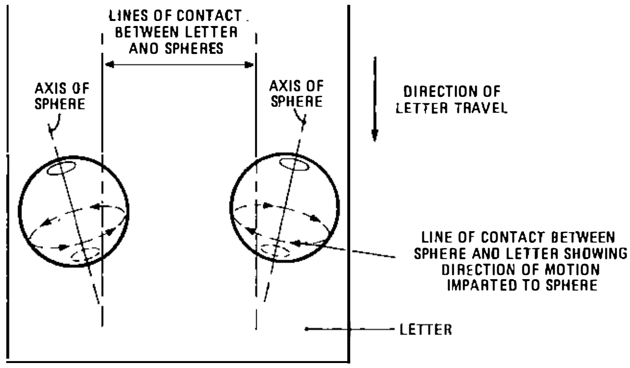


FIG. 6—Principle of operation of the destacker retarding spheres

Printer

After leaving the presentation zone, letters pass into the printing unit where the luminescent code dots are applied. The printer is of the hot-transfer type[6], similar to that used on earlier designs of coding desks, but having the print pins actuated by pneumatic cylinders. A letter has to be stationary

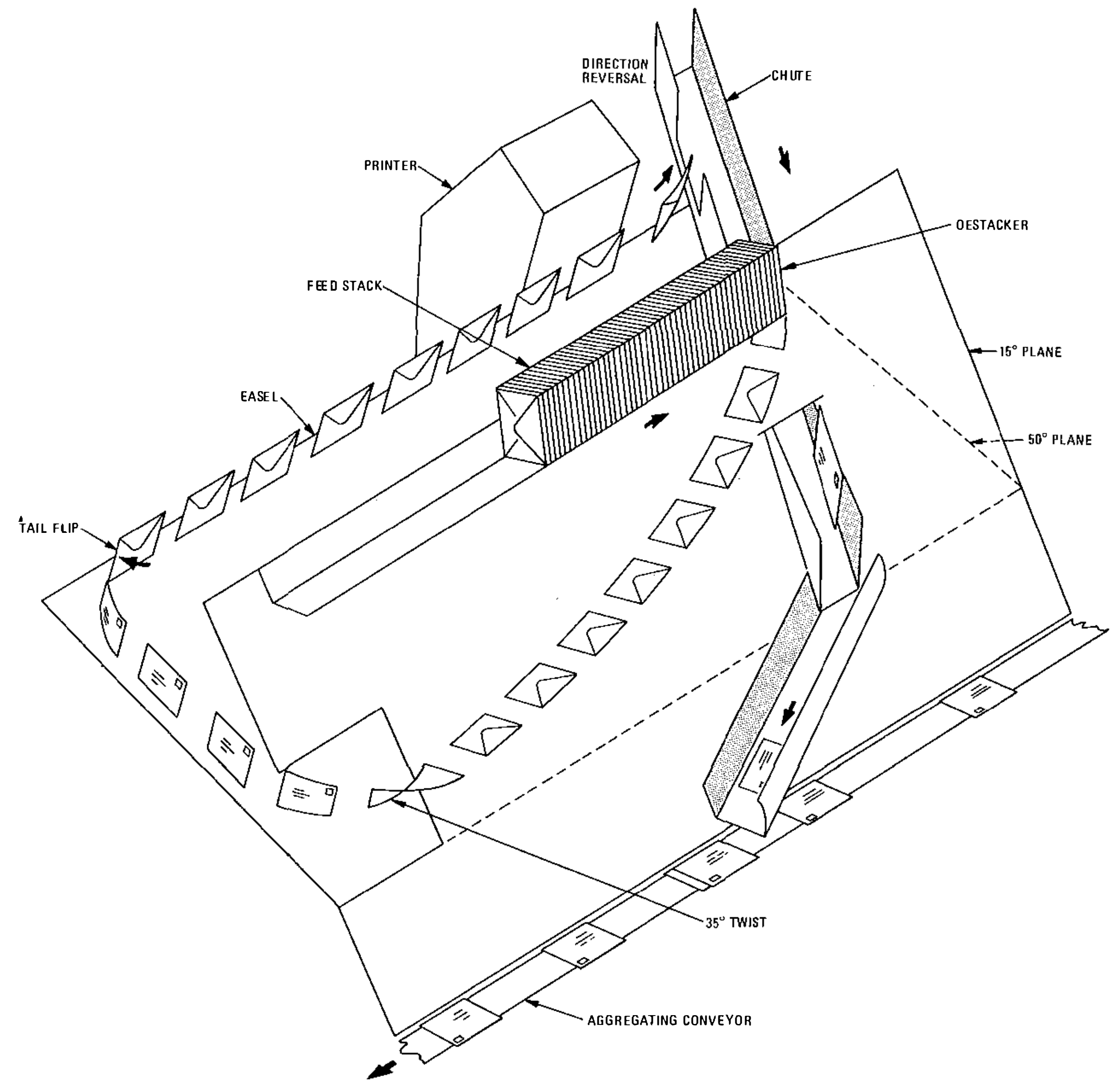


FIG. 7—Letter path of hand-loaded coding desk

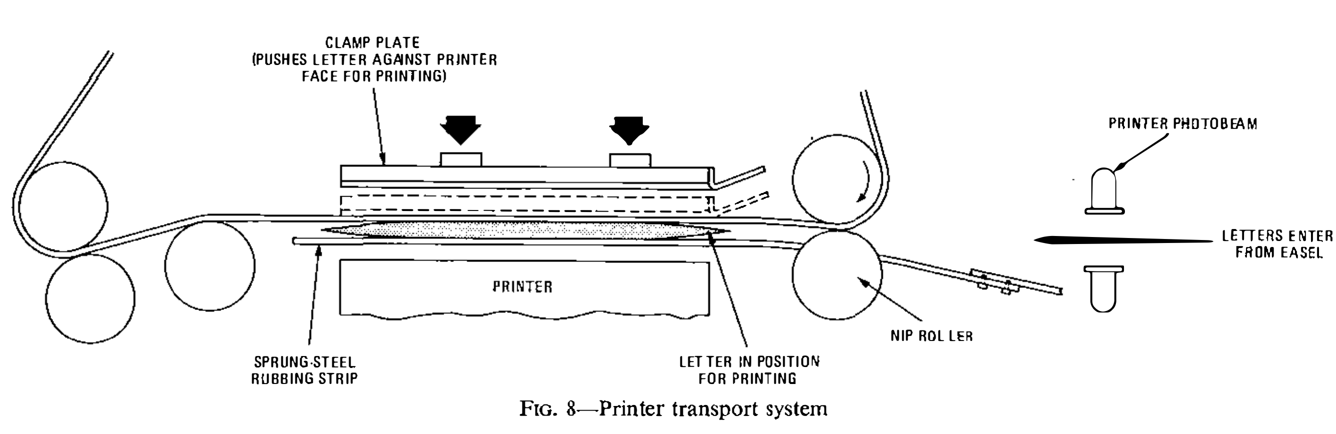


FIG. 8—Printer transport system

for printing to take place. From the end of the easel, letters are transported through the printer sandwiched between a single narrow-belt and a fixed strip of sprung steel lying between the 2 code-mark row positions (see Fig. 8). The fixed steel-strip is anchored at the entry end and an idling roller is provided to ensure take-up of the letter. The printer belt runs at the higher speed of 60 m/min so that a letter leaving the easel is snatched ahead of the moving stream to separate overlapping items, and to gain sufficient time for each letter to be stopped and printed before the next one catches up. All letters stop with their trailing (right-hand) edges at the same fixed point since it is from this edge that the code-marks are referenced. At the end of the print cycle, the printer belt restarts to eject the letter from the printer and to snatch the next one from the easel. On ejection, the letters are reversed into a gravity chute which guides them beneath the replenishment system onto an aggregating belt conveyor.

CONTROL SYSTEM

The electronic control for the earlier designs of coding desks, which had specific discrete letter locations for pre-viewing, keying and printing, was straightforward insofar as ensuring that the 28 bit code-mark information from the translator was assigned to the correct letter. In the easy-view desk, the ever-changing number of letters between the one being keyed and the printer, together with the need for an indicator to keep the operator informed as to the letter being keyed, requires electronic control which, at the time of the design of the first generation desk, would have been very bulky. It was the introduction of integrated circuits that made practicable the easy-view principle with its simple mechanical design.

To register each letter into the electronic control system, a photoelectric monitor beam is provided just prior to the presentation unit. This photo-beam facility operates in conjunction with a tail-flip device, which is illustrated in Fig. 9. The monitoring beam is detected between spaced or overlapping letters when the trailing edge of the leading letter flips straight as it changes direction round the curve. After passing through the tail-flip device, the letters emerge from between the sandwich belts onto the easel conveyor.

Two designs of engineering for the electronic control system were tested: the automatically-replenished coding desk (designated *Type C13*) was equipped with a microprocessor (the first to enter on-line operational service in the BPO); the hand-loaded version of the coding desk (designated *Type C17*) uses committed logic elements. The control methods of the 2 coding desks follow broadly similar philosophies, both systems being required to

(*a*) track all letters from the tail-flip photoelectric beam across the presentation easel to the printer,
(*b*) indicate to the operator the item to be coded,
(*c*) store the code-mark information received from the translator,
(*d*) recall the code-mark information at the printing section,
(*e*) control the speed of the main drive motor, and
(*f*) control other minor peripheral functions.

Microprocessor Control

In the microprocessor version of the desk, the desk control program is stored in a non-volatile read-only memory (ROM), and a random-access memory (RAM) is used for variable data storage such as positional tracking of letters, error detection and lamp displays. The microcomputer runs specific control programs for letter tracking, lamp displays and run-up–run-down control functions, stimulated by signals generated by the desk sensing devices. A block diagram of the microprocessor control-system is shown in Fig. 10. As each letter is sensed at the entry to the easel, it is allocated its own location in the memory, which is loaded with a number representing the distance of the trailing edge of the letter from the output point of the easel. As each 19 mm movement of the easel belt is detected, the count value in the letter storage location is decremented to leave a value representing the current letter position. Thus, as each 19 mm synchronizing pulse is detected, all letter-count values are decremented, thereby keeping track of all letter positions. The control system keeps track of the letter to be keyed by marking a storage area associated with its counter. The letter-position indication required for the operator is generated from the counter associated with the keying letter. Answers from the translator, received by the coding desk, are stored in separate locations, which are linked in software with the counter location. In this way, code-mark printer data is strictly associated with the appropriate letter in the keying sequence. The speed of the easel is calculated from the value of the counter associated with the keying letter. Speed commands from the microcomputer are sent in digital form to a digital-to-analogue converter in the motor control circuit. When a letter nears the output end of the easel, a specified series of

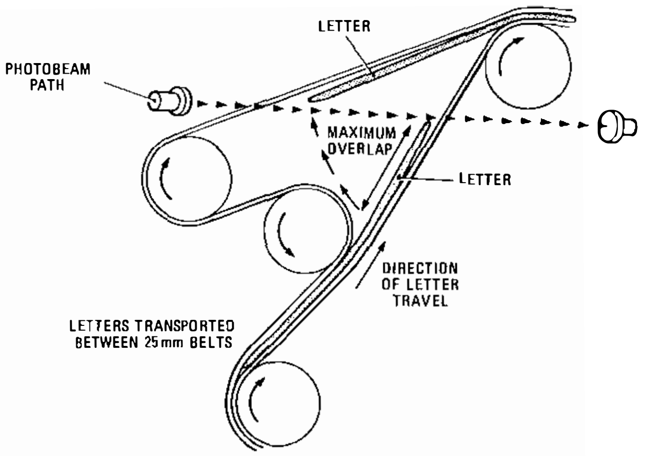


FIG. 9—Principle of operation of the photoelectric monitor

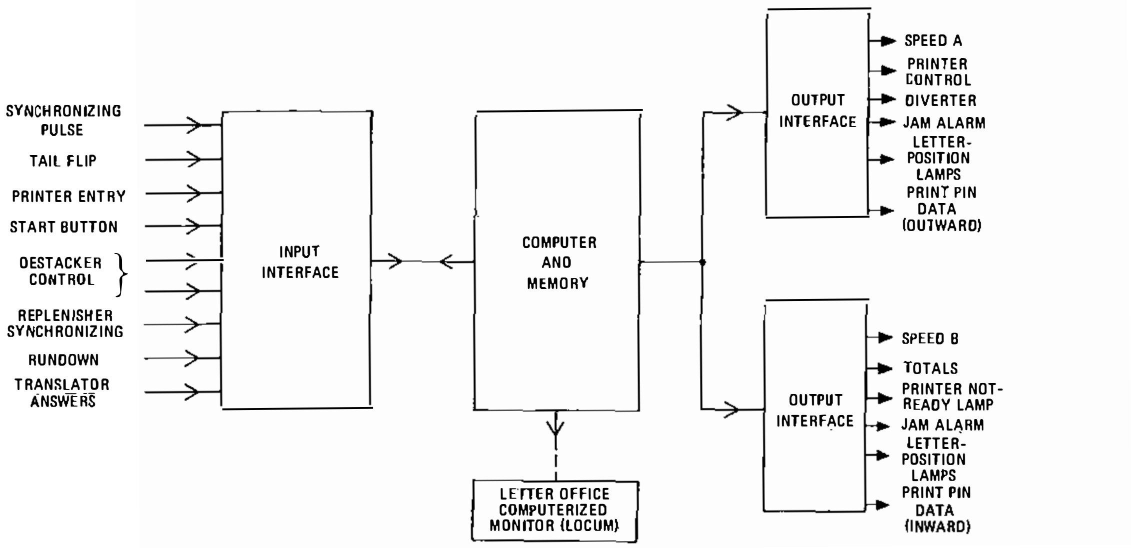


FIG. 10—Block diagram of the microprocessor control system

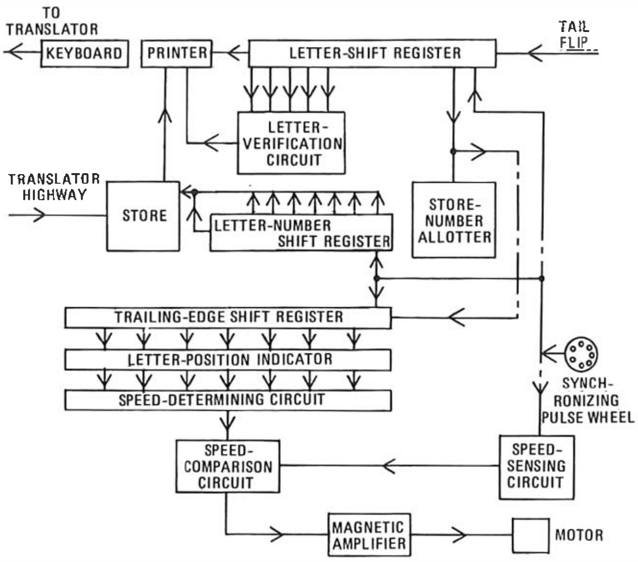


FIG. 11—Block diagram of the hard-wired logic control system

letter positional tests are applied, and printing is aborted if a chance of incorrectly code-marking a letter is possible.

Hard-Wired Control

In the hard-wired logic system, each letter is tracked across the easel by a number of shift registers, which are stepped by pulses obtained from a light-chopping disc driven by the letter transport belt. Of these registers, the electronic letter-shift register is of primary importance. Its data is received from the tail-flip beam and it carries electronic letters which are each represented by a series of logic *ones* separated by logic *zeros*. When a letter arrives at the easel, it is allocated a numbered-store location of its own, in which its code-mark pattern will be held between completion of keying and arrival of the item at the printer. The number of the store location is recorded on 3 parallel shift registers, which step in unison with the electronic letter. Fundamental to this hard-wired approach to desk control is the trailing-edge shift register, in which a trailing edge is represented by a single logic *one*. The position of the trailing edge of each letter is derived from the electronic-letter register. When the printing information is received from the translator, after the completion of keying, the relevant trailing edge is expunged. Thus, the trailing-edge register contains only those letters which have not yet been dealt with by the operator and makes the most advanced trailing edge the keying item. It is this shift register that controls the letter-position indicator lamps and the letter-transport speed. Code-mark information from the translator is put into the code-mark store bearing the number of the letter. When a letter arrives at the printer, its code-mark pattern is recovered from the code-mark store, as directed by the letter-number shift register. Immediately before printing, the position of the electronic letter is compared with that of the actual letter as detected by the printer entry beam. Printing is aborted if the letter has lost station, is excessively overlapped, or has been removed from the easel by the operator. A block diagram of the hard-wired logic control system is shown in Fig. 11.

Speed Control

The main drive-motor is a single-phase induction motor, fitted with a high-resistance rotor. The speed is controlled by a magnetic amplifier[7] connected in series with the power supply to the motor, the controlling signal to the magnetic amplifier being derived from the position of the keying letter on the viewing easel. Accurate speed control is achieved by means of a feedback signal, derived from the actual speed of the letters, which is obtained from the synchronizing pulse wheel driven by the letter conveyor belt.

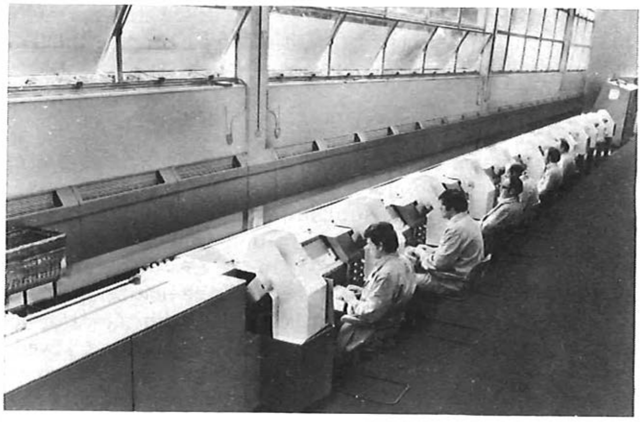


FIG. 12—Coding-desk suite at the Redhill mechanised letter-office

FAULT LOCATION AND REPAIR

An extra printed-wiring board can be inserted into the microprocessor chassis to enable the keyboard to be used to address the microprocessor for fault-location purposes, and the committed logic controller has a test panel by which much of the electronics can be tested periodically. These facilities are intended only to identify faulty printed-wiring boards so that they can be sent for repair to a centre at which computerized automatic test equipment is available.

CONCLUSION

The easy-view coding desk has substantially met the initial design aims. The key factor has been the method of presentation, which brings a two-fold benefit over other types of desk. Firstly, pre-reading is provided free as far as mechanical components are concerned and, secondly, the transport system from the destacker to the easel operates as a single simple-system without the need for mechanisms to stop, start, or change the axis of movement of the letters.

A suite of easy-view coding desks (see Fig. 12) has been in operational service at the Redhill mechanized letter-office since October 1975, and over 300 000 letters a day are being processed by the coding-desk operators. The operation of the automatic and the manually-loaded replenishment systems has been evaluated, and the latter type has been adopted as a preferred means of letter replenishment.

The success of the operation at the Redhill mechanized letter-office has led the BPO to adopt the easy-view letter-coding desk as a standard for new mechanized letter-offices.

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The authors wish to thank the staffs of the BPO Postal Headquarters and the South East Postal Region for their assistance in this development.

References

1 PHILLIPS, K. H. C., and STURGEON, A. W. A New Code Translator for the Letter-Sorting System. *POEEJ*, Vol. 66, p. 2, Apr. 1973.

2 HEWETT, J. W., and MUCKETT, R. G. Code Marking of Letter Mail. I.Mech.E. Postal Engineering Conference, May 1970, Paper 38.

3 COPPING, G. P., and BEAK, K. L. A Single-Operator Letter-Sorting Machine. Part 1—Introduction and the Experimental Machine. *POEEJ*, Vol. 51, p. 104, July 1958.

4 COPPING, G. P., WHITTINGTON, K. W. H., and MUCKETT, R. G. Prototype Coding Desks for Letter Mail. *POEEJ*, Vol. 64, p. 98, July 1971.

5 Patent Application Number 37612 (12-9-75) Destacker Spheres.

6 FORSTER, C. F. Use of Phosphorescent Code Marks in Automatic Letter-Facing and Sorting Machines. *POEEJ*, Vol. 54, p. 180, Oct. 1961.

7 MILNES, A. G., GALE, H. M., and ATKINSON, P. D. Magnetic Amplifiers. *JIEE*, 1949, Vol. 96, pt. 1.

Operators' Telephone Instruments used in the UK: Past, Present and Future

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UDC 621.395.6

It has been nearly 100 years since the first telephone exchange was opened in the UK; this article reviews the development of operators' telephone instruments during this period and considers possible future designs.

INTRODUCTION

The first public telephone exchange in the UK was opened in August 1879, at 36 Coleman Street, London. It was soon apparent that, to work efficiently, an operator needed both hands free. The type of microphone and earphone available at that time could not readily be adapted for support on an operator's head or body.

An early example of an operator's headset is shown in Fig. 1. This instrument was made in the USA in about 1880 and used the *Gilliland harness*, supported on the shoulders by a yoke. The total weight of the headset was approximately 3 kg. It was not, however, adopted for use in the UK.

From 1895 to 1960, the practice was for an operator to use an earphone mounted on a headband and a microphone coupled to a horn mounted on a breastplate; such instruments were known as *head-and-breast sets*. Eventually, in 1960, the British Post Office (BPO) introduced a combined headset, known as the *Operator's Headset No. 1*,[1] and this headset instrument is still in current use.

† Research Department, Telecommunications Headquarters.

FIG. 1—The Gilliland harness, c 1880

((c) National Geographic Society)

REVIEW OF OPERATORS' TELEPHONE SETS

The operators at the first UK telephone exchange used a telephone set consisting of a Blake microphone and a Bell earphone; an example of this type of set can be seen in Fig. 2. The microphone was mounted on a brass pole and the earphone was held in the operator's hand. (The early designs of microphones and earphones were named after their inventors).

A modified form of the Blake microphone, fitted in a round nickel-plated metal case, was designed for use with telephone switchboards. Being much smaller than the original version, it did not conceal such a large part of the switchboard from

FIG. 2—The Blake microphone and the Bell earphone

(In use at the Croydon telephone exchange in 1884)

FIG. 3—The Ericsson head-and-breast set, c 1908

FIG. 4—The Peel-Connors head-and-breast set, c 1905

FIG. 5—The Transmitter No. 23A and Receiver Headgear No. 10A

FIG. 6—The Operator's Headset No. 1

the operator's view. This microphone was suspended by 2 flexible conductors in front of the operator, and the height could be adjusted by a pulley system.

The first operators' set that allowed both hands to be free for making connexions is shown in Fig. 3. In this design, made by Ericssons in about 1895, the microphone was mounted on a breastplate and a small and lightweight earphone (termed the *watch* type and described later in this article) was mounted on a headband. The microphone was pivoted on its axis with the diaphragm vertical; thus, the distance of the horn from the mouth was variable. The breastplate was made of aluminium and was suspended from the operator's neck by an adjustable band. A switch was incorporated on the case and designed such that, when the horn was revolved to a non-speaking position, the microphone current was disconnected.

In the mid-1890s, handsets were also used on the smaller telephone exchanges (as they are today on PBXs). These were often Deckert or White microphones, and a watch-type earphone was mounted upon a suitable handle on which a button was provided to enable an operator to monitor (listen only) a conversation, thus preventing noise from the switch room or the operators' breathing being overheard by the customers engaged in conversation. These sets were all used on local-battery (LB) working but, with common-battery (CB) working becoming more prevalent, these sets were found to be unsuitable for general use.

The Peel-Conners design (see Fig. 4), introduced in 1905, fulfilled all the operational requirements of the day, and used the White (commonly termed the *solid-back*) microphone. To this microphone was fitted a long flared-horn that incorporated a universal ball-joint that enabled the position of the horn to be adjusted, or positioned away from an operator's mouth if required. Again, the watch-type earphone was used and was mounted on a headband. This particular type of head-and-breast set was in use for nearly 40 years.

The next generation of head-and-breast set incorporated features of 2 earlier types of set, namely a flared-horn and a horn-adjustment system similar to the Ericsson instrument. This set used an inset-type microphone (the BPO Transmitter Inset No. 13), which was cheaper, and afforded easier maintenance because the microphone could be replaced easily; the whole assembly was known as the *Transmitter 23A*. The earphone used was a new type of acoustically-equalized earphone, the Receiver Headgear No. 10A; this was similar to the Receiver Inset No. 2P (used in the BPO Telephone No. 332), but it had a bakelite frame instead of the diecast aluminium frame to minimize its weight. The complete head-and-breast set can be seen in Fig. 5.

The first combined headset used by the BPO was the Operator's Headset No. 1 (shown in Fig. 6); this headset was first brought into service in 1960. The development of small, light and highly-sensitive transducers, namely the Transmitter Inset No. 15 and the Receiver No. 3T, led to their being incorporated in the same housing; for example, in the Telephone No. 722 (the *Trimphone*). The horn is positioned by means of a springy joint at the elbow that joins the horn to the body of the set. The Headset No. 1 weighs about 140 g, being one-third of the weight of the head-and-breast sets (450 g).

DEVELOPMENTS IN TRANSDUCERS [2,3]

Microphones

The Blake microphone was patented in the UK on 20 January 1879. This microphone was developed by Francis Blake of the Bell organization in the USA, and was a carbon microphone that operated on the principle of detecting the variation in contact pressure between a platinum pellet and a disc of hard carbon; this proved to be a very reliable microphone, although the static pressure had to be adjusted fairly carefully.

As telephone lines became longer there was a need for more-sensitive transducers. These transducers had to be small and light in weight to be suitable for use with head-and-breast sets.

The success of the Ericsson design of microphone was achieved by ensuring that the front electrode (formed by the diaphragm) operated over the whole surface of the granules and the back electrode (fixed) had a raised surface, thereby increasing the surface area in contact with the granules and preventing them from forming a compact mass.

From 1905, the solid-back microphone displaced nearly all the other types for CB working. This was the first type of microphone to use more fully immersed electrodes. The separation of the diaphragm from the granule chamber enabled each to be designed independently, using the most suitable materials available at the time. Although carbon microphones were unsurpassed for sensitivity, they remained highly susceptible to changes in orientation, to excessive current and the way they were handled; impairments to performance due to the granules becoming too tightly packed, or because they were too loose, were not uncommon. Therefore, to achieve optimum performance, the solid-back microphone needed to be operated in an upright position.

The performance of the Transmitter Inset No. 13 was a great improvement on any previous type of microphone. The electrodes of the microphone were totally immersed in the granules (the immersed-electrode principle had been established as the most satisfactory arrangement for carbon-microphone design). The granules were packed around the electrodes in such a way that, in whatever position the microphone was held, the required pressure of granules on the surface of the electrodes was maintained. Thus, whatever the position of the microphone, its resistance was kept within reasonable limits, and the conversion of variations in sound pressure on the diaphragm to variations of pressure on the granules remained substantially constant. In some of the earlier types of microphone, if the contact surface between the granules and the electrodes was reduced (even if the granules did not actually leave the electrode), the result was that transmission suffered badly and noise was introduced.

The microphone in present use, the Transmitter Inset No. 15, was developed in 1957 and uses concentric hemispherical-shaped electrodes; in this unit, the almost complete filling of the granule chamber ensures greater stability and freedom from severe positioning effects.

Earphones

The original operators' set used the single-pole Bell earphone, but this was superseded in 1890 by the double-pole (bipolar) Bell earphone. The principle of operation of the bipolar earphone was that a steel diaphragm was held at a small distance from 2 pole-pieces, which were subjected to a magnetic flux derived from the poles of a permanent magnet. Coils of wire, wound on each pole-piece, carried the electric currents that were to be transformed into sound.

However, the earphone used for all subsequent operators' sets up to 1940 was an adaption of an earphone produced by Clement Ader, in France, in 1879. This earphone, called the *watch* type, used a flat circular-magnet terminating in 2 soft-iron pole-pieces around each of which was wound a coil of wire; this design proved to be the prototype of all the compact lightweight-headgear earphones in use until 1940. There was little change in the basic construction of the earphone during this period; those changes that were made related to improvements in the quality of the materials used, especially in the magnetic circuit.

The first dramatic change to earphone design occurred in 1940 when the Receiver Headgear No. 10A was introduced. The sensitivity/frequency response of this design of earphone was much flatter than that of previous designs, this improvement being achieved by acoustic equalization.

Further improvements in frequency response and in

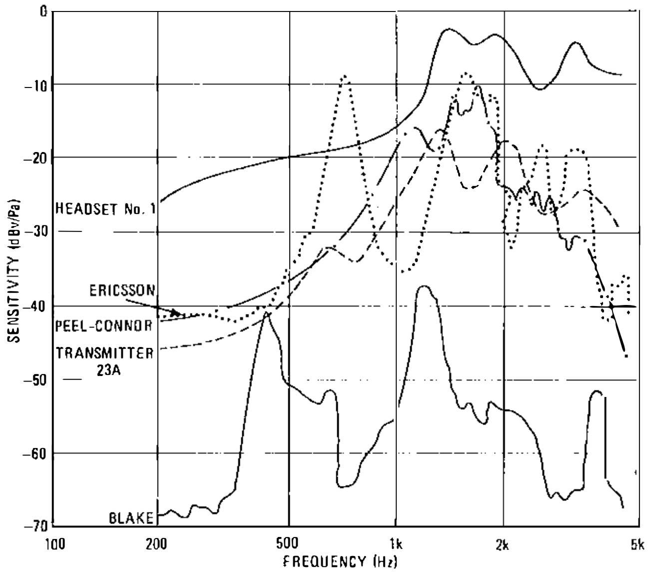

FIG. 7—Microphone open-circuit sensitivity/frequency characteristics

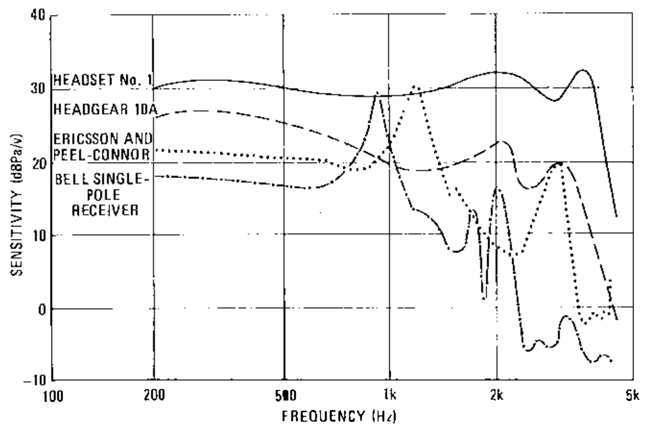

FIG. 8—Earphone constant-voltage sensitivity/frequency characteristics

sensitivity were achieved in 1956 by the rocking-armature design of the capsule earphone, which is now used in almost all BPO telephone sets.

IMPROVEMENTS IN TRANSMISSION PERFORMANCE

The improvement in sensitivity and frequency response between various types of microphone can be seen in Fig. 7. The measurements recorded in Fig. 7 were made by placing each microphone, coupled to its associated horn, 25 mm from the lip plane of an artificial voice; each microphone was fed with a current of 50 mA (except for the Blake microphone which, because of its low resistance, was fed with 90 mA). The *free-field sound pressure* (FFSP) was also measured at 25 mm; this point of measurement is known as the *mouth reference point* (MRP). The expression for microphone sensitivity is

$$\frac{\text{microphone open-circuit voltage}}{\text{FFSP at MRP}} \text{ dBV/Pa.}$$

In the early years, improvements in microphone performance were quite substantial; large increases in sensitivity were achieved, especially with the introduction of the carbon-granule type microphone. Until 1960, further improvements were confined to improving the shape of the frequency response. Finally, in 1960, the design of the Headset No. 1 afforded a further increase in sensitivity.

Improvement to earphone performance followed a similar pattern to that of microphones; Fig. 8 shows the improvement in performance that has been achieved between the various types. The earphones were all measured on an artificial ear. The expression for earphone sensitivity is

$$\frac{\text{sound pressure developed in an artificial ear}}{\text{input voltage}} \text{ dBPa/V.}$$

In general, in the early years, transducers were rated on volume (loudness) comparison methods. That is to say, each new transducer was rated against a known standard, and attenuation was inserted in the test circuit until equal loudness was obtained. The difference in loudness between the standard test circuit and the transducer under test was expressed in terms of the loss. As a result of this method of comparison, defects of the standard components were perpetuated in new or modified designs of transducers. At one time, the standard earphone was the double-pole Bell earphone, which had a pronounced mechanical resonance and a correspondingly high output in the neighbourhood of 1 kHz. New earphones that were assessed against this standard earphone compared badly on loudness tests, unless they too had a similar mechanical resonance. The fallacy of this method of testing was realized and, in the 1930s, the performance of transducers was determined purely by articulation tests, which are a measure of the information capacity of a path when it is transmitting information in the form of speech signals. However, this method was not accepted internationally and, with the improvement in frequency response of the transducers, volume tests were continued as the means of assessment. The internationally-accepted reference system in use at the present time is the NOSFER[4] system (*noveau système fondamental des equivalents de référence*—new basic system of reference equivalents), and volume efficiencies are subjectively measured in terms of reference equivalent decibels.

The NOSFER method of evaluation causes some difficulty in quantifying the efficiency of telephone instruments because the reference system is a high quality, wideband system (approximately 0–8 kHz) and is balanced against systems that are band restricted (300–3400 Hz). As a result, the BPO is using an intermediate reference system (IRS), which has frequency responses similar to those of practical telephone sets and the results are termed *loudness ratings* (LRs).

Sensitivity and frequency characteristics do not indicate precise performance characteristics because the responses are determined in a rather artificial way. The designer of operators' telephone instruments has to take into consideration the natural position of the horn in front of an operator's mouth. The position of the horn is determined by the contours of both body and face for the head-and-breast sets, and by the face only for the Headset No. 1. An estimate of the performance of the instruments mentioned in this article has been made and is recorded in Table 1. The performance assessments were made using a typical present-day exchange operators' transmission bridge circuit, as shown in Fig. 9. Of course, apart from the Headset No. 1, any movement in the position of the head of a talker alters the output voltage of the microphone.

HEADSETS OF THE FUTURE

Recent developments in both the transducer and micro-electronics fields have made it possible to design headsets that have a significant reduction in weight and size compared with present-day designs. The physical appearance of the present type is an external earphone and a horn-loaded microphone, whereas the new types might be an insert-type

TABLE 1

Comparison of Operators' Telephone Instrument Performance

Date	Type of Operators' Instrument	Microphone Resistance (Ω)	Send Loudness Rating (Estimated) (dB)	Earphone Impedance at 1 kHz (Ω)	Receive Loudness Rating (Estimated) (dB)
1880–1900	Blake and Bell	15	+50·0	250	+4·0
1895–1915	Ericsson	68	+19·0	700	0
1905–1940	Peel-Connors	82	+19·0	250	0
1940–1960	Transmitter 23A, Headgear 10A	50	+19·0	220	−4·0
1960–	Headset No. 1	70	+11·0	150	−11·0

Notes: In the sign convention for loudness rating, a positive sign indicates that a connexion is quieter than the reference standard; a negative sign indicates that the connexion is louder.
By comparison, a BPO Telephone No. 706 connected to a transmission bridge has a send loudness rating of +4 dB and a receive loudness rating of −6 dB.

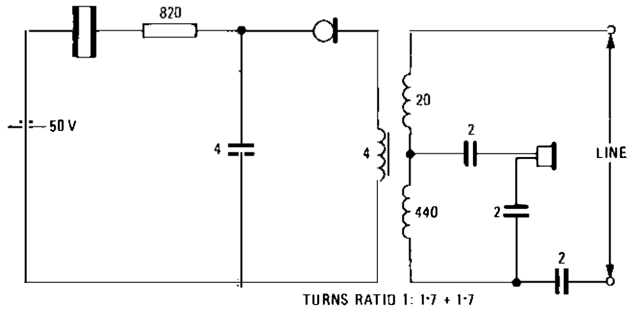

FIG. 9—Operators' transmission-bridge circuit

earphone and a voice tube. In some designs, the microphone can be carried on a small boom near to the lips. The earphone could be a true insert-type which enters the ear canal, thus providing an acoustic seal: alternatively, a slightly bulbous tip could be designed to rest on the entrance of the ear canal; this design would not necessarily ensure a good acoustic seal. As a result of the reduction in weight, modern-type transducers can be contained in a small assembly and carried on a head-band or spectacle frame, or even mounted such that they fit behind the pinna of the ear and require no other form of support. Any combination of microphones and earphones and assemblies could be achieved.

However, the sensitivity of the new microphones is substantially lower than that of the Transmitter Inset No. 15, and amplification would be required to maintain performance. Apart from the basic amplification function, amplifiers can be designed to reduce the microphone sensitivity at low sound pressures. (These amplifiers are called *switch-gain* amplifiers, and their purpose would be to reduce the pick-up of environmental room-noise.) An advantage of the carbon-granule type of microphone is that it is unaffected by the polarity of the line; these new types would require a diode bridge to maintain the correct polarity to the amplifiers.

With the continuing reduction in the cost of microelectronics, it is possible that such new headsets could become an attractive proposition in the future.

CONCLUSIONS

This article has followed the development of the operators' telephone instrument from its infancy in 1879. Because of their size and weight, early instruments could not readily be supported on the head. As microphones became smaller, lighter and more sensitive, they could be more easily mounted on the body in front of the mouth, thus leaving both hands free.

Finally, even greater reduction in weight has led to both transducers being mounted in the same housing on the head, thus creating the headset as we know it today.

ACKNOWLEDGEMENTS

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References

[1] SPENCER, H. J. C., and ROBERTSON, J. S. P. A Light-weight Headset for Telephone Operators. *POEEJ*, Vol. 53, p. 177, Oct. 1960.

[2] HERBERT and PROCTOR. Telephony, Vol. 1, Chapter VIII, (Sir Isaac Pitman and Sons, Ltd., 1934).

[3] ATKINSON, J. Telephony, Vol. 1, Chapter III, (Sir Isaac Pitman and Sons, Ltd., 1948).

[4] FRY, R. A., and BANNER, I. H. Transmission Performance Assessment of Telephone Networks. *POEEJ*, Vol. 65, p. 145, Oct. 1972.

30-Channel Pulse-Code Modulation System

Part 2—2·048 Mbit/s Digital Line System

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UDC 621.376.56

Part 1 of this article explained why the British Post Office has adopted a 30-channel pulse-code modulation standard, and included a technical description of the multiplex equipment. This part describes the operating principles, equipment, and installation of the associated 2·048 Mbit/s digital line system.

INTRODUCTION

The 30-channel pulse-code modulation (PCM) system adopted by the British Post Office (BPO) provides the basic multiplexing unit upon which the UK future digital-transmission network hierarchy[1] will be based. In the proposed integrated digital transmission and switching network, at the digital-switching exchanges (known as *System X*) the basic switching function will be performed between 30-channel PCM system output and input ports. Thus, switching centres in the BPO junction network[2] will be interconnected by 2·048 Mbit/s digital line systems. Therefore, in the future, these systems will be provided on the grounds of necessity, as distinct from the policy adopted hitherto whereby 1·536 Mbit/s digital line systems were used as a means of augmenting the capacity of existing junction cables.

Over the past decade, the BPO has installed some 7000 24-channel PCM[3] systems and their associated 1·536 Mbit/s digital line systems; these systems have many years of service life remaining. With such a large capital investment in 1·536 Mbit/s PCM systems, the BPO wished to minimize the cost of the changeover to the new line systems; therefore, the following terms of reference for the design of 2·048 Mbit/s line systems were adopted:

(*a*) the use of cables and repeater housings were to be shared with those existing for the 1·536 Mbit/s systems if required,

(*b*) the spacing of regenerators was to be similar for both systems,

(*c*) a common fault-location system was to be used, and

(*d*) the power-feeding arrangements were to be identical.

The most significant differences between the new regenerators and those used for 1·536 Mbit/s line systems are that automatic equalization is provided over a range of more than 30 dB, and the size of regenerators has been reduced to allow an increase of 50% in the number that can be accommodated in a repeater case. Additionally, the specification of the 2·048 Mbit/s regenerators ensures that full compatibility will be achieved between units from all UK manufacturers, a factor which provides complete freedom to interconnect non-homogeneous systems and the rationalization of spare plant for maintenance replacement. Because 2·048 Mbit/s line systems will form an integral part of the future digital-transmission network, the regenerators have been specified to minimize the cumulative jitter to ensure compatibility with higher-order multiplexing equipment; a significant parameter in this context is the choice of line code.

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LINE CODE

To transmit binary information over a digital line system the information must be encoded into a sequence of symbols known as a *line code*. There are a number of fundamental properties which a line code must possess for transmission over tandem-connected regenerators:

(*a*) The line code must have an energy spectrum with very small low-frequency components and no zero-frequency component. This feature enables dependent regenerators (that is, those remote from terminal stations) to be energized from a relatively simple constant-current source. The absence of significant low-frequency energy has the advantage that miniature coupling transformers, which reduce low-frequency noise and facilitate rejection of common-mode signals, can he used at regenerator input and output ports.

(*b*) The line code must contain adequate timing information. The regenerative process requires re-timing of the transmitted signal, the timing information being derived from transitions in the received line signal. To minimize the generation of jitter, the timing content is maintained at a high level and this is achieved by imposing a constraint on the number of consecutive zeros transmitted.

(*c*) The structure of the line code must be such that monitoring of the error rate of the transmitted information is possible.

The above properties can be achieved by translating the unipolar binary information into a sequence of bipolar symbols. An example of this form of encoding is known as *alternate-mark inversion* (AMI), which is used on 1·536 Mbit/s line systems. The main disadvantage of AMI is that zeros in the binary input are not encoded and hence adequate timing information cannot be derived when long sequences of

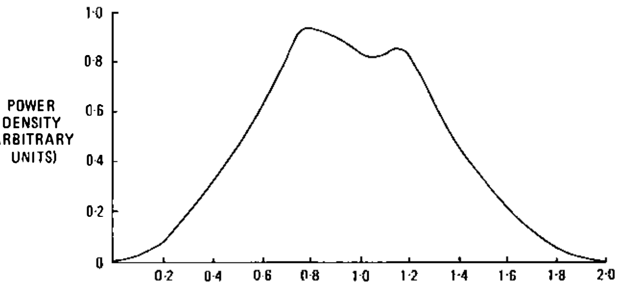


FIG. 11—Power spectrum for HDB3-encoded random binary signal

consecutive zeros occur. The AMI code can easily be modified to overcome this problem by inserting a mark (pulse) when a predetermined number of consecutive zeros occurs; this form of encoding is known as *high density bipolar n* (HDB*n*). (In the HDB*n* code, *n* is the maximum number of consecutive zeros transmitted by the code converter.) As *n* decreases, the low frequency content of the energy spectrum increases and, for 2·048 Mbit/s line systems, a compromise value of $n = 3$ has been chosen. Thus, the line code adopted is HDB3 and details of the encoding method have been described in Part 1[4] of this article. The power spectrum for an HDB3-encoded random binary signal is shown in Fig. 11.

INTERFACE CONDITIONS

To ensure compatibility between items of interconnected 2·048 Mbit/s transmission equipment, it is essential that certain interface conditions are defined.

The digital signal presented at the 75 Ω coaxial output port of the line system complies with CCITT† recommendations[5]: the binary digit rate is 2·048 Mbit/s (± 50 ppm*); the signal is HDB3 encoded; pulse shapes are nominally rectangular with a peak voltage of 2·37 V and a 50% duty cycle to minimize inter-symbol interference; a zero is transmitted as zero voltage.

The signal presented at an input port will have a different pulse shape due to the effect of the coaxial cable interconnecting with preceding transmission equipment; the length of such cabling is limited to ensure that the loss at 1·024 MHz does not exceed 6 dB.

LINE TERMINAL APPARATUS

Terminal apparatus is engineered in BPO 62-type equipment practice and is normally located in telephone exchange equipment areas. Line terminal equipment for up to 48 digital line sections (DLSs) can be accommodated in one apparatus rack. A DLS comprises appropriate elements of 2 consecutive line terminal equipments and the interconnecting transmission path which, together, provide the means of transmitting and receiving a bothway digital signal between 2 terminal stations.

† CCITT—International Telegraph and Telephone Consultative Committee

* ppm—parts per million

Line Terminating Equipment

An equipment shelf (see Fig. 12—lower shelf) accommodates 4 DLSs and contains 4 terminal regenerator units, 4 line power-feed units and a centrally-mounted patching (cross-connexion) panel. The 75 Ω input and output ports of each DLS appear on the patching panel; U-links are provided for maintenance access and also provide a convenient flexibility point to enable a faulty line or multiplex to be taken out of service. A buffered monitor-point is provided on each input and output port to enable in-service performance checks to be carried out.

A terminal regenerator unit (see Fig. 13) carries both directions of transmission of a DLS. A signal incoming from line is regenerated using circuitry identical to that contained in a dependent regenerator. In the transmit direction, the signal from the preceding multiplex equipment is transmitted to line via a transformer, which matches the impedance of the interconnecting cable (75 Ω unbalanced) to the nominal impedance of the line (120 Ω balanced). Connexions to the main distribution frame and thence to an external cable are made using balanced-pair cables, which are double screened to minimize the effect of crosstalk from exchange-switching transients.

Ancillary Equipment

Two shelves of ancillary equipment (see Fig. 12—centre and top shelves) are included in a fully-equipped line terminal rack. One shelf contains power distribution equipment and includes a fuse unit and associated alarm equipment to monitor the power outlets to each of the remaining shelves. A second shelf contains units that provide access to the line system fault-location pairs and engineering speaker circuits; also included on this shelf is a digital pattern generator (see Fig. 14), which provides 8 outputs to drive spare DLSs.

Spare DLSs are always powered to prevent oxidization of metallic contacts and, in the absence of a digital drive signal, the crosstalk noise from working systems in the cable may cause regenerators to free run. In a free-running mode, regenerators

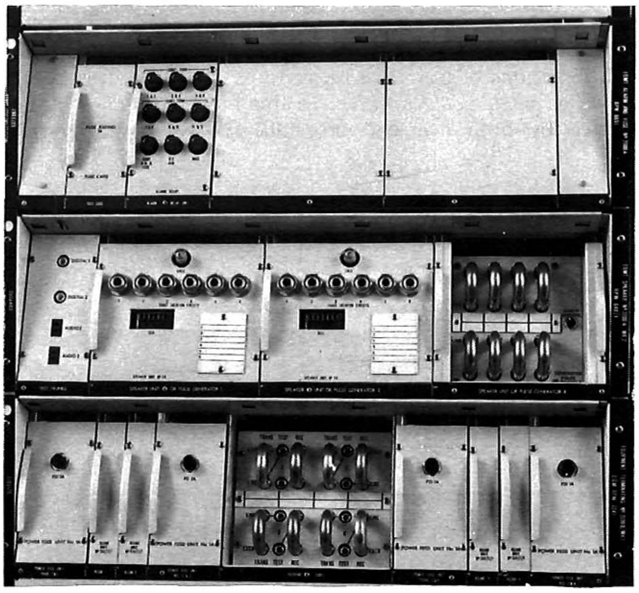


FIG. 12—Line terminal equipment

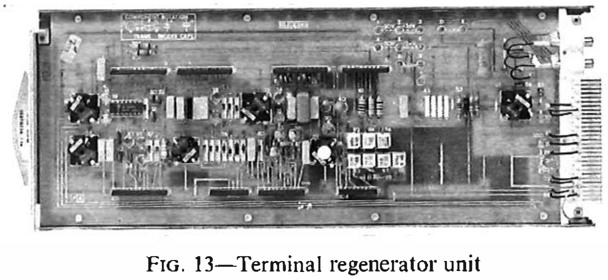
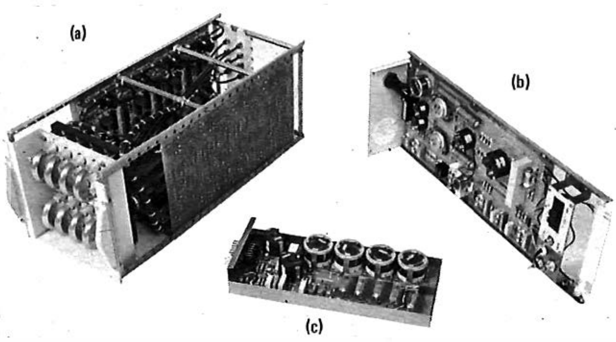


FIG. 13—Terminal regenerator unit



(*a*) Digital pattern generator
(*b*) Line power-feed unit
(*c*) Fault-location filter

FIG. 14—Ancillary equipment

can produce spurious outputs with large-amplitude high-frequency components, which may interfere with adjacent traffic-carrying systems. Therefore, to minimize crosstalk interference from spare DLSs, a pseudo-random pattern generator is used and each of the 8 output signals is evenly distributed in sequence phase.

POWER FEEDING

Terminal regenerator units are powered from the station battery supply. Each line-coupling transformer within a regenerator unit is centre-tapped on the line side to provide access to the phantom circuit of the 2 digital-bearer cable pairs. A line power-feed unit (shown included in Fig. 14) is connected across the phantom circuit and supplies, with an applied voltage of up to 75–0–75 V, a constant current within the range 48 ± 2 mA to the dependent regenerators. To comply with safety requirements, the maximum current is restricted to 50 mA DC, and the applied voltage must not exceed 150 V between pairs or 75 V between any conductor and earth.

On 0·63 mm cable pairs, up to 7 dependent regenerator units, each containing 2 regenerators, can be power fed from one terminal station. On longer DLSs, the power is fed from each terminal station; a block diagram of the arrangement is shown in Fig. 15.

TRANSMISSION MEDIA

The 2·048 Mbit/s DLSs operate over existing audio cables[6] in the junction and main networks, circuit lengths are normally 10–40 km. There are many different types and gauges of cable in the BPO junction network but, since the late 1960s, the standard cable provided for audio transmission has been paper-core quad trunk (PCQT) with 0·63 mm diameter copper-conductors. This type of cable accounts for over 50% of the junction cables now in service, and is supplied in 13 sizes, ranging in capacity from 14–1040 pair. Consequently, 2·048 Mbit/s regenerator designs are optimized for operation over 0·63 mm PCQT cable, although they will operate satisfactorily over different cable types, such as unit-twin type cable and on cables with larger conductor gauges.

SITING OF REGENERATORS

Regenerators are normally sited at loading-coil points, which corresponds to a nominal spacing of 1·83 km. The choice of regenerator spacing was influenced by the following factors:

(*a*) The spacing is compatible with the regenerator spacing used for 1·536 Mbit/s systems so that existing regenerator housings can be shared.

(*b*) Audio cable pairs must be de-loaded to provide the required transmission bandwidth; therefore, on new routes, it is convenient to co-site the regenerators with the loading coils to minimize the amount of jointing work required.

(*c*) Using 0·63 mm cable pairs, at the maximum regenerator spacing the insertion loss of the cable pair is approximately 36 dB at 1 MHz, which is the significant frequency of the signal spectrum. This magnitude of insertion loss enables a cost-effective regenerator design to be realized and, in a crosstalk dominated environment, it enables an economic cable-utilization factor to be achieved.

The distance between a terminal station and a first dependent regenerator is limited to a maximum of 1·1 km in the transmission direction towards the terminal. This restriction increases the level of the received signal and so reduces the effects of impulsive noise from exchange switching equipment which may be propagated over audio cable pairs and result in crosstalk interference into adjacent digital-bearer pairs. A similar limit also applies to regenerators on either side of a spur cable connected to exchange equipment.

CABLE UTILIZATION

The number of systems that can operate satisfactorily over symmetric pair cable is limited by mutual crosstalk interference. The cables are manufactured and installed to specifications that are concerned solely with audio transmission performance and they are used for digital transmission on an 'as found' basis. The crosstalk performance at high frequencies is influenced by the cable construction and installation techniques, and it is useful to summarize these to appreciate their significance in relation to digital transmission at 2·048 Mbit/s.

PCQT cables have a central core of 4 quads with one or more concentric layers built around it. Crosstalk is reduced by giving different twist lengths to adjacent quads in a layer, and by giving a different direction of lay to adjacent layers. For use at audio frequencies, the cable is divided into balancing groups of approximately equal size, each comprising one or more layers. Crosstalk at audio frequencies occurs due to capacitive coupling, and capacitive unbalance is minimized during installation by jointing quads in a predetermined sequence. Quad integrity is maintained throughout a cable and only quads in the same balancing group are allowed to be connected together. As a consequence, it is not possible to guarantee spacial separation between any 2 pairs in a balancing group, but the jointing sequence does ensure that no 2 quads adjacent in one length are joined to quads that are adjacent in any other length in a loading section. Crosstalk at high frequencies is dominated by inductive coupling and so the crosstalk between pairs in a balancing group is worse than the average for the whole cable, within-quad crosstalk being the worst case.

The 2·048 Mbit/s line systems are normally installed with both directions of transmission in the same cable, this being known as *single cable working*. In this mode, systems will be affected by both near-end crosstalk (NEXT) and far-end

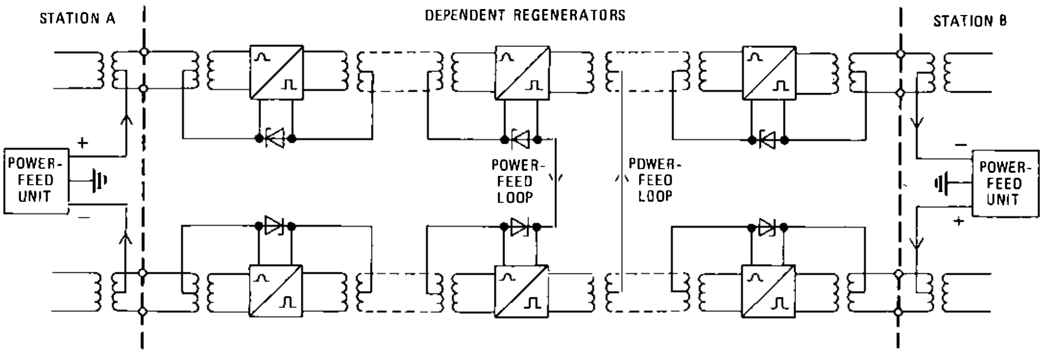
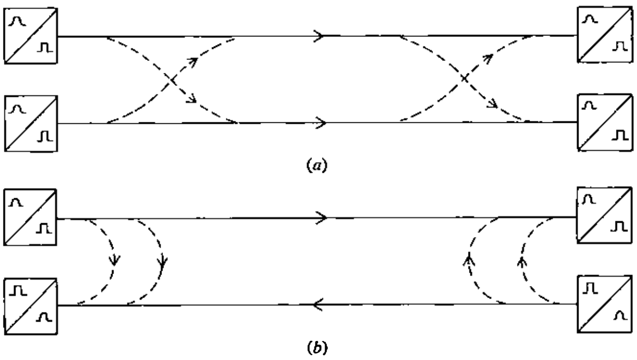


FIG. 15—Power-feeding arrangements



(a) Far-end crosstalk path
(b) Near-end crosstalk path

FIG. 16—Crosstalk paths

crosstalk (FEXT), as shown in Fig. 16. Due to the large differences in signal level (high-level output to low-level input), NEXT is far worse in its effect than FEXT and, hence, the signal-to-noise ratio at the regenerator decision point will be largely determined by the number of NEXT disturbing signals.

Balancing Group Allocation

To mitigate the effects of NEXT interference, it is necessary to segregate opposite directions of digital transmission in accordance with the following rules:

(a) In each balancing group, all cable pairs carry transmissions in one direction only.
(b) Adjacent balancing groups carry the same direction of transmission.
(c) Where possible, opposite directions of transmission are separated by a balancing group not used for digital transmission.

Utilization Factor

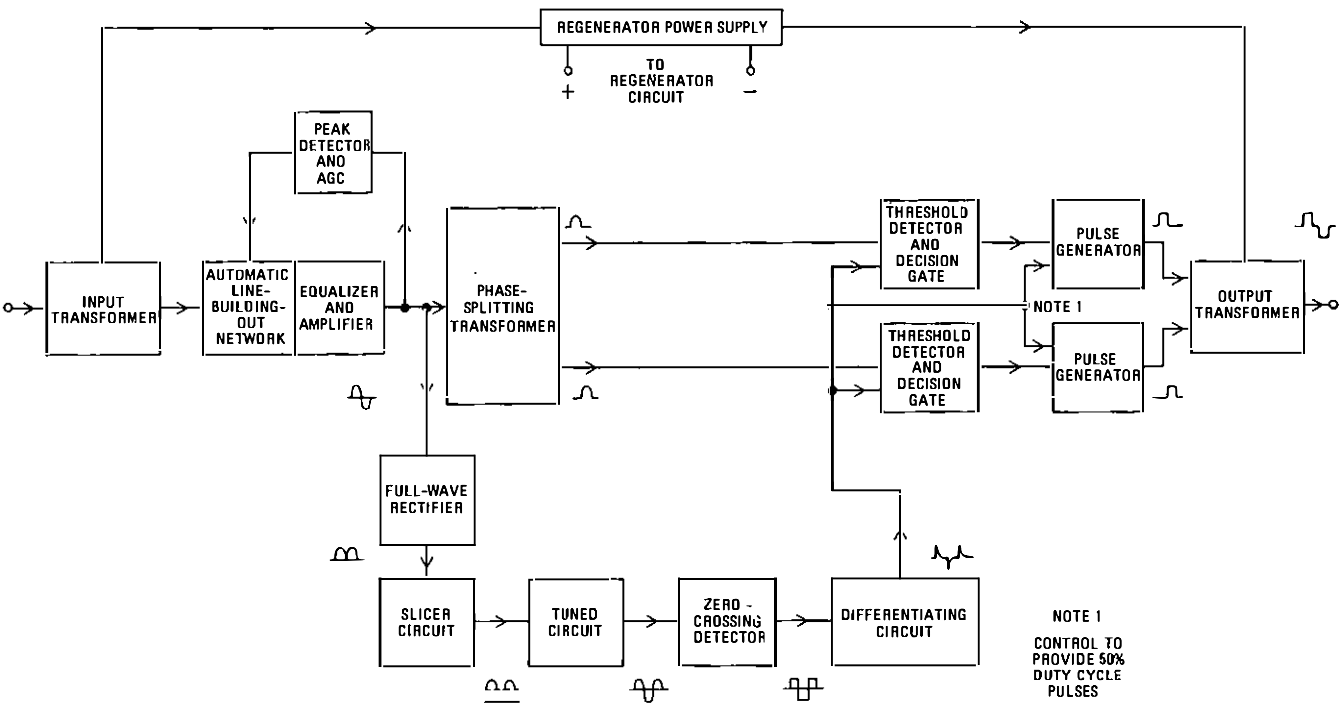
The maximum number of systems that can be installed on a cable is a function of the high-frequency crosstalk characteristics of the cable, which is not a controlled parameter. It is uneconomic to perform crosstalk measurements prior to each installation and therefore a sample number of cables has been measured to characterize the network in terms of the mean and standard deviation of the crosstalk attenuation. The maximum number of systems that can be provided in a given cable has been computed from this data on the basis that the probability of any one regenerator exceeding an unacceptable error rate is less than 1%. Accordingly, the maximum number of 2·048 Mbit/s bearer pairs per balancing group is restricted in accordance with the following rules:

(a) 50% of the pairs, up to a maximum of 24, may be provided in the inner balancing groups, and
(b) 25% of the pairs, up to a maximum of 24, may be provided in the outer balancing group. This restriction is due to the presence of the cable sheath which degrades the crosstalk attenuation.

The utilization factor depends upon the size of cable; for example, a fully utilized 504 pair cable can accommodate 72 DLSs (144 pairs).
The utilization factor can be increased by using 2-cable working; in this mode, each direction of transmission is in a separate cable and the limiting factor is FEXT. However, this mode of working is rarely used since it is difficult to justify on economic grounds due to the extra line plant required and the worsening of the system reliability due to cable faults.

REGENERATOR UNIT

A regenerator is required to reproduce at its output a replica of the original transmitted pulse sequence. The operation of a regenerator is best described by considering the 3 basic functions which must be performed on the received signal: equalization, re-timing and regeneration. A block diagram of a typical regenerator is shown in Fig. 17.



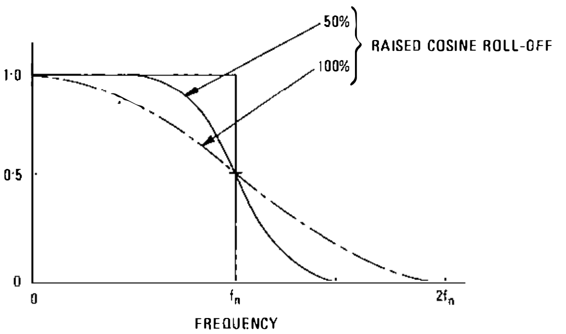
AGC: Automatic gain control

FIG. 17—Block diagram of a 2·048 Mbit/s regenerator

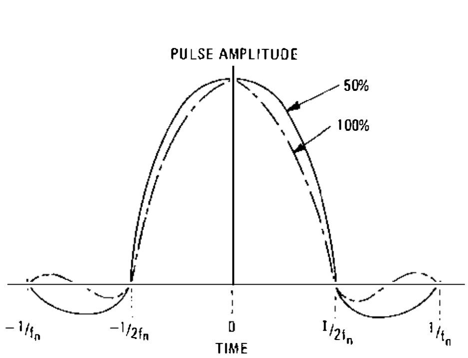
Equalization

The high frequency content of the transmitted signal will be severely attenuated, relative to the low frequencies, by the preceding cable section. The function of the equalizer is to compensate for this distortion and to shape the pulses into an acceptable form for regeneration. The equalized pulse shape should be such as to minimize the probability of incorrect threshold detection due to the combined effects of crosstalk noise and deviations from the optimum sampling instant. The equalizer can provide only a finite amount of gain; therefore, the channel frequency response, as measured between the equalizer output and the output of the preceding regenerator, will be that of a low-pass filter. The transmission of a rectangular pulse through such a network results in an oscillatory tail that can interfere with adjacent pulses, giving rise to inter-symbol interference. This effect can be overcome by using a channel response derived from the theory according to Nyquist, whereby the peak amplitude of any pulse is coincident in time with the zero crossings of the secondary lobes from adjacent pulses. Sampling at exactly this instant will ensure that there is no inter-symbol interference.

The minimum channel bandwidth which satisfies the Nyquist criteria is achieved by use of an ideal low-pass filter, which has a cut-off frequency (Nyquist frequency) numerically equal to half the digit rate and a linear phase characteristic. This ideal response cannot be realized in practice, but a useful approximation is achieved by using a raised cosine roll-off characteristic which is skew symmetrical about the Nyquist frequency at the half-amplitude point. The raised-cosine response is characterized by its cut-off frequency, which is normally between 50% and 100% greater than the ideal filter; the frequency and impulse responses are shown in Fig. 18. In practice, deviations from the ideal sampling



(a)



(b)

f_n: Nyquist frequency
$2f_n$: Digit rate

(a) Frequency response
(b) Impulse response

FIG. 18—Raised-cosine pulse shaping

instant are inevitable and, hence, the optimum raised-cosine channel shaping is a compromise between inter-symbol interference, crosstalk-noise bandwidth, and the complexity of the circuit realization.

Automatic-Equalization Range

Because the insertion loss of the cable preceding the regenerator may vary between 4–37 dB (at 1 MHz) due to differences in cable conductor gauge, cable length and cable temperature, it was considered advantageous to provide automatic equalization to simplify the installation procedure. Automatic equalization is achieved by use of an automatic line-building-out (LBO) network, followed by a fixed equalizer, a shaping network and an amplifier; these components provide the means whereby the desired channel response is achieved. The LBO is adjusted automatically (via an automatic gain-control loop, which is controlled by the peak amplitude of the equalized signal) to build out the preceding cable section to the maximum loss. The subsequent fixed-equalizer network simulates the inverse characteristic of a maximum-loss cable section, and the shaping network provides the required roll-off characteristic.

Re-Timing

The regeneration process requires a timing signal which is used to ensure that the equalized pulse sequence is sampled at the optimum instant in time, and to control the pulse width and repetition rate in the output stage. The frequency of the required timing signal is numerically equal to the digit rate and, in general, there is no component at this frequency in the spectrum of the transmitted signal. The timing signal is therefore derived from the equalized line signal by means of a non-linear process involving rectification and slicing, the resultant signal is passed through a resonant circuit (Q-factor 35–50) tuned to 2·048 MHz. The narrow timing-pulses required can then be obtained by differentiating the output of a limiting amplifier which operates on the output of the tuned circuit.

Regeneration

To produce a regenerated output, it is necessary to make a decision as to whether or not a pulse is present at a particular timing instant. The narrow timing-pulses are used to sample the equalized signal; samples being fed into a threshold detector which produces an output whenever the signal exceeds half the nominal pulse amplitude. Two threshold detectors are used, one for each pulse polarity, and the outputs from each are fed into separate output logic-gates which are triggered from the timing signal to produce rectangular output pulses of the correct duration and repetition rate. Finally, the outputs are combined and the regenerated signal (6 V peak–peak in 120 Ω) is transmitted to line via the output transformer.

Immunity to Noise

Crosstalk noise will be present at the regenerator decision point and there is a finite probability that, at the sampling instant, the noise will be of sufficient amplitude to cause the threshold detector to produce an incorrect output, thereby generating an error in the transmitted signal. When several systems share the same cable, the resultant crosstalk power has a near Gaussian amplitude distribution. Fig. 19 shows the relationship between the error probability† and the peak signal-to-RMS-Gaussian-noise ratio at the decision point of an ideal regenerator. A salient feature of digital transmission is illustrated in Fig. 19 in that a small change in signal-to-noise ratio will cause an order of magnitude change in the error

† An error rate of m in 10^n is an error probability of $m10^{-n}$

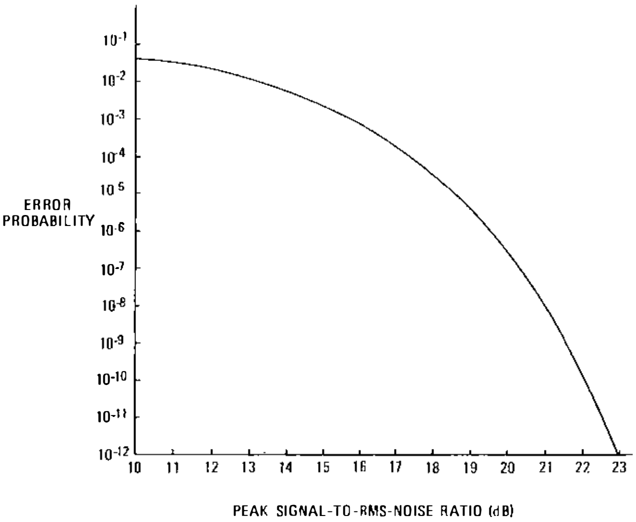


FIG. 19—Error-probability for given signal-to-noise ratios at decision point of ideal regenerator

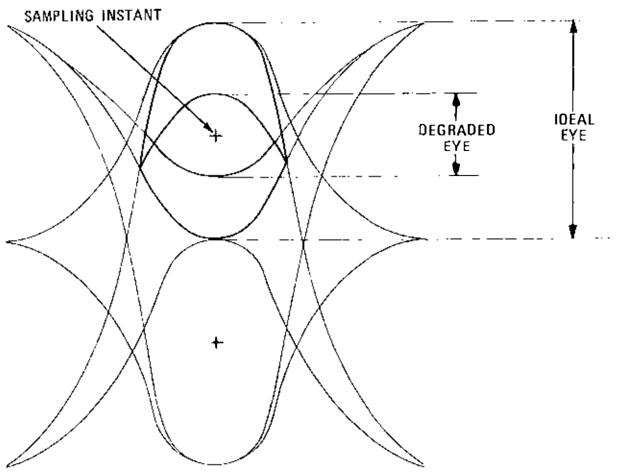


FIG. 20—Eye diagram

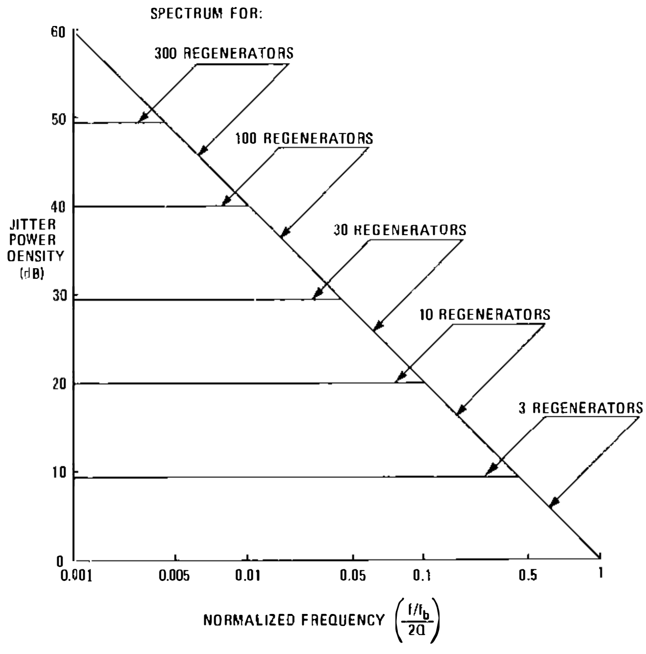
rate. It is for this reason that adequate performance margins are incorporated in the cable-utilization rules to allow for the deterioration of regenerator performance due to ageing effects. The design error rate for 2·048 Mbit/s regenerator sections is 1 in 10^7 but, on field-trial installations, an error rate of better than 1 in 10^{11} was achieved; the operating margin of 3 dB (see Fig. 19) is sufficient to cater for the addition of further systems on these partially utilized cables.

No allowance is made in Fig. 19 for regenerator imperfections and, in practice, the signal-to-noise ratio must be enhanced by several decibels to ensure satisfactory performance. A convenient technique that enables regenerator imperfections to be assessed, is the *eye* diagram. This is obtained by observing the waveform at the input to the threshold detectors on an oscilloscope, whose time base is triggered at the digit rate. If a random signal is being transmitted, successive sweeps of the trace build up a picture of all possible transitions, as shown in Fig. 20. The decision-making process is portrayed by the intersection of a vertical line, representing the decision instant, and a horizontal line representing the threshold level. The optimum intersection point is in the middle of the eye and, to reduce the probability of errors due to noise, the eye opening must be as large as possible. Regenerator imperfections will cause the eye opening to contract and are termed *eye impairments*. Vertical displacement of the eye will be caused by amplitude degradations due to inter-symbol interference and variations in equalizer output level. Horizontal displacement will be caused by timing degradations due to static timing-errors and jitter. The magnitude of the eye opening relative to the theoretically-derived eye provides a measure of both the regenerator impairments and the enhancement in signal-to-noise ratio required to keep the probability of errors the same.

Jitter

Jitter is defined as the short-term variation of pulses from their ideal position in time, and is a form of random pulse-width/position modulation arising from imperfections in the equalization and timing-recovery process. Jitter is due to fluctuations in the timing-pulse repetition rate, which is a function of the period between successive zero crossings of the timing signal. The most significant forms of jitter are those which add on a cumulative basis with the number of regenerators. These sources of jitter are caused by inter-symbol interference in the equalizer and amplitude-to-phase conversion in the limiting amplifier, both of which produce fluctuations in the time interval between zero crossings of the timing signal. Jitter from both these sources are dependent on variations in the pulse density of the transmitted signal and, consequently, the jitter will accumulate in proportion to the number of regenerators on a route. The tuned circuit in each timing-recovery loop acts as a low-pass filter to the jitter produced by preceding regenerators and hence it is the low-frequency pattern-dependent jitter that is cumulative,as illustrated in Fig. 21. The performance of digital multiplex equipment may be degraded by large amounts of jitter, but the route length of 2·048 Mbit/s systems in the BPO network is unlikely to cause any significant problems.

The pattern-dependent jitter produced by each regenerator is restricted by design to ensure that the displacement of the timing instant does not significantly impair the noise margin and, provided all regenerators have a similar tuned-circuit Q-factor, the jitter from preceding units can be tracked without incurring any additional impairment. The most



f_b: 2·048 MHz
Q: Q-factor of the timing-recovery circuit

FIG. 21—Simplified jitter-power spectrum

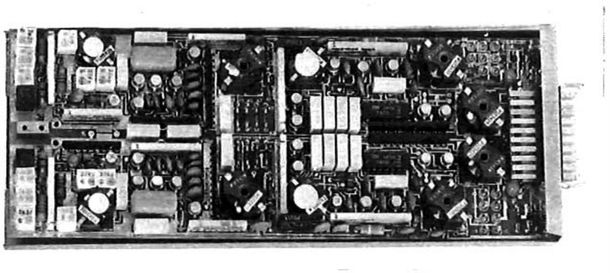


Fig. 22—Dependent regenerator unit

Fig. 23—Partially-equipped repeater case

significant form of non-cumulative jitter is due to drift in the resonant frequency of the tuned circuit, caused by environmental and ageing effects. As mentioned, 2·048 Mbit/s regenerators are specified to have a Q-factor between 35 and 50, and the frequency stability must be better than 0·3% to prevent a significant reduction in noise margin caused by the consequent phase shift of the sampling instant.

Design Features

Construction

A dependent regenerator unit (see Fig. 22) consists of 2 regenerators constructed on a single printed-wiring board which is totally enclosed in a metal container. External connexions are made via a common multi-way plug, which is compatible with that used for 1·536 Mbit/s regenerator units. The 2 regenerators operate in the same direction of transmission over 2 cable pairs which form a quad and therefore DLSs are installed in pairs.

Power Supply

The power to each dependent regenerator is derived from a remote constant-current supply via a Zener diode, which is connected between the centre taps on the line side of the input and output transformers. The 2 regenerators are power fed from the same line power-feed unit and, at the end of a power-feed loop, the Zener diodes are connected in series to complete the return path. The maximum power consumption is restricted to 700 mW per regenerator unit.

Components

The volume occupied by a dependent regenerator unit has been minimized to make efficient use of the space available within a repeater case. This has led to designs which have a very high component packing density on double-sided, plated through-hole printed-wiring boards. Extensive use is therefore made of thick-film resistor modules, and standard transistor-transistor logic circuits are normally used in the output stages and are required to comply with the standards of BPO specification D3000.

Reliability

Failures of dependent regenerators are expensive to correct and therefore reliability is a very important aspect. Experience with 1·536 Mbit/s line systems in the BPO has shown that the mean-time-between-failure (MTBF), estimated from fault returns, for a regenerator unit comprising 2 regenerators, is approximately 110 years. The production specification for 2·048 Mbit/s regenerator units calls for a 120 year MTBF design objective calculated from component failure rates.

HOUSING OF DEPENDENT REGENERATORS

Dependent regenerators are normally housed in pressurized repeater cases which are installed in footway boxes or manholes along a cable route. Up to 36 regenerator units can be accommodated in the standard BPO Case Repeater Equipment No. 1[7]. A 160-pair tail-cable assembly, of length up to 10 m, is jointed to the main cable and enters the repeater case through an air-tight gland. At the point of entry, the tail cable contains an epoxy-resin air block, which allows the repeater case to be separately pressurized via a Schrader valve in the lid. Connexions to the regenerator units and auxiliary services are made via flexible leads and mating sockets. Auxiliary items are mounted on a separate plate and comprise a fault-location filter (shown in Fig. 14), a pressure contactor, engineering-speaker-circuit loading coils and access plugs. A partially-equipped repeater case is shown in Fig. 23.

FAULT LOCATION FACILITIES

Fault location techniques are similar to those provided on 1·536 Mbit/s line systems and have been described in detail in a previous article.[8]

Error-Rate Alarm

The first indication of a fault is provided by an in-service error monitor on the associated PCM multiplex equipment,[4] further measurements at the line terminal patching panel will isolate the fault to either the line or the multiplex equipment.

Consideration is being given to providing a line-system error monitor at the line terminal, this would assist in localizing faults on tandem-connected DLSs, and also enable standby DLSs to be permanently monitored.

Power Feed Faults

A fault in a power-feed loop can be located by reversing the polarity of the applied voltage by means of a switch in the power-feed unit. This causes diodes, which are connected across the power-feed loop at each regenerator unit, to conduct. All diodes between the terminal test point and the fault will conduct, and measurement of the line current enables the location of the faulty section to be determined.

Regenerator Faults

To enable 2·048 Mbit/s regenerators to share the same repeater housings and associated ancillary equipment as 1·536 Mbit/s regenerators, a compatible fault-location method has been adopted. The *trio* method[8] is therefore used to locate a faulty regenerator, the only new item of equipment required being a 2·048 Mbit/s trio-pattern generator. Up to 18 dependent locations can be interrogated from a terminal station. Catastrophic failures can be accurately located by this method, but a regenerator with a marginal performance or intermittent fault, is difficult to locate. Serious consideration will therefore be given to an improved fault-location method for any second-generation line systems.

Service Communication Circuits

Several different terminal stations may be served by equipment at dependent locations. Therefore, access to a telephone exchange line is provided in each repeater case to enable calls to be set up via the public switched telephone network. In addition, a local speaker circuit is provided for communication between dependent locations.

SURGE PROTECTION

Line transmission equipment is susceptible to damage from high-voltage surges that are induced longitudinally along cable pairs due to lightning, or fault conditions on power-supply lines.[9] Present BPO policy requires that surge-protected 2·048 Mbit/s line equipment is provided only in areas of high risk to damage but, from 1980 onwards, all such line equipment will be of the protected variety.

The dependent regenerators are protected from damage by providing a shunt path for the surge energy between the centre taps of the input and output transformers. The surge energy is thus dissipated to some extent by the loss of the cable pair, and a discharge path to earth is provided via a protection device in the line power-feed unit at the terminal station. Circulating currents via the cable pairs between successive terminal stations are prevented by the isolation provided between the 2 power-feed loops at the power turn-round point.

Protection Arrangements

The cable pairs have only a finite degree of balance and therefore a residual transverse voltage will result from any longitudinal surge. Zener diodes are placed across both input and output circuits to prevent damage from excess transverse voltages. The shunt-protection devices comprise either gas-discharge tubes or high-power-rated Zener diodes, or a combination of both. A typical arrangement for surge protection of a dependent regenerator is shown in Fig. 24.

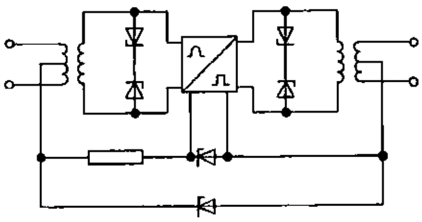


FIG. 24—Voltage-surge protection arrangement for dependent regenerator

Testing of Protection Circuitry

Lightning Protection

The applied waveform used to simulate lightning-induced surges is recommended by the CCITT, and has rise and decay times of 10 μs and 800 μs respectively. A regenerator is designed to withstand transverse voltages of up to 100 V, and longitudinal voltages of up to 2 kV applied between input and output ports or either port and earth.

50 Hz Induction

A regenerator is designed to withstand a longitudinal surge-voltage of amplitude 1300 V RMS and frequency 50 Hz, applied between input and output for a period of 0·1 s; the surge current is limited to 1 A.

CONCLUSIONS

Part 1 of this article described the multiplex equipment, and this part has described the operating principles, equipment and installation of the 2·048 Mbit/s line system. Part 3 will describe the signalling equipment.

When 2·048 Mbit/s line systems from 5 UK manufacturers were installed on several field-trial routes, the subsequent evaluation programme revealed interference problems which were investigated and rectified. As a result, the regenerator production specification includes additional clauses to enhance the immunity to crosstalk interference from 1·536 Mbit/s line systems, and impulsive noise from exchange switching equipment. Compatibility trials were successful and over the past 2 years the field-trial systems have performed satisfactorily under operational conditions. Contracts for production equipment have been placed and the first systems will be brought into service during 1978.

ACKNOWLEDGEMENTS

Photographs taken with the co-operation of GEC Telecommunications Ltd. and the Eastern Telecommunications Region.

REFERENCES

1 BRIGHAM, E. R., SNAITH, M. J., and WILCOX, D. M. Multiplexing for a Digital Main Network. *POEEJ*, Vol. 69, p. 93, July 1976.

2 KYME, R. C. The Junction Network. *POEEJ*, Vol. 69, p. 175, Oct. 1976.

3 BOLTON, L. J., and BENNETT, G. H. Design Features and Application of the British Post Office. 24-Channel Pulse-Code Modulation System. *POEEJ*, Vol. 61, p. 95, July 1968.

4 VOGEL, E. C., and MCLINTOCK, R. W. 30-Channel Pulse-Code Modulation System, Part 1–Multiplex Equipment. *POEEJ*, Vol. 71, p. 5, Apr. 1978.

5 CCITT Recommendation G703. Orange Book, Vol. III–2.

6 LAWRENCE, A. J., and DUFOUR, I. G. Audio Junction Cables. *POEEJ*, Vol. 69, p. 239, Jan. 1977.

7 CLINCH, C. E. E., and STENSON, D. W. Housing of Repeater Equipment Underground. *POEEJ*, Vol. 56, p. 158, Oct. 1963.

8 BENNETT, G. H. Testing Techniques for 24-Channel PCM Systems. *POEEJ*, Vol. 65, p. 182, Oct. 1972.

9 HALL, B., and LILLY, C. J. The Effects of Power-Supply Transients on British Post Office Cable Communication. *POEEJ*, Vol. 67, p. 240, Jan. 1975.

Teleconferencing: A Service for the Businessman

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UDC 621.395.348: 061·3

A number of different teleconferencing systems have appeared in recent years, and studies suggest that audio and visual teleconferencing systems have the potential to replace a significant proportion of business meetings that today involve travel. To realize this new business potential and the call revenue it represents, new systems must be economic, effective and compatible. This article briefly surveys the potential of teleconferencing systems and proposes some fundamental design requirements which are derived directly from the needs of the user. The design of an experimental audio-teleconference terminal is then described as an example of the design principles in action.

INTRODUCTION

In this era of energy crises, job dispersal, devolution and the paradoxical increase in central administration, serious consideration is being given to the use of new and existing telecommunications services as a substitute for business activities that involve travelling. A major example of such services falls within the scope of the generic term *teleconferencing*. Unfortunately, teleconferencing can also be a confusing term which conjures up different, yet highly specific, pictures in the minds of both customers and telecommunications staff alike. This article tries to allay some of that confusion by outlining for audio-teleconferencing systems, common fundamental requirements which are derived from the needs of the user and are not dictated by particular engineering restrictions. Teleconferencing, with its considerable savings in time, money and energy, could become as familiar a means of communication as the ubiquitous telephone, a natural aid and adjunct to the conduct of business. In the longer term, by improving group-to-group communication, it may influence management methods and improve business efficiency.

TYPES OF SERVICE

The enormous variety of teleconference services and equipment available can be broadly divided into 3 categories. However, the categories are not mutually exclusive, and some services are hybrids falling between the categories.

Category 1—Multi-party handset telephony

A group of individuals, each in a separate location, confer using their normal telephone instruments (or loudspeaking telephones) connected via the public switched telephone network (PSTN) and a special conferencing bridge.

Category 2—The group-to-group audio conference

Two groups of people, each in a separate location, confer using special audio equipment connected by the PSTN or private circuits and supplemented, if necessary, by document facilities.

Category 3—The group-to-group audio/visual conference

Two groups of people, each in separate locations, confer using special audio terminal equipment connected by private circuit links and a 2-way vision link.

Each category has its own peculiar engineering problems and, particularly in categories 1 and 2, which together constitute audio-teleconferencing, a wide variety of solutions has been demonstrated or proposed. These solutions are not necessarily compatible; it is therefore important at this early stage in the development of audio-teleconferencing to determine if there are fundamental requirements that are linked to the needs of users and which must be satisfied if the service is to become widespread. In many cases, audio-teleconferencing systems are adaptations or extensions of loudspeaking, or *hands-free*, telephone equipment. It is, however, dangerous to assume that the principles and restrictions accepted for hands-free telephony are equally acceptable when considering audio-teleconferencing.

THE POTENTIAL FOR TELECONFERENCING SERVICES

Although some teleconferencing equipment has existed for many years in a number of countries, its impact has been small and the pattern of use very variable. As no clear picture of the potential for these services was emerging from the corporate experience, extensive studies have been undertaken which try to predict the potential market in the UK for audio and visual teleconferencing. These studies examined the nature and content of business meetings, the number of participants and the economics, to determine which meetings could, without loss of effectiveness, be substituted by a telecommunications service. The projection is expressed in terms of the anticipated number of business meetings in 1990, and the surprising conclusion is, using conservative assumptions, that 53% of all meetings that involve travel could be substituted by teleconferences. The concept of effective substitution is, of course, a key issue. The participant in a remote teleconference has no access to the subtle forms of interpersonal communication that are available in a face-to-face meeting, and it is therefore natural that the inexperienced user should initially consider the teleconference as inferior. Research work suggests, however, that the value of the face-to-face nuances and subtleties is much less than is commonly supposed. Except in a small number of special cases, they are unlikely to influence the outcome of a meeting. The special cases of counselling, bargaining and getting-to-know-people, favour the face-to-face meeting as would be expected, but these constitute a minor part of the totality of meetings. For the routine meeting, between people already well known to each other and jointly concerned with project co-ordination, information exchange and discussion, the telecommunication alternative is wholly adequate and does not colour the outcome. Indeed, in as much as teleconferences tend to be more businesslike, shorter and not prone to after-lunch effects, they are more efficient than the face-to-face encounter in addition to saving time and travel.

Recent studies show that even these conclusions may be pessimistic and that bargaining activity can be, and is being

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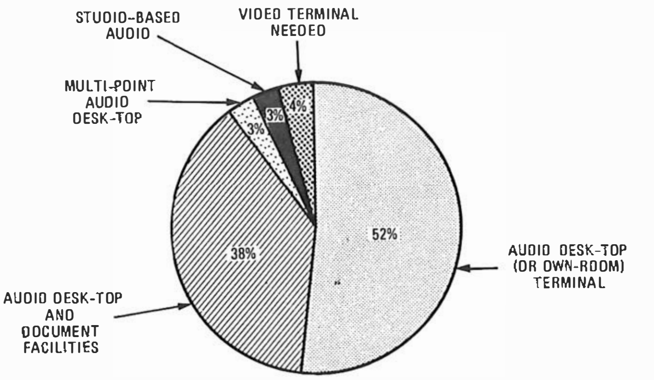


FIG. 1—Projected 1990 terminal requirements for the UK

accomplished by teleconference, particularly where immediacy is a vital factor. The study of general potential, however, is deliberately conservative, and where doubt existed concerning the suitability of the telecommunication alternative, the study has opted for face-to-face contact. Similarly, in subdividing the teleconferences between different types of communications media, if doubt existed, the more sophisticated service was assumed. The division of types of services predicted for 1990 is shown in Fig. 1. The striking result is the percentage of teleconferences that can be conducted solely by audio, or audio with document, facilities. If this prediction is borne out in practice it is an important result for telecommunications administrations, for it indicates a growth potential for the new audio-teleconferencing services on the existing planned network and represents a source of additional income that will not require capital outlay on a completely new, special-purpose overlay network dedicated to teleconferencing.

In absolute terms, the potential market is large. Of the predicted 170 million meetings in the UK in 1990, it is suggested that some 67 million could be carried by audio-teleconferencing services, thereby earning a revenue in the region of £100M at today's prices. The bulk of these teleconferences would use conventional telephone line and switching plant, but would originate from new types of terminal equipment.

At this stage it must be emphasized, however, that these are projections of market potential and give no indication of how the market can be won, or the way in which it will evolve.

FUNDAMENTAL REQUIREMENTS FOR AUDIO-TELECONFERENCING

Though there are a number of teleconferencing systems available, they differ widely in their method of operation, sophistication and facilities. The necessary blend of features and facilities for successful audio-teleconferencing is not, therefore, immediately apparent, and care must be taken to distinguish between fundamental requirements and optional extras. Important technical questions on quality, loudness and bandwidth still need to be answered and, in each case, the ultimate choice must be made in the light of users' reactions. In examining the current ideas on the fundamental requirements, it must be stressed that there is no consensus to the views expressed. Opinions still vary widely on the direction that future development will take.

Loudness of Received Speech

Loudness is a key parameter. An audio-teleconferencing system that is too quiet is extremely tiring and frustrating to use, even when there is little background noise in the terminals. Not only does the attention of the participants wander as the strain of listening becomes unpleasant, but the teleconference tends to break up into 2 sub-meetings, one in each location. Surprisingly, the subjective effect of loudness is not a linear function of the sound level of the received speech in given room conditions. There appears to be a critical level for the received speech in a teleconference, above which conferees relax and participate freely and below which the strain of listening swiftly takes its toll and conferees tire rapidly and cease to concentrate. The desired listening level for received speech (which is clearly above the critical level) depends on the room noise, the room acoustics, hearing acuity of the participants and perhaps on their attitude. As a first postulate, it was assumed that the received speech level heard by an individual over the system should be the same as the speech level he would hear from a colleague in the same room. Such a criterion is, however, simplistic, for at his own terminal the conferee receives speech from a colleague without bandwidth limitation or distortion and can also take advantage of binaural hearing effects. The reported work on preferred listening levels[2,3] for conference telephony is not, unfortunately, directly applicable as it has concentrated on monophonic, broadband speech, which does not indicate the effect of limiting the received speech signal to the telephone bandwidth of 0·3–3·4 kHz. A tentative interpretation of the available evidence suggests, however, that, if the speech is restricted to the telephone bandwidth, it must be presented in the terminal at a level some 3–6 dB† greater than if it were broadband. The advantage of a national service obtained by using the PSTN for transmission must be balanced, therefore, against the increase in the loudness required in the terminals. Furthermore, a conference system that has been designed for use on broadband private circuits will not necessarily perform satisfactorily on circuits having telephone bandwidth, nor may it be entirely straightforward to add the extra loudness required.

Room noise also has an important effect on the preferred listening level. Fig. 2 shows the variation in the received level of telephone-band loudspeaker speech preferred by a conferee and the change in his speech level as the room noise increases. Between a room noise level of 35 dB(A)* which represents a quiet room, and a level of 55 dB(A), the preferred listening level increases by approximately 8 dB and the vocal level by about 5 dB. The most unfavourable situation occurs when a quiet room is connected to a noisy one via a telephone link. The speech level of the conferees in the quiet room is relatively unaffected by their ambient noise level, whereas that of the conferees in the noisy room is increased. Similarly, the conferees in the quiet room are content with a

† dB (sound pressure level)—20 times the logarithm to base 10 of the ratio of the RMS sound pressure level to the reference sound pressure level; unless otherwise stated the reference sound pressure level in air is taken as 2×10^{-5} Pa

* dB(A)—The sound pressure level as measured in decibels, by an instrument having a shaping filter, reference *A*, which approximates to the response of the human ear

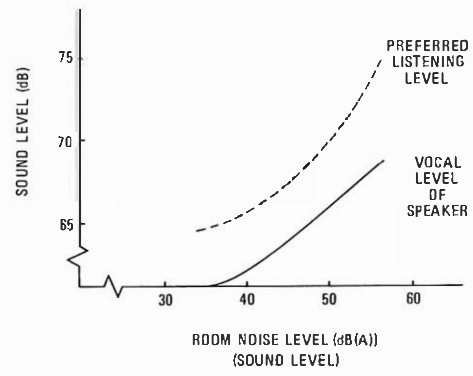


FIG. 2—Variation of preferred listening level and vocal level with room noise

lower listening level than those in the noisy room. In practice, however, the conferees in the quiet room receive the raised speech level from the conferees in the noisy room, and the latter receive the speech of the conferees in the quiet room at too low a level. This is clearly an unsatisfactory situation and a method of rebalancing the conferencing-system gains to take into account the ambient noise in the terminals is desirable.

Noise is a function of the terminal room's environment, but the acoustic properties of the room itself also affect the speech level received by the conferee. In a typical room with reflecting surfaces, the received sound reaches the listeners' ears not only via a direct path from the sound source, but also by indirect paths resulting from single or multiple reflections at the walls. This reverberant sound-field adds to the direct sound-field and may enhance the total received sound, as those who sing in the bath are aware. While it may be straightforward to design a stable audio-teleconference system considering only the direct received sound-field, few meetings take place in completely damped or anechoic conditions. Indeed, it is not desirable that they should, as the typical reverberant rooms (in which we all live) give rise to a quality of speech that we all consider pleasant and normal. In an audio-teleconference system, it is the reverberant sound-field in the conference room that governs the stability of the system and, hence, the loudness performance that can be achieved. This sound-field varies from room to room depending on the amount of acoustic absorbent the room contains.

In a simple system, a compromise loudness setting can be arrived at that will be satisfactory in a range of room conditions but will be unsatisfactory in others. Alternatively, a volume control can be provided, but this may bring other serious performance disadvantages. The most satisfactory approach is to make the audio-conferencing terminal equipment adapt its performance to the environment in which it is used. With modern technology, this is becoming feasible and economic, but there remains much work to be done to quantify the interdependence of the factors affecting loudness to ensure that the performance provided by the equipment is optimum in all conditions. It is unlikely that the final analysis will become available for several years, as the work must involve the subjective response of groups of individuals, and there are a large number of permutations of possible conferencing situations to be examined.

Bandwidth and Speech Quality

Ideally, a bandwidth in excess of 10 kHz should be provided if a high-quality audio-teleconference is required. Providing such a service via BPO networks is difficult for, although a limited number of special music-quality private circuits exist for broadcast use, the overwhelming majority of links, both in private-circuit networks and the PSTN, are limited to the telephone bandwidth. If, however, the extra loudness mentioned earlier is available, conferees presented with a properly optimized telephone-bandwidth system are usually unable to recognize that the system is quite severely band-limited, unless they are deliberately allowed to listen to a wideband system and to make a direct comparison. The quality of speech produced by the receiver in a telephone handset is widely, and wrongly, taken as the quality to be expected from the telephone network. Indeed, naïve users are often surprised by the network performance revealed by the electro-acoustic devices forming an audio-teleconference system. From the evidence to date, it can be concluded that telephone bandwidth is adequate for audio-teleconferencing, provided the available bandwidth is used to full advantage. In particular, attention must be paid to intelligibility, as this may suffer badly if there is a marked loss of frequencies at the upper end of the telephone band.

Closely linked to the questions of bandwidth and intelligibility is the ability of the listener to recognize the voices of individual participants at the remote terminal. The need for speaker identification varies from meeting to meeting. Some organizations may require a separate system of speaker identification so that appropriate rank protocols can be observed; others may wish to identify the affiliation of a particular speaker. There is a suggestion that some people find voice recognition difficult under normal circumstances and would, therefore, have particular difficulty in recognizing the distant people in an audio-teleconference by the timbre of their voices alone. This could be particularly embarrassing if, for example, the person in difficulty was also taking the notes of the meeting. The provision of better quality or wideband audio systems will not necessarily help. They may merely make it easier for the people with recognition ability to make an identification. Further human factors work is desirable in this area, for the implications of speaker identification are not trivial, either technically or economically.

Voice-Activated Switching

Voice-activated switched attenuation (often called voice-switching) has been introduced extensively into loudspeaking telephones and audio-teleconferencing units to maintain overall stability and prevent *howling*; for 2-wire operation on the present PSTN it is an essential feature. Fig. 3 illustrates a typical situation which could be found with existing equipment. The attenuation between the 2 parties in a telephone call can be large and require considerable amplification of the speech signal received from the local telephone line. As the trans-hybrid loss is usually quite low, under listening conditions a correspondingly large amount of switched attenuation must be included in the transmit path of the instrument to prevent instability around the acoustic path and hybrid loop. Similarly, when transmitting speech, the attenuation must be placed on the receive path of the instrument. Other solutions to the problem are difficult to find. The range of attenuation to be expected on calls made via the national network is unlikely to change significantly while analogue transmission remains in the local network, and only minor improvements can be effected in the acoustic feedback between the microphone and loud-speaker. Increasing the trans-hybrid loss would make an important difference, but no cheap and effective method of automatic balancing has yet been found. From a technical point of view, therefore, voice-switching is a necessity but, from the marketing and users' point of view, is it a serious restriction on the acceptability of widespread and efficient audio-conferencing? Opinion is divided. One recent study has shown that the present BPO loudspeaking telephone (LST 4D) is used by some customers for teleconferencing, particularly when it enables an urgent need to be satisfied. The range of views expressed was nonetheless wide: from the enthusiast who regularly used the equipment for teleconferencing, to the individuals who resorted to telecommunications only if the need was imperative. A severely voice-switched system was used in one country by a government department as

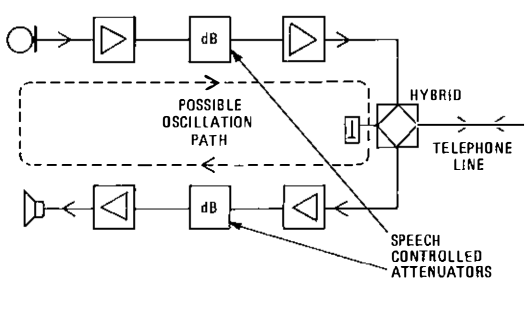


FIG. 3—Voice-switched attenuation

an alternative to a difficult journey of several thousand kilometres over near-arctic terrain in the winter months. Such examples cannot in themselves constitute proof that the principle of the equipments' operation is satisfactory for the majority of meetings; they merely suggest that given the need, a conference can be achieved.

At this stage in the development of audio-teleconferencing, it would seem sensible to regard any system, whose mode of operation makes the user continuously and consciously interact with it, as potentially less suited to the task than the *invisible* system, which neither intrudes nor inhibits the users' normal behaviour.

In summary, therefore, it is suggested that the basic requirements for any audio-teleconferencing system should be:

(*a*) adequate loudness of the received speech from the distant terminal,
(*b*) intelligibility,
(*c*) unobtrusive equipment whose operation and presence does not alter the user's normal behaviour,
(*d*) an ability to operate in a wide range of rooms without special acoustic room treatment, and
(*e*) an ability to provide a range of special facilities; for example graphics or document facilities as optional extras.

To these specialist requirements can be added the general requirements:

(*a*) simplicity in use, both during call set-up and conferencing,
(*b*) high reliability, and
(*c*) modest cost.

HABITS, ATTITUDES AND DEMAND

The potential market for audio-conferencing can be realized only if naturally conservative conferees can be persuaded that teleconferencing is an attractive alternative to travel. The subject of attitudes must be approached carefully for, in many cases where teleconferencing will be provided as an alternative to travel, the potential conferee will still be able to make a choice. Enthusiasm on the part of efficiency experts or communications managers may ultimately prove quite spurious, whatever the economic and technical merits of the system, if potential users vote with their feet and shun the service. Unfortunately, such experience is commonly found in many existing audio-teleconference systems. An initial burst of use gradually dies away, to leave only a residual core of enthusiasts and, although the equipment earns its keep, its use is far below the potential the system can offer. Avoiding complicated procedures and distortion of a meeting by the intrusion of the equipment clearly has an important effect, not just in an objective view of the efficiency of the audio equipment, but in influencing the reaction of first-time users. Perhaps the best that the teleconference equipment designer can achieve at this stage in the development of the concept is a system without irritating and manifest weaknesses. Not only must the equipment work well and be economic, it must be reliable, easy to access and simple to operate. With these features, the initial complaints and criticisms can be allayed and the real business of education and attitude changing tackled as part of the marketing process. After all, our grandparents naturally communicated by letter with only special recourse to the telephone. So today the businessman turns to the car, the aircraft and the train to go to that meeting and probably never pauses to consider the alternative.

It is not surprising that the explicit demand for audio-teleconferencing should at present come from organizations with particular specialist needs who are not necessarily representative of the ultimate users. Several room-to-room audio-teleconferencing systems have come on to the market in recent years, despite the prediction in the survey of market potential of an overwhelming ultimate demand for small desk-based terminals. The apparent contradiction is easily resolved, however, for, in the main, the present demand is seen as originating from the large companies with many locations who are seeking to improve managerial cohesion and efficiency by the use of teleconferencing. Such organizations often also require multi-point facilities, particularly between their head and regional offices and their manufacturing units. Moreover, their very size and complexity generates specialist departments concerned with managerial efficiency and costs, part of whose function is to be aware of new developments such as audio-teleconferencing and to provide detailed economic analysis of their possible benefits to the company. Early systems will be tailored to the perceived needs of such customers and the success of these initial attempts at teleconferencing will, inevitably, greatly influence business attitudes. Nonetheless, the ultimate aim must be kept in mind and, if possible, the first-generation equipment should be capable of logical development towards the ultimate market need. In particular, a determined attempt must be made to define the preferred principle of operation at the outset so that each new generation of equipment can be functionally compatible with its predecessors. If this can be achieved, the momentum of the teleconferencing concept can be steadily built up with business increasing as the availability of compatible terminals increases.

A ROOM-TO-ROOM AUDIO-TELECONFERENCE SYSTEM

As an example of the application of the foregoing principles, a room-to-room audio-teleconference system, developed at the BPO Research Centre and capable of connexion via the PSTN is briefly described. The service provided by the system meets the requirements for group-to-group facilities (category 2) and it is aimed at the initial market requirements.

To ensure minimum intrusion of the equipment into the conference, an open-loop audio system has been adopted. The simple configuration is shown in Fig. 4, and a 4-wire circuit (or its equivalent), with separate transmit and receive paths, is required. This can be most readily and simply achieved on the PSTN by using a pair of exchange lines, one line for each direction of transmission.

A typical terminal could take the form shown diagrammatically in Fig. 5 and pictured in Fig. 6. There are 3

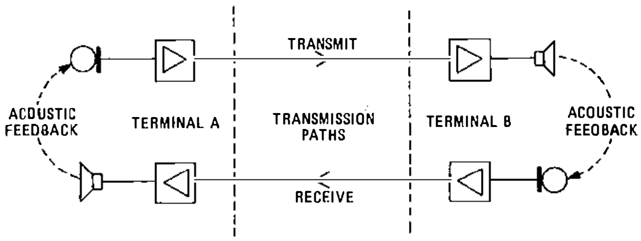


FIG. 4—A 4-wire circuit for an open-loop audio-teleconference

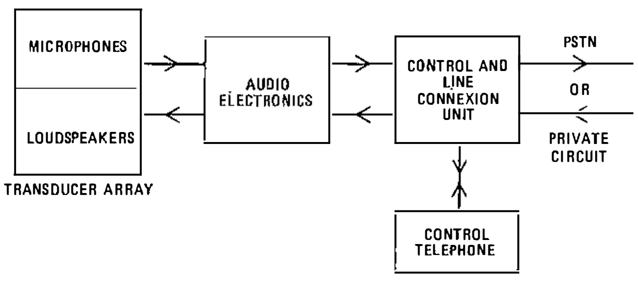
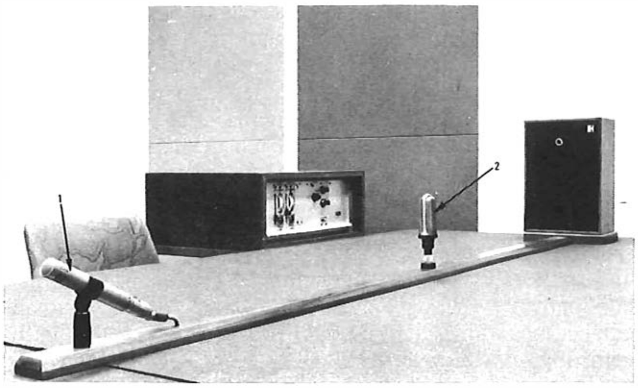


FIG. 5—Simplified block diagram of an audio-teleconferencing terminal



1: Cardioid microphone. 2: 'Figure-of-eight' microphone.

FIG. 6—View of a typical experimental terminal

major elements: the transducer array, the audio electronics and the line connexion unit.

The transducer array shown in Fig. 6 has been designed for use on existing tables. The loudspeaker is placed at the end of the table and conferees are free to sit at any other position around the table. The microphone pick-up field is approximately uniform around the table and conferees can talk naturally without speaking to a particular microphone, without raising their voices, without having to sit upright or to press access keys. Anyone who is seated at the table and is within 0·75–1 m of the microphone array can be heard at the distant terminal at a speech level some 3–6 dB higher than that heard by his colleagues at his own terminal, thereby fulfilling the loudness requirement. People in a second row around the table are also audible to the distant conferees, but their speech is quieter and slightly more reverberant. The arrangement is, moreover, fairly tolerant of the acoustic conditions of the room in which the terminal is used. Generally, acoustic treatment is only required to isolate the conference room from excessive external noise.

Measurements indicate that the arrangement should perform well in the range of acoustic conditions that have been found in a survey of UK offices and conference rooms.

The audio electronics unit is built from conventional pre-amplifiers, mixers and power amplifiers. A limiter is included in the transmit chain to limit the maximum speech level sent to line in accordance with BPO requirements, and a filter is provided to restrict the outgoing signal to the telephone band. The only unusual feature is a 5 Hz frequency shifter. The frequency response of the complete audio loop without the frequency shifter in operation is very irregular, caused mainly by interference effects from the many sound modes in the terminal rooms. As the loop gain is increased the system oscillates, or howls, at the frequency where there is the largest peak in the response. The frequency shifter capitalizes on the rapid fluctuation of the loop characteristic with frequency and attempts to reduce or cancel a peak in the response by merging it into the adjacent trough. Each time a particular signal passes round the teleconference loop it is shifted by 5 Hz. A signal occurring at a peak in the frequency response is, on average, shifted to a trough for the second transit and the probability of a regenerative build-up of oscillation is reduced. In practice, the addition of the frequency shifter, the operation of which cannot be detected by the conferees, allows a small amount of extra gain to be added to the audio loop before 'colouration' of the received speech commences. The actual oscillation point is raised by approximately 10 dB. The benefit in increased speech level is therefore modest, but the margin against howling is increased and this further improves the tolerance of the system to extreme acoustic environments.

The line connexion unit can take 2 forms:

(*a*) a simple unit to connect the system to private circuits, or

(*b*) a more complex control unit which connects the terminal to the PSTN.

The simple unit merely contains the line isolation and protection features specified by the BPO for equipment connected to private circuits. Adequate gain is provided within the unit to compensate for the loss of the different types of speech-band private circuits which are in use.

The unit used for connexion to the PSTN has several additional functions. This unit supervises the setting-up of 2 telephone calls over the normal telephone network to the distant terminal, and equalizes the resulting transmission paths to make them suitable for audio-teleconferencing. The absolute attenuation, the attenuation/frequency characteristic and the circuit noise of each connexion will differ, and the equipment must therefore measure and correct each factor. Extra signal gain can be added only at the receivers at either end of the transmission path, as the signal transmitted to the telephone line must not exceed a predetermined level to avoid overloading of frequency division multiplex (f.d.m.) line equipment. Thus, although the absolute attenuation of the transmission path can be quite simply corrected, the additional gain at the receiver amplifies any line noise and crosstalk, both of which will impede the teleconference. It is therefore necessary to apply a noise-reduction technique to the transmission path by processing the transmitted signal and subsequently restoring it to its original form at the receiver.

CONCLUSIONS

This article has identified a number of basic requirements for an audio-teleconferencing system which may be essential if customers are to adopt audio telecommunications services as an alternative to business travel. The factors discussed do not necessarily form a comprehensive list of all the relevant parameters but they include many of the principal problems and dilemmas that face the equipment designer.

It is recognized that opinions among designers still differ widely on the way in which audio-teleconferencing could, and should, develop. There is a need for a consensus, however, if the new systems and services in this area are to be complementary and compatible for both national and international applications.

ACKNOWLEDGEMENTS

The author would like to thank all those colleagues who have contributed by their detailed work or in discussion to the philosophy presented in this article.

References

[1] TYLER, M., CARTWRIGHT, B., and COLLINS, H. A. Demand for Teleconference Services. BPO Telecommunications Long Range Intelligence Bulletin, No. 9, 1977.

[2] GARDNER, M. B. Effect of Noise on Listening Levels in Conference Telephony. *J. Acoust. Soc. America*, Vol. 36, No. 12, p. 2354, 1964.

[3] GLEISS, N. Preferred Listening Levels in Telephony. 7th International Symposium on Human Factors, Montreal, 1974.

Dual Polarization Technology

Part 3—The Influence of the Propagation Path

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UDC 621.3.095.1: 621.37.029.6

This article, which concludes a series of articles on dual-polarization technology, discusses the effects of the propagation path on the overall cross-polar performance of 4/6 GHz communication-satellite systems. A programme of slant-path propagation measurements, which commenced in January 1977 at the Goonhilly earth station site, is also described and some of the interim results are discussed in detail.

INTRODUCTION

The propagation path between an earth station and a communication satellite includes the ionosphere and the troposphere. Both of these regions may cause polarization changes, but by different mechanisms. The polarization changes are frequency dependent and therefore assume different levels of importance according to the operating frequency. This article is concerned primarily with the introduction of dual polarization into the 4 GHz and 6 GHz frequency bands.

The main object of studying propagation effects in dual-polarized systems is to enable the overall performance of a satellite link to be predicted as a function of time. To this end, COMSAT*, acting as the manager of the International Telecommunications Satellite Organization (INTELSAT) system, has undertaken a number of theoretical and practical studies of rain depolarization. However, relatively small changes in location can result in significant variations in propagation conditions and it is therefore desirable that measurements are made at particular earth station sites, thereby minimizing the uncertainties involved in predicting the performance of dual-polarized systems operating via those earth stations.

IONOSPHERIC EFFECTS AND THE CHOICE OF POLARIZATION STATE

The ionosphere is a region of space surrounding the earth, and consists of electrons and ions that move in a gyrating manner. The principal effect of the ionosphere on wave polarization is known as *Faraday rotation*. This effect is attributable to anisotropy in the ionosphere, and results in the rotation of the plane of polarization of linearly-polarized waves. The angle of rotation is proportional to the square of the wavelength and is very dependent on the electron density and, therefore, on the level of solar activity. The direction of rotation is determined by the orientation and the direction of the signal path relative to the earth's magnetic field, and is opposite for up-link and down-link signals. Typically, at 4 GHz, a linearly-polarized wave can suffer a rotation of 3° and, unless some means is available of separately controlling the transmit and receive polarization angles of the earth-station aerial, the cross-polar discrimination (XPD) is limited to about 25 dB for a down-link operating at 4 GHz, and 32 dB for an up-link operating at 6 GHz. Methods of counteracting Faraday rotation by adjustment of the earth-station aerial-feed polarization were described in Part 2 of this series[16].

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However, with circular polarization, the effects of Faraday rotation are avoided and it is the natural choice for systems operating in the 4 GHz and 6 GHz frequency bands. At higher frequencies, the choice of polarization state is more likely to be influenced by tropospheric effects.

TROPOSPHERIC EFFECTS AND RAINFALL DEPOLARIZATION

Precipitation (which includes rainfall, snow, ice and hail) occurs within the first 10 km or so of the earth's atmosphere; this region is known as the *troposphere*. Precipitation is the main source of disturbance to microwave propagation on earth–satellite transmission paths, and a number of papers have been published[17] on the effects of rainfall on polarized waves.

The underlying mechanisms of rain depolarization and, in particular, how they affect dual polarized transmissions at 4 GHz and 6 GHz are of interest. Raindrops falling through the atmosphere take up a non-spheroidal shape, owing to the effects of air resistance. A linearly-polarized wave which is polarized parallel to an axis of symmetry of a raindrop has its amplitude and phase changed, but its polarization state remains unaltered. When the same linearly-polarized wave is incident at some other angle, the 2 axes of symmetry of the drop produce different attenuations and phase shifts, and these differences (differential phase-shift and differential attenuation) change the polarization state of the wave. The change in the polarization state is determined by the magnitudes of the differential phase-shift and the differential attenuation and by the relative orientations of the wave and the raindrop. For the more general case of an elliptically polarized wave, having the major axis of its polarization ellipse inclined at an angle to the axis of symmetry of a rain drop, differential phase-shift and differential attenuation introduce further ellipticity.

There is an important difference between the effects of differential phase-shift and differential attenuation. A given amount of differential phase-shift always rotates the axes of the wave by the same amount and in the same direction, irrespective of the hand of polarization of the incident wave. Thus, for frequency re-use systems, orthogonality is preserved. Differential attenuation, however, introduces non-orthogonality, because the direction of rotation of the polarization ellipse depends on the direction of rotation of the incident wave. This aspect has important implications regarding methods of cancelling rainfall depolarization. However, in the 4 GHz and 6 GHz frequency bands, both theoretical calculations and measurement results show that depolarization in rainfall is mainly due to differential phase shifts.

In practice, of course, the incident wave is perturbed by an ensemble of raindrops in the path, and each of these drops tends to be oriented slightly differently. The size of individual raindrops varies, and statistical distributions, which change with rainfall rates, are used to describe the dimensions of the raindrops.

In the case of orthogonal linearly-polarized waves being transmitted through a volume of rainfall there will be maximum and minimum values of XPD, according to the alignment of the incident waves and the mean canting-angle of the raindrops. (The tilt angle of the raindrops, with respect to the local vertical, is referred to as the *canting angle.*) In theory, if all the raindrops were aligned in the polarization plane then no cross-polar coupling would occur. This is a factor in favour of choosing linear polarization for frequency re-use and, at frequencies above 10 GHz, where the Faraday rotation problem is less significant than at 4 GHz or 6 GHz, there is usually a clear preference for using linear polarization.

It can be shown[18] that the minimum XPD (that is, maximum coupling) between linearly-polarized waves is always numerically equal to that for circular polarizations under the same conditions. Thus, since it is simpler to derive an expression for XPD for the case of circular polarization than for linear polarization, circular polarization is assumed in the following discussion.

Assuming that all raindrops are equally oriented, the XPD for circular polarizations

$$= -20 \log\left\{\left(\frac{T_2 - T_1}{T_2 + T_1}\right)|e^{j2\theta}|\right\} \text{ decibels} \quad \ldots \quad (1)$$

in which

$$T_1 = e^{\{-(\alpha_1 - j\beta_1)L\}}, \quad \ldots\ldots \quad (2a)$$

$$T_2 = e^{\{-(\alpha_2 - j\beta_2)L\}}, \quad \ldots\ldots \quad (2b)$$

and $|e^{j2\theta}| = 1$, where

α_1, α_2 are attenuation coefficients (nepers per kilometre), β_1, β_2 are phase-change coefficients (radians per kilometre), θ is the canting angle in degrees, and L is the total distance (kilometres) through the rainfall region.

By re-arranging equations 1 and 2,

$$\text{XPD} = -20 \log\left\{\left(\frac{1 - T}{1 + T}\right)|e^{j2\theta}|\right\} \text{ decibels,} \quad \ldots \quad (3)$$

where

$$T = e^{\{-(\Delta\alpha - j\Delta\beta)L\}}, \quad \ldots\ldots \quad (4)$$

$$\Delta\alpha = \alpha_1 - \alpha_2,$$

$$\Delta\beta = \beta_1 - \beta_2.$$

In practice, not all the raindrops are equally oriented and the canting angles of individual raindrops will be scattered either side of some mean value. Components scattered by oppositely canted raindrops tend to cancel out, thus improving the XPD. This is taken into account in equation (1) by replacing the factor $|e^{j2\theta}|$ with $<e^{j2\theta}>$, where the brackets denote averaging over the whole distribution. Unlike the factor $|e^{j2\theta}|$, the averaged value has a value of less than unity; this results in a canting-angle reduction factor (ϵ), where

$$\epsilon = -20 \log < e^{j2\theta} > \text{ decibels.} \quad \ldots\ldots \quad (5)$$

The resultant expression for XPD becomes:

$$\text{XPD} = -20 \log\left\{\left(\frac{1 - T}{1 + T}\right) < e^{j2\theta} > \right\} \text{decibels} \quad \ldots \quad (6)$$

The reduction factor has been evaluated in a number of different experiments. In one case, it was evaluated by photographing raindrops and reading off the canting angles of individual drops, and estimating ϵ theoretically[19]. In other experiments, the reduction factor has been estimated by comparing measured results with theory; for example, in the comparison of simultaneously measured linear and circular depolarization[18]. The value of ϵ for circular polarizations appears not to be very sensitive to changes in frequency, and a value of about 7·5 dB at 4 GHz has been suggested[20].

For frequencies of 4 GHz and 6 GHz, $\Delta\beta$ is much larger than $\Delta\alpha$ and, therefore, T in equation (6) is imaginary. The magnitude of T is nearly unity for typical XPD ratios and therefore the phase relationship between co-polar and cross-polar circularly polarized signals is approximately $90 \pm 2\theta$ degrees. The significance of this result is that it is possible to determine how the mean canting-angle of the rainfall is varying, according to the changes observed in the relative phase measured. In situations where it is thought that the depolarization measured is attributable to sources other than rainfall, determination of the canting angle can assist in the identification of the depolarizing medium. This method has been used in several instances to identify ice-particle depolarization at higher frequencies[21, 22].

For linear polarizations, the relative orientation of the polarization plane and the canting angle determines the relative phase angle and, in this case, the relationship between the measured relative phase angle and changes in canting angle is more complicated[23].

THE ROLE OF MEASUREMENTS

The main incentive in developing theories and models for rain depolarization has been to enable predictions to be made of the cross-polar discrimination achievable over 4/6 GHz satellite links at earth stations in the INTELSAT system.

The following outlines the kind of methods used in a typical prediction process. Such a process usually starts with a measured or estimated statistical distribution of rainfall rate for the particular site in question. Experimental relationships can be derived between point rainfall rate and total attenuation A over a slant path, of the form

$$A = kR^{\alpha} \text{ decibels,} \quad \ldots\ldots \quad (7)$$

in which R is the rainfall rate in millimetres per hour, and k and α are constants. The constants, k and α, are dependent on a large number of highly variable factors, and are usually found by equating measured statistical distributions of rainfall rate and fade depth, using regression methods. Such simple relationships can hide a multitude of approximations, and therefore particular values of k and α can be regarded as being average values only. Experiments show that there is usually a good correlation between statistics of rainfall intensity and fade depth, although the instantaneous correlation between the 2 variables may be poor. A second relationship, between XPD and attenuation, can be derived theoretically[24], and takes the form

$$\text{XPD} = X - Y \log A \text{ decibels,} \quad \ldots\ldots (8)$$

in which X and Y are constants and A is the attenuation.

The instantaneous correlation between these 2 variables is likely to be rather better than that between point rainfall rate and path attenuation, and a relationship of this form can provide quite a close bound to rain depolarization measurements. From this relationship, it is possible to obtain an estimated statistical distribution of XPD from slant-path attenuation statistics. However, this is a gross simplification of the prediction problem, for the following reasons:

(*a*) the variability of point rainfall-rate statistics from year to year,

(*b*) the presence of depolarizing influences other than rainfall,

(*c*) the wide range of variability of the spatial characteristics of rainfall,

(*d*) the variability of rainfall with the topography of the earth station site and its surrounding area, and

(*e*) elevation-angle dependency.

All these factors, to a greater or lesser extent, call for fairly comprehensive measurements to be made which, experience has shown, may well reveal effects not considered prior to carrying out measurements. For example, measurements at 20 GHz of the depolarization on a slant path on the ATS–6 satellite revealed significant depolarization effects that were not attributable to rainfall.[21, 25] Attempts have since been made to develop a theoretical model for these effects, and it has been concluded that high-altitude ice particles could give rise to large differential phase shifts[26]. The measurements made by the BPO indicate that, at 20 GHz, ice-particle depolarization may be a more significant source of depolarization than rainfall.

Fairly extensive programmes of measurements have been conducted by COMSAT at a number of earth stations. A severe restriction on the usefulness of some of these measurements has been the poor residual XPD, due mainly to the ellipticity of the satellite aerial. Unless the component due to the satellite aerial can be eliminated, it is not possible to measure accurately rainfall depolarizations of similar magnitude to the residual XPD. Typically, residual clear-weather XPD values of 40–45 dB are required over the link to permit the rainfall component to be measured accurately, principally because XPD components add vectorially. It is possible to improve the residual XPD by using a more elaborate measurement system, and this has been done in some experiments. Table 1 gives a brief summary of 4 GHz depolarization experiments in progress, or completed.

TABLE 1

Results of Slant Path XPD Measurements taken at 4 GHz

Location	Elevation Angle	Clear-sky XPD (dB)	Test Period
COMSAT Laboratories, Clarksburg, USA	20°	24 37	1972–73 (1 year) 1974 to date, intermittent
Cairns, Australia	52°	18–25	1975 (4 months)
Taipei, Taiwan	20°		1975–76 (1 year)
Yamaguchi, Japan	8°	23·5	July 1975–July 1977
Ibaraki, Japan	35°		March 1976 (1 year)
Lario, Italy	16°		Planned October 1976, not yet started
Goonhilly, UK	28°	39–40	Commenced January 1977

Topographical features which are particularly likely to have a strong influence on the rainfall statistics at a given site are hills and valleys (particularly where the prevailing wind first traverses sea and then hills); this situation can result in heavy rainfall in a hilly region. This phenomenon is known as *orographic rainfall*, and can be responsible for significant localized effects[27].

The second UK earth station site at Madley is located eastwards of several mountainous areas in Wales, including the Brecon Beacons and the Black Mountains; it is thought that, in this location, a heavier rainfall may occur than at the site at Goonhilly. These factors all point towards the desirability of conducting depolarization measurements at the UK earth station sites, and therefore a programme of measurements of XPD and associated parameters has been embarked upon at both sites.

When making measurements of depolarization, it is highly desirable that simultaneous measurements of other variables should be made (for example, fading and rainfall rate) to identify the causes of the depolarization. However, at 4 GHz and 6 GHz, the fades accompanying depolarization have, in general, magnitudes of only fractions of a decibel. The fade depths at higher frequencies (11 GHz and above) are deeper, and the measurement of fading of a co-polar signal at these frequencies is therefore a more practical means of identifying rain depolarization. An alternative to the direct measurement of fade depth is the use of a radiometer, coupled to a small aerial pointing at the sky in the direction of interest. The radiometer measures the noise temperature of the sky, from which the corresponding fade depth can be directly estimated. Recent comparisons[13] of radiometer measurements with direct measurements via a satellite source at 20 GHz and 30 GHz[28] show that, despite doubts about the magnitude of scattering effects, it is possible to obtain good instantaneous correlation between directly-measured fade depth at a given frequency and that derived from a radiometer measurement of sky noise-temperature at the same frequency. Therefore, in situations where it is required to measure absorptive fading effects, the radiometer, properly calibrated, provides a low-cost alternative to satellite measurements.

At these higher frequencies, fairly accurate theoretical relationships have been developed and thus, assuming different bounding statistical distributions for the canting angles of individual drops within the rainfall, it is possible to derive upper and lower bounding curves relating XPD and fade depth.

It is vital, during the course of any propagation measurement, to ensure that observed effects can be attributed to propagation changes over the path being measured, rather than being due to measurement problems. Since the radiometer records atmospheric changes only, it provides a useful indicator that the measured effects are a result of atmospheric variations and not due to failure of the satellite source or problems elsewhere in the measurement chain. Additionally, radiometer data, accumulated over a long period, can provide a useful means of extrapolating depolarization measurements in time for a given site. It may also be practicable to use radiometer data to extrapolate measurements to other sites and other frequencies, although with a limited degree of confidence. Extrapolation of measurements by this means does not, as previously pointed out, cover depolarization due to causes other than rainfall; where evidence exists of such depolarizations they must be treated separately.

Accurate measurements of XPD could not be made by the BPO until a satellite source with a high degree of polarization purity became available (or alternatively, access to an earth station aerial with facilities for cancelling the residual satellite-ellipticity). The availability of an INTELSAT IV-A satellite in the Atlantic Ocean region during 1976, and the refitting of a 9 m experimental aerial at Goonhilly, enabled the BPO to commence 4 GHz XPD measurements in January 1977. To assist in the identification of the causes of depolarization, a 12 GHz Dicke-type radiometer was installed alongside the 9 m aerial, both aerials pointing towards the INTELSAT IV-A Flight 2 satellite. In addition to the measurements already mentioned, rainfall is recorded by means of a tilting-syphon rainfall gauge. Fig. 21 shows a block diagram of the experimental measurement arrangements.

This experiment is intended to provide the following information:

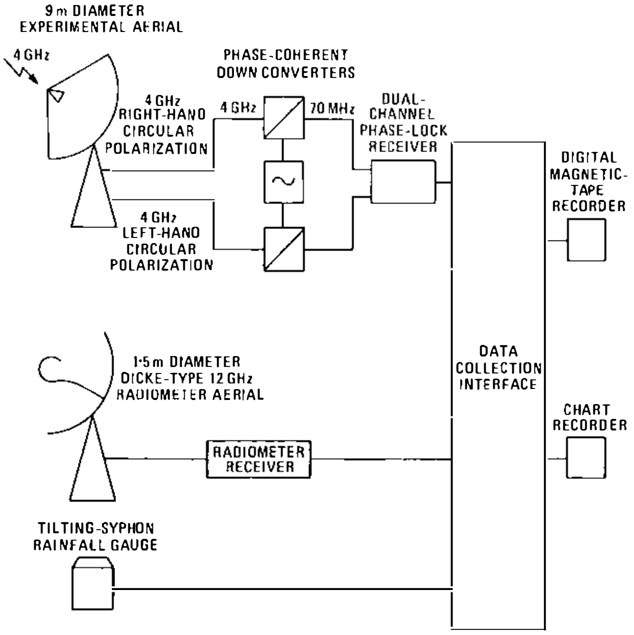


FIG. 21—Block diagram of measurement arrangement

(*a*) statistical distributions of XPD at 4 GHz, to provide an indication of the performance to be expected from a dual-polarized link using an earth station in the UK;

(*b*) statistical distributions of fade depth at 12 GHz, to provide data for the determination of aerial size for systems using frequencies of this order;

(*c*) indications of depolarization effects not already revealed by theoretical studies;

(*d*) the relationship between fading at 12 GHz and XPD at 4 GHz, to test the possibility of predicting XPD at 4 GHz at new sites from fade-depth data measured at 12 GHz at these sites;

(*e*) the statistical relationship (if any) between point rainfall rate, fade depth at 12 GHz and XPD at 4 GHz.

These measurements are providing useful results applicable to a 28° elevation-angle path at the Goonhilly site and may have some application to other sites. However, the first aerial system at the new earth-station at Madley will be used with an Indian Ocean satellite, and will, therefore, operate at a low elevation-angle (around 6°). This difference in elevation angle, together with the additional possibility of topographical influences on rainfall at Madley, indicates a need for a separate programme of measurements at 6° to be made at the Madley site.

To this end, a 12 GHz radiometer will be installed at Madley early in 1978 to enable measurements to be made of fade depths at 12 GHz. Later in 1978, with the introduction of Madley 1 to the Indian Ocean region, it is hoped to extend the experiment to the measurement of XPD using the Madley 1 aerial system, which has a dual-polarized feed with high polarization purity.

UK MEASUREMENT RESULTS

Since the start of these measurements, early in 1977, severe depolarizations have been observed on a number of occasions. Table 2 gives a list of the dates and times of the most significant of these events. It can be seen that low levels of XPD have occurred in conjunction with large fades measured with the radiometer on 2 of the occasions (on the 11th July and the 31st October), whereas on 2 other occasions the low levels of XPD were measured at times of relatively low depths of fading. These results indicate that, on the 2 occasions on which only low fades were observed, the primary source of depolarization was something other than rainfall, and was possibly ice particles at high altitudes. To investigate these occurrences further, scatter diagrams were produced for each of the events in Table 2, and are shown in Figs. 22(*a*)–22(*d*); the rain depolarization predicted by theory is also shown on these scatter diagrams.

TABLE 2

Severe Depolarization Events Observed at Goonhilly (21 January–31 October 1977)

Date	Minimum XPD (dB)	Accompanying Attenuation at 12 GHz (dB)	Time that XPD fell below 25 dB (s)
11–12 July 1977	22	15	192
23–24 August 1977	25	2·3	48*
5–6 October 1977	21·5	5·5	288
31 October 1977	19·9	21	156

* XPD recorded at 25·5 dB

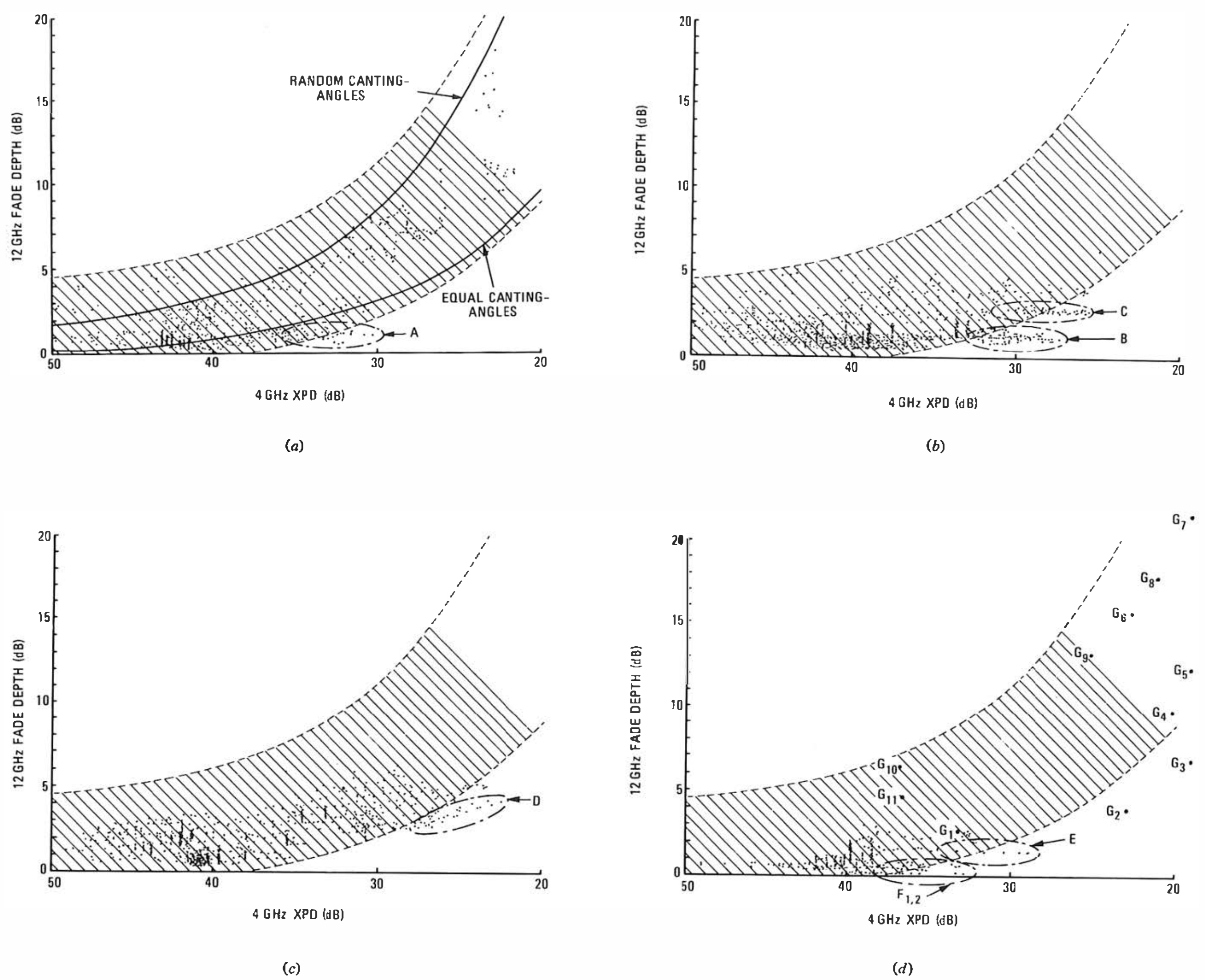
The 2 solid lines in Fig. 22(*a*) indicate theoretical relationships between XPD at 4 GHz and fade depth at 12 GHz for circular polarization over a slant path at 28° elevation angle; those are shown for equal rainfall rates assuming (*a*) equal canting angles, and (*b*) random canting angles, to represent 2 extremes of the same model. It is also necessary to consider the effect of the residual XPD, which can add vectorially to the rainfall or ice-particle depolarization; to account for this, maximum and minimum bounds were computed for the total XPD, allowing for the residual XPD at the start of each event. The shaded area on each diagram therefore represents the area in which there is correspondence between the data and the rain depolarization theory.

Referring to Fig. 22(*a*), it can be seen that the majority of points lie well within the shaded area. However, there is a cluster of points (marked A) parallel to the XPD axis, with a low accompanying fade (less than 1·5 dB). This effect is shown quite strongly in Fig. 22(*b*) where the clusters of points marked B and C clearly do not agree with the rain depolarization theory. The cluster marked C corresponds to a fade depth of just over 2 dB, indicating the presence of both rainfall and another depolarizing medium in the propagation path.

The results recorded in Fig. 22(*c*), while showing a cluster of points outside the shaded area are, in general, in agreement with the rain-depolarization theory. The cluster of points marked D do not (unlike those shown on Figs. 22(*a*) and (*b*)) move out parallel to the XPD axis, indicating that the main source of depolarization was probably rainfall.

The scatter diagram for the most severe depolarization event observed up to the beginning of November 1977 is shown in Fig. 22(*d*). The clusters of points marked E and $F_{1,2}$, show little correlation between XPD and fading for those periods. Fig. 23 shows extracts from the chart recordings over the same period and it can be seen from the signal traces that there is little correlation between the 2 measurements for the periods E and F. However, one stretch of recording (marked G in Fig. 23) showed a very strong correlation, and when plotted on the scatter diagram (points marked G_1–G_{11} on Fig. 22(*d*)), the observed values largely agree with the rain depolarization theory.

These observations appear to show that depolarizations of comparable magnitude to those attributable to rain can probably be caused by high altitude ice particles. Although



(a) 11–12 July 1977 (b) 23–24 August 1977
(c) 5–6 October 1977 (d) 31 October 1977

Notes: Sampling interval 30 s
The shaded area indicates the area in which rain depolarization theory is valid

FIG. 22—Scatter diagrams of fade depth and XPD measurements for given periods

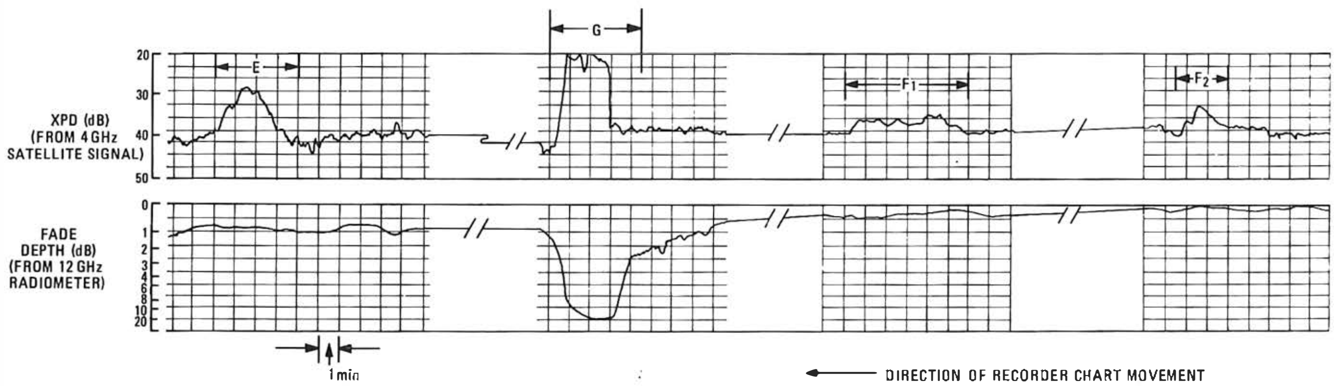


FIG. 23—Chart record of 4 GHz XPD and 12 GHz fade depth (recorded 31 October 1977)

similar effects have been observed elsewhere at higher frequencies, this is the first time that such observations have been made at 4 GHz or 6 GHz. Clearly, the presence of depolarization from causes other than rainfall significantly decreases the accuracy of the prediction of dual-polar performance based on the propagation models used until recently. Over the 10-month measurement period at Goonhilly, the XPD fell below 25 dB on 4 occasions, for a total of 11 min 24 s, representing approximately 0·003% of the total time. On 2 of these occasions, totalling 5 min 35 s, it was considered that the principal source of depolarization was other than rainfall, and it is concluded that non-rainfall depolarization is at least as significant as rainfall for XPD ratios corresponding to the smaller percentages of time.

SYSTEM PERFORMANCE REQUIREMENTS

Part 2 of this article mentioned the requirement for an overall clear-sky XPD of 27 dB or better for dual-polarized satellite links, but little has been said about the minimum acceptable XPD that can be tolerated for very small periods of time.

Because of the difficulty in predicting the likely cross-polar performance for satellite links, INTELSAT decided that a provisional lower limit would be set on the XPD of 12 dB, to be exceeded for at least 99·99% of time. This limit is intended to enable most of the earth stations in the INTELSAT system to operate to INTELSAT V without requiring adaptive polarization compensation.

From the results of the UK measurements, it appears certain that this requirement will be more than adequately met for Atlantic Ocean operation, without providing polarization compensation. However, the ability of the majority of earth stations to meet a more stringent limit without polarization compensation may result in a better performance being specified at some time in the future. It is therefore important that the BPO should have accurate information on the actual performance of satellite links both for Atlantic Ocean operation at 28° elevation angle, and also for Indian Ocean operation at 6° elevation angle.

A more stringent limit on XPD would have advantages in terms of satellite capacity since, with the existing limit, it will be necessary to introduce constraints in the allocation of carriers on opposite polarizations.

CONCLUSIONS

The introduction of dual-polarized frequency re-use systems into the INTELSAT satellite network requires a detailed knowledge of the problems arising from the propagation path. Until recently, only the effect of depolarization due to rainfall itself was considered, but there is now strong evidence that other causes are as significant, even at the relatively low frequencies of 4 GHz and 6 GHz.

The measurements carried out at Goonhilly have contributed to an increased understanding of depolarization effects attributable to the propagation path. It is intended to continue with the measurements at Goonhilly to obtain statistically significant data, and to proceed with low elevation-angle depolarization measurements at the Madley earth station. The first objective of the low-angle measurements will be to determine the need or otherwise for polarization compensation devices for low-angle operation. To supplement these measurements, the results of theoretical studies elsewhere will be used to try and improve methods of predicting cross-polar performance over slant paths. It is hoped that these methods will also be of value in predicting the performance of 11/14 GHz dual-polarization satellite links, with the accuracy of prediction improving as measured results of slant-path propagation at 11 GHz and 14 GHz become available.

ACKNOWLEDGEMENTS

The valuable work of the Telecommunications Development Department staff at Goonhilly in gathering the depolarization data, is acknowledged. Acknowledgements are also made to INTELSAT for the free use of the satellite and to COMSAT for the provision of a test carrier from the Andover earth station.

References

16 Marsh, A. L. Dual Polarization Technology, Part 2—The Introduction of Dual-Polarization Operation at UK Earth Stations. *POEEJ*, Vol. 70, p. 178, Oct. 1977.

17 Hogg, D. C., and Chu, T. S. The role of Rain in Satellite Communications. *Proc. IEEE*, Vol. 63, 9 Sept. 1975.

18 Chu, T. S. Rain Induced Cross-Polarization at Centimetre and Millimetre Wavelengths. *BSTJ*, Vol. 53, pp. 1557–1579, Oct. 1974.

19 Saunders, M. J. Cross-Polarization at 18 and 30 GHz due to Rain. *IEE Trans.* Antennas Propagation. Vol. AP—19, pp. 273–277, Mar. 1971.

20 Uzunoglu, N. K., Evans, B. G., and Holt, A. R. Scattering of EM Radiation by Precipitation Particles, and Propagation Characteristics of Terrestrial and Space Communication Systems. *Proc. IEE*, Vol. 124, p. 5, May 1977.

21 Howell, R. G. Cross-Polar Phase Variations at 20 and 30 GHz on a Satellite-Earth Path. *Electronics Letters*, No. 14, pp. 405–406, 1977.

22 McKewan, N. J. Phase of Cross-Polarized Signals on Microwave Satellite Links. *Electronics Letters*, 16, pp. 489–491, 1977.

23 Thirlwell, J., Knox, D. M., and Howell, R. G. Cross-Polar Variations on a Satellite Earth Path. *Electronics Letters*, No. 17, pp. 516–518, 1977.

24 CCIR (1974–1978) Doc. 5/206 (Canada) Techniques for the Prediction of Rain Depolarization Statistics at Microwave and Millimetre Wavelengths, May 1977.

25 Shutie, P. F., Allnut, J. E., and McKenzie, E. C. 30 GHz Satellite—Earth Signal Depolarization in the absence of Significant Fading. *Electronics Letters*, No. 13, pp. 1–2, 1977.

26 Evans, B. G., and Holt, A. R. Scattering Amplitudes and Cross-Polarization of Ice Particles. *Electronics Letters*, No. 12, pp. 342–344, 1977.

27 Allnutt, J. E. Slant Path Attenuation and Space Diversity Results Using 11·6 GHz Radiometers. *Electronic Letters*, No. 12, pp. 346–347, 1977.

28 Bell, R. R. The Calibration of 20 and 30 GHz Radiometers using the ATS-6 Satellite. URSI Symposium 'Propagation in Non-Ionized Media' La Baule 1977.

Book Review

Microwave System Engineering Principles. S. J. Raff, PH.D. Pergamon Press. xi + 120 pp. 17 ills. £3·90.

The author based this book on his graduate courses on microwave engineering principles. He has made the subject interesting by describing, in simple terms, the physical principle behind the concepts treated. His treatment of the application of the second law of thermodynamics and the properties of thermal noise to certain circuit theorems is particularly interesting.

Subjects treated are: thermal noise, statistics, signal processing and detection, antennae, propagation and transmission lines, reflection and refraction and some system characteristics. However, perhaps not surprising for a book of this size, the treatment is at an elementary level only, and a microwave engineer would need to build on this information. The book would be suitable for someone requiring an introduction to the subject, or for someone wishing to know something about the physical aspects of the concepts involved.

In the opinion of the reviewer, the print is not attractive to read and the printing used on the figures is too small.

M. C. D.

The Selection of Microprocessors

M. D. ROWE, B.Sc.†

UDC 681.31–181

A simple classification of microprocessors into 3 types, supported by specification details of representative microprocessors, is used in this article to aid a discussion of the technical factors relevant to the selection of microprocessors. The weaknesses of commonly used methods of selection and the commercial constraints for telecommunications applications are used to show the difficulty of the task. The need for a continuing supply of components of assured reliability is highlighted to stress the importance of the choice of semiconductor manufacturer.

INTRODUCTION

Articles[1,2] already published in this *Journal* have given an introduction to microprocessors by showing how their design could be built up from the principles of programmable logic, and described the techniques used in writing, and in eliminating errors from, programs for microprocessors. This article discusses the selection of a microprocessor from the many available, particularly when a large range of applications is to be satisfied.

In this article, the microprocessor field is divided into 3 distinct classes, and the broad characteristics of each class are described. The wide meaning of the term *performance* when applied to microprocessors, and the technical factors which a user would consider, are discussed. Survey tables giving technical details of some modern microprocessors from each class are presented as a convenient reference for the discussions. The commercial influences on the selection of a microprocessor are introduced and used to put the technical considerations into perspective.

Since 1971, when only 2 designs of microprocessor were available, each year has brought an increasing number of new designs; today, more than 50 distinct designs of microprocessor are on sale. The choice seems bewildering at first sight, but fortunately, as the number of designs has increased, so has their diversity. For any particular application, a choice of microprocessor need only be made from perhaps as few as a dozen broadly comparable designs. In this article, the classification scheme and discussion of the technical factors, and the survey tables (Tables 1–4), are intended to show up the distinguishing features.

The microprocessor is a common block of logic which can be adapted to any of a wide variety of tasks by a specific program. This ability allows some of the economies of semiconductor large-scale integration (LSI) to be brought to tasks where production numbers are only moderate[3]. The microprocessor therefore answers a special need in the field of telecommunications system engineering, which is characterized by many types and variants of equipment in numbers that are small by semiconductor standards.

In other respects, semiconductors and telecommunications are less well matched. Telecommunications, with its lengthy design cycles and long equipment lifetimes, contrasts starkly with the volatile semiconductor industry, which has seen 20 or more designs of microprocessor stillborn or extinct. Most of the present designs have been on sale for less than 2 years. The difficulty of maintaining equipment in 30 years' time in an era of rapid technological obsolescence is not merely a philosophical problem. It will be taken up later but is mentioned here as a hint that a satisfactory selection cannot be made on technical considerations alone.

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MICROPROCESSOR PERFORMANCE

If their speed of operation allows, most microprocessors, by virtue of their inherent versatility, can accomplish most application requirements, though with varying ease and cost. Each microprocessor has some strengths and some weaknesses when viewed in the light of typical applications. In specific applications, one microprocessor's strong point may prove worthless, while another's weakness may be unimportant.

There are many aspects of microprocessor performance that need to be considered in the selection process:

Speed of Operation
Speed of operation is a primary measure of performance since it determines whether a given application can be tackled by a particular microprocessor. Speed of operation is hard to predict, since it depends upon so many other factors, such as clock rate, timing scheme, word size, addressing modes, architecture and instruction set. These factors are discussed later in this article.

Program Space
Program space is the number of program binary digits required to perform a given task. The program space requirement is a complex combination of many factors, but its chief importance lies in the influence it has on the number of memory packages required in the system, and hence upon the unit production cost of the equipment.

Chip Count
The chip count is the number of major devices required to make a working system. The chip count has a direct effect on the production cost of the equipment, and depends strongly on the architecture of the microprocessor.

Cost
Another part of performance is the production cost of the equipment. This depends upon the chip count and, to a lesser extent, on the price of the chips. That is not the whole story, however, for the development and operational costs cannot be ignored. These additional costs are discussed later.

Ease of Use
Ease of use is a less obvious facet of performance. If a device is easier to use, it is likely to take less time to understand and to program, thereby reducing the cost of training and of writing the program. The programmer is also likely to make fewer mistakes. The quality of the manufacturer's literature

and development aids may be as influential as the instruction set and architecture of the device.

Versatility

Versatility could also be considered an aspect of performance. The more versatile a microprocessor is, the wider the variety of tasks which it is likely to be able to tackle, and hence the more scope there is for sharing the training and development costs by using the same device for several applications.

The point in mentioning these progressively more obscure aspects of performance is to encourage consideration of the measurable parameters for their significance in the overall selection process.

CLASSIFICATION OF MICROPROCESSORS

The microprocessor field can be divided into 3 classes within which the devices are intended broadly for the same levels of performance. The survey tables (Tables 1–4) follow the same classification. The general characteristics of each class are described first, followed by an introduction to the survey tables.

Compact Microprocessors

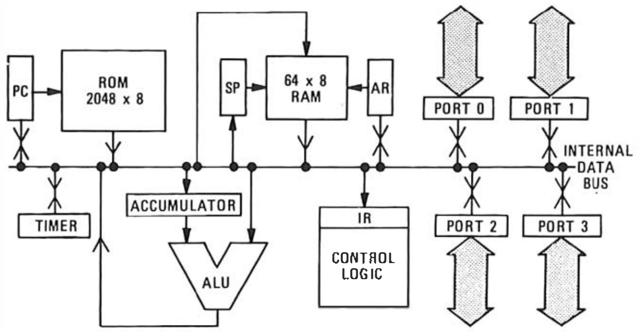
Compact microprocessors are intended for applications for which low cost and simplicity in the external circuitry are more important than speed of operation or sophistication of facilities. Architectures are used which result in a minimum chip count for small systems, sometimes at a penalty for large systems. Some have a program read-only memory (ROM), data memory, input/output ports and system clock all on the central processing unit (CPU) chip; that is, a complete microprocessor system (or microcomputer) on a chip. A typical internal architecture of a compact microprocessor is shown in Fig. 1(a); a typical minimum system structure is shown in Fig. 1(b). The instruction set need only be fairly basic, with an emphasis more on compactness in the program than on speed of operation. The programs tend to be short (typically 4 kword or less) and are usually written in assembly language.

Applications are in wired-logic replacement or dedicated control. Many applications are new to electronics (for example, use in cash registers) or new altogether (for example, games played on domestic television receivers). Production volumes tend to be high, so that production costs predominate. A single power supply and on-chip clock generator are likely to be more highly regarded than sophisticated features or long-term availability. Word widths are commonly 4 bit or 8 bit. Compact microprocessors are constructed almost exclusively in metal-oxide semiconductor (MOS) technologies[4] (see Fig. 2).

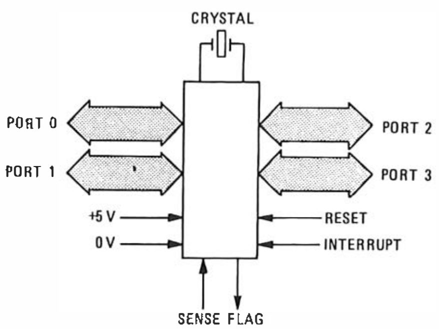
Versatile Microprocessors

In versatile microprocessors, architectures are used which give a wide addressing range and easy control of larger systems. Typical internal architecture and minimum system structure are given in Figs. 3(*a*) and 3(*b*) respectively. Maximum performance, within the constraints of accommodating the CPU on a single chip, is sought by designers, though chip count in the system as a whole may be less important. Sophisticated system facilities are normally provided for interrupt and direct memory access. The parts family often includes special chips for controlling large memory systems constructed from industry-standard memory devices. The instruction sets tend to be extensive, sometimes with multiplication and division facilities, with a choice of several modes of addressing. Larger programs are feasible (even 32 kword or more), so that a high-level language is usually offered (by means of a compiler). Versatile microprocessors are often powerful enough to form the basis of their own development system by running the development software, such as assemblers and compilers, as resident programs[2].

The applications of versatile microprocessors are often those for which a conventional minicomputer would have been used (but for cost) before the advent of the microprocessor. Indeed, some of the microprocessors use the instruction sets of successful minicomputers. Word lengths are generally 8 bit or 16 bit; a few 12 bit devices are available. Most versatile microprocessors are produced in MOS technologies, mainly n-channel; some are produced in complementary metal-oxide semiconductor (CMOS) technology and some in bipolar technology (see Fig. 4).



(*a*) Typical internal architecture



(*b*) Typical minimum system structure

SP: Stack pointer
AR: Address register
ALU: Arithmetic logic unit
PC: Program Counter
IR: Instruction Register

FIG. 1—Compact microprocessor

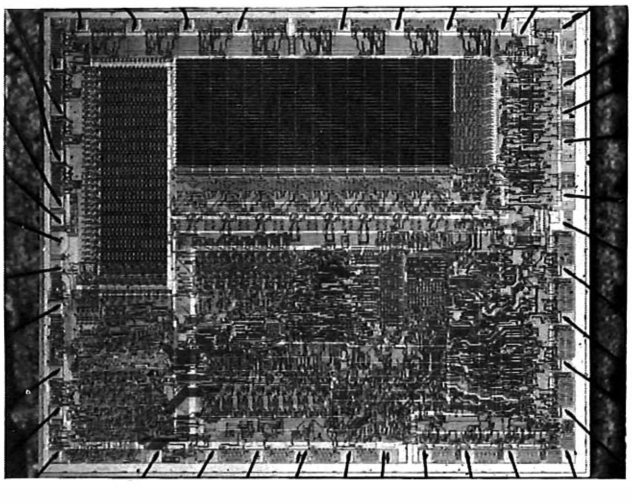
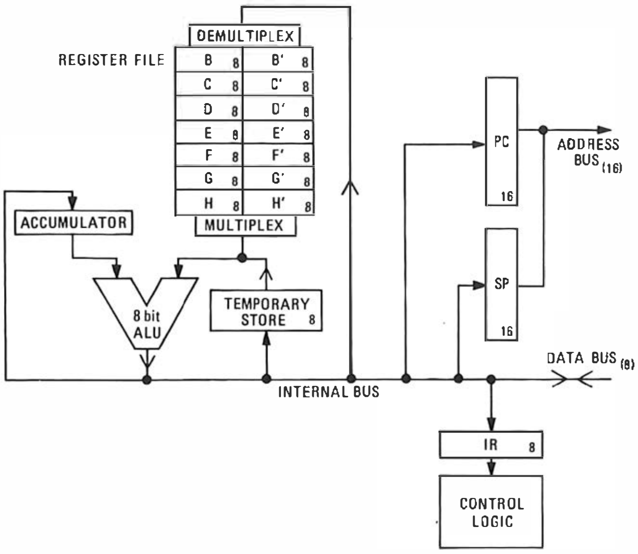
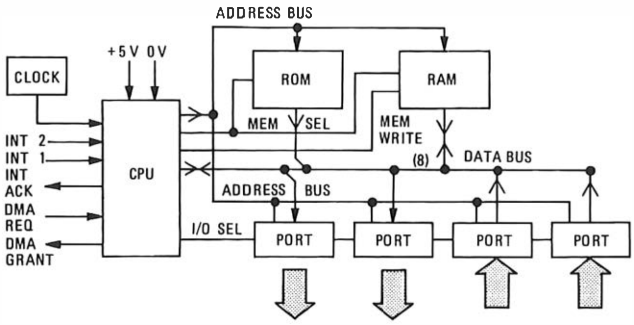


FIG. 2—A self-contained microprocessor with on-chip ROM

(Photograph by courtesy of Intel Corporation UK Limited)



(*a*) Typical internal architecture



(*b*) Typical minimum system structure

INT: Interrupt
ACK: Acknowledge
REQ: Request
SEL: Select
DMA: Direct memory access
I/O: Input/output
MEM: Memory

FIG. 3—An 8 bit versatile microprocessor

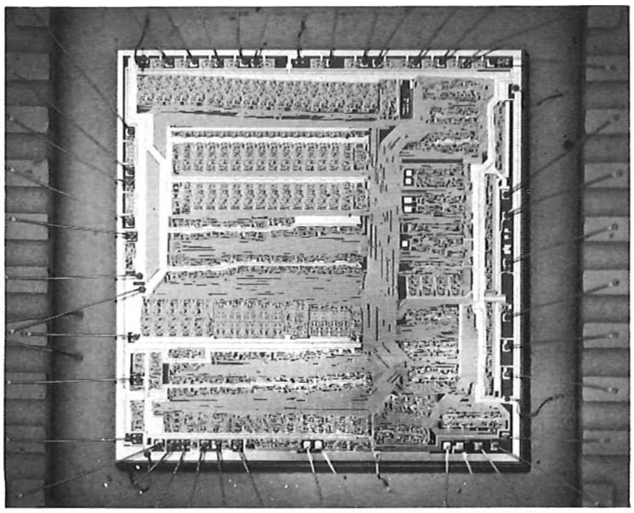


FIG. 4—A 16 bit bipolar microprocessor

(Photograph by courtesy of Ferranti Limited)

High-Performance Microprocessors

The highest performance in terms of processing speed is demanded from these devices. Historically, this has dictated the use of bipolar technologies which, with their lower packing density, led to the use of the bit-slicing technique mentioned in an earlier article[2]. These bit-slice devices, in addition to their high intrinsic speed, permit also architectural flexibility, in that the physical paths of the system can be fitted to the application. Microprogramming by the user, which allows the instruction set to be tailored to the application with beneficial effects on the performance, is also usually offered.

The systems which use these devices place a premium on performance at the expense of chip count, power dissipation and development costs. They are often used for functions previously carried out with medium-scale logic parts. Typical applications are in making high-performance minicomputers, high-speed control or real-time information-processing, such as speech analysis.

SURVEY TABLES

The survey tables (Tables 1–4) give the author's interpretation of manufacturers' data for a small selection of modern microprocessors. Table 1 shows data for the compact types of microprocessor. Tables 2 and 3 refer to examples of the versatile microprocessor types (Table 2 for 8 bit devices and Table 3 for 12 bit and 16 bit devices). Table 4 includes data for the high-performance types of microprocessor.

The information in the tables has purposely been restricted in scope because the tables are not intended as a source of data but only to illustrate points made in this article. Manufacturers' own data should always be consulted as the only authoritative source for working decisions.

The data given for the compact and versatile microprocessors (Tables 1–3) covers, roughly, resources on chip, software characteristics, performance indications and hardware aspects.

The data given for the high-performance bit-slice microprocessors (Table 4) is much simpler, since many of the characteristics which are fixed by the chip designer in the other types are at the hands of the user of the high-performance devices. It is not intended to add much to the information given in an earlier previous article[2], which was sufficient to give some meaning to the parameters presented; further description of these devices is available[5].

SOFTWARE DESIGN

Word Size

The word size of the instruction and data paths has some influence on the ease of use and efficiency of a microprocessor but, in general, is one of the least important parameters. Short-word machines (for example, 4 bit) generally need instruction words longer than the data words, and address manipulation is particularly difficult. As explained in an earlier article[1], an address of 12–16 bit is needed for memory sizes of 4 kword to 64 kword. In principle, therefore, 16 bit devices ought to be more straightforward to program. In practice, the limitations of package pins and chip layout, and a preoccupation with program size, has made most 16 bit microprocessors less easy to use than 8 bit versions. The 8 bit machines generally show a slight advantage in program size, except where 16 bit precision is needed.

Address Capacity

The addressing capacity indicates the size of program which can be used without resorting to special techniques. If the range is too small, extra hardware will be required to extend the addressing, whereas spare capacity results in superfluous address binary digits being carried in instructions which refer to the memory. Most general-purpose microprocessors give an addressing range of 32–64 kwords, but some ROM-on-

TABLE 1

Sample Data for a Selection of Compact Microprocessors

Data	Device A (8 bit, nMOS)	Device B (8 bit, nMOS)	Device C (8 bit, nMOS)	Device D (4 bit, pMOS)
RESOURCES ON CHIP				
Number of working registers	RAM	8 (2 banks)	16	2
RAM on-chip/maximum	32, 8 bit/[3]	64, 8 bit/320	64, 8 bit/64 kword	128, 4 bit/[5]
Stack size	2	8	1 (expandable)	2
Timer/event counter	8 bit event	8 bit	8 bit	No
Interrupts	No	1	1	No
PROGRAMMING CHARACTERISTICS				
ROM on-chip/maximum	512, 12 bit/[3]	1 kword 8 bit/ 4 kword[4]	2 kword 8 bit/ 64 kword	2 kword 8 bit/[5]
Instruction word size	12 bit	8 bit × 1, 2	8 bit × 1, 2	8 bit
Main operand addressing modes	DIR, REG-IND	REG, REG-IND	REG, REG-IND	REG-IND
Full-range and relative jumps[1]	Full-range	Full-range	Both	Both
Number of conditional jumps	3	14	9	30
PERFORMANCE INDICATIONS				
Add register to accumulator	4 μs	2·5 μs	1 μs	12·5 μs
Full-range AND-mask in memory[1], [2]	8 μs	10 μs	6·5 μs	37·5 μs
Local increment, test, and branch[2]	28 μs	10 μs	5·5 μs	112·5 μs
Bit set and test instructions	Yes	Yes	No	Yes
Decimal arithmetic instructions	No	Yes	Yes	Yes
HARDWARE ASPECTS				
Power supplies	5 V	5 V	5 V	− 15 V[6]
Clock requirements	On-chip 1 MHz	On-chip 6 MHz	On-chip 4 MHz	On-chip 100 kHz
I/O on-chip/maximum	32/[3]	27/No limit	32/2048	31/[5]
Number as input/as output	32/32	24/24	32/32	31/18
Separate I/O and memory expansion	No[3]	Yes	Yes	No[5]
Notes (1) Full-range in minimum system only (2) Author's solutions—see text (See note in Table 4)	(3) Not expandable (Slave microprocessor version also made)	(4) ROM or PROM on-chip (Slave microprocessor version also made)		(5) Not expandable (6) +5 V, −10 V for TTL compatibility (Many versions with various ROM/RAM and I/O options)

REG: Register DIR: Direct REG-IND: Register-indirect nMOS: n-channel MOS pMOS: p-channel MOS TTL: transistor-transistor logic

TABLE 2

Sample Data for a Selection of 8 bit Versatile Microprocessors

Data	Device A (nMOS)	Device B (CMOS)	Device C (nMOS)	Device D (nMOS)
RESOURCES ON CHIP				
Number of accumulators	2	1	1	1
Number of working registers	0	16, 16 bit including program counter defined by pointers	3, 8 bit (2 banks)	6, 8 bit (2 banks)
Number of address registers[1]	2, 16 bit (Index)		3, 8 bit of above	4, 16 bit of above
Size of stack or pointer (SP)	16 bit SP		8 × 16 bit stack	16 bit SP
Number of interrupts	2 (1 maskable)	1	1 (maskable)	2 (1 maskable)
PROGRAMMING CHARACTERISTICS				
Address range	64 kword	64 kword	8 kword[4]	64 kword
Instruction word size	8 bit × 1, 2, 3	8 bit × 1, 2, 3	8 bit × 1, 2, 3	8 bit × 1, 2, 3, 4
Main operand address modes	DIR, Indexed	REG-IND	REG, DIR, Indexed, Page 0	REG, REG-IND, Indexed
Full-range and relative jumps	Both	Both	Both	Both
Number of conditional jumps	14 relative	18	8	8 full-range
PERFORMANCE INDICATIONS				
Add register to accumulator	1·5 μs	2·5 μs[3]	2·4 μs	1 μs
Full-range AND-mask in memory[2]	5 μs	17·5 μs[3]	12 μs	8·25 μs
Local increment, test, branch[2]	4·5 μs	7·5 μs[3]	9·6 μs	6·25 μs
Bit set and test instructions	Yes	No	Yes	Yes
Decimal arithmetic instructions	Yes	No	Yes	Yes
HARDWARE ASPECTS				
Power supplies	5 V	3–12 V[3]	5 V	5 V
Clock requirements	2 phases	On-chip	1 phase TTL	1 phase TTL
System and standard memory	Both	Standard	Standard	Standard
Memory access time	300 ns	300 ns	500 ns	230 ns
Architecture	Single-address	Register-based	Register-based	Register-based
Notes (1) Not including program counter or stack pointer (2) Author's solutions—see text (See note in Table 4)		(3) Times at 10 V; TTL compatible at 5 V	(4) Paged to 32 kword	

REG: Register DIR: Direct REG-IND: Register-indirect Page 0: Page addressing (zero assumed for high-order address)

TABLE 3

Sample Data for a Selection of 12 and 16 bit Versatile Microprocessors

Data	Device A (bipolar)	Device B (nMOS)	Device C (CMOS) (12 bit)	Device D (nMOS)
RESOURCES ON CHIP				
Number of accumulators	1	6, 16 bit	1	16, 16 bit in RAM(7)
Number of working registers	0		1, 12 bit	
Number of address registers(1)	0		RAM(5)	
Size of stack or pointer (SP)	16 bit SP	16 bit SP	Not applicable(6)	16 bit WP(7)
Number of interrupts	1 (mask, vector)	2 (mask, vector)	2 (1 masked)	16
PROGRAMMING CHARACTERISTICS				
Address range	32 kword	64 kword	4 kword	32 kword(8)
Instruction word size	16 bit × 1, 2	10 bit(4) × 1, 2, 3, 4	12 bit × 1	16 bit × 1, 2
Main operand address modes	DIR, REG-IND, Page 0	REG-REG, DIR	Page 0, Indexed	REG-REG, DIR, Indexed, REG-IND
Full-range and relative jumps	Full-range(3)	Both	Relative	Relative
Number of conditional jumps	16+	16 (+ 16 external)	14	12
PERFORMANCE INDICATIONS				
Add register to accumulator	3·4 μs	2·4 μs	2·5 μs(5)	4·7 μs
Full-range AND-mask in memory(2)	11·8 μs	16·4 (13·2)(4) μs	14 μs(5)	10 μs
Local increment, test, branch(2)	17·7 μs	10·8 μs	16·3 μs(5)	11·3 μs
Bit test and set instructions	Yes	Yes	No	No
Decimal arithmetic instructions	No	No	No	No
HARDWARE ASPECTS				
Power supplies	5 V	12 V, 5 V, −3 V	4-11 V(5)	12 V, 5 V, −5V
Clock requirements	1 phase 10 MHz	2 phase 5 MHz	On-chip 4 MHz(5)	4 phases 3 MHz
System and standard memory	Standard	Standard	Both	Both
Memory access time	100 ns	700 ns	250 ns(5)	300 ns
Architecture	Single-address	Register-register	Single-address	Memory-memory
Notes (1) Not including program counter or stack pointer (2) Author's solutions—see text (See note in Table 4)	(3) Also abbreviated, jumps to page 0	(4) Faster time for 16 bit memory	(5) Times at 10 V; TTL compatible at 5 V (6) Address modification in sub-routine	(7) WP defines working registers in RAM; stack and SP in RAM. (8) Byte addressing (Bipolar version also made)

DIR: Direct REG-REG: Register to register REG-IND: Register-indirect

TABLE 4

Sample Data for a Selection of High-Performance Microprocessors

Data	Device A (Shottky TTL)	Device B (Shottky TTL)	Device C (Shottky TTL)	Device D (ECL)
ALU SLICE				
Word size	4 bit	4 bit	2 bit	4 bit
Number of instructions	64	64	40	Over 100
Number of general-purpose registers	16	8	11	0
Maximum clock rate	10 MHz	10 MHz	10 MHz	20 MHz
SEQUENCER				
Number of address binary digits	4	4	9	4
Number of commands	12	4	11	16
Stack size	4 × 4 bit	16 × 4 bit	0	4 × 4 bit
Maximum clock rate	10 MHz	10 MHz	Over 10 MHz	20 MHz
Power supplies	5 V	5 V	5 V	−2 V, −5·2 V
Note: The values in this table represent the author's interpretation of manufacturers' data, and are for illustration only. The manufacturers' literature should always be consulted for working information				

chip devices cannot be extended beyond the memory which is on the chip.

Processing Speed

The faster the microprocessor executes each instruction the higher the processing rate. For any particular machine, doubling the clock rate could double the processing speed. The clock rate is not, however, a good guide when comparing different microprocessors, since designs differ in the way the execution cycle is subdivided into clock periods. Inspection of Tables 1–3 shows the varying relationships between the clock rate and the register-accumulator addition times which are given. This addition time is often used for comparison, but can also be misleading, since the ratio of the execution time of one instruction to that of another varies from one microprocessor to another. This is indicated in Tables 1–3, which include the author's attempt at deriving the execution times for common tasks which use other instructions. The effective processing speed depends upon the mixture of instructions presented to the microprocessor.

Instruction Set

The performance of a microprocessor also depends upon the amount of useful work carried out by each instruction. This is affected by many architectural details, such as whether the operands† can be drawn from locations anywhere in the memory, or whether one must be in a local register and the other in the accumulator. In the latter case, the data may have to be moved from the main memory first, and returned there afterwards. Some microprocessors implement space-saving compound instructions, useful in common tasks, such as INCREMENT-AND-SKIP-IF-ZERO, used for loops, Others, conversely, omit a basic instruction, requiring extra program steps to circumvent the problem.

Tables 1–3 show 3 execution times: one is the register-accumulator addition time, or its nearest equivalent where there are no local registers; the others are short routines which are commonly used, as attempted by the author. One of these short routines is the time required to modify data in a location anywhere in the main memory; for example, by executing an AND function with fixed data. (Devices without direct addressing, or at least a direct load/store instruction, fare badly on such a test.) The other routine is incrementing or decrementing a local register, comparing it with an arbitrary value (not zero) and jumping a short distance conditionally upon the result. The tests are not extensive enough for evaluation, but show the variety of results obtained from different devices: this is largely a result of the addressing schemes used. The tests also show how a given microprocessor varies in its ability to do different tasks, and should serve as a warning against placing too much faith in tests unrelated to the job at hand.

The number of instructions said to be possessed by a microprocessor is not a good guide to the power of the instruction set. So much depends upon the way in which the instructions are counted that no significance can be attached to this information, which is therefore not given in the survey tables. A microprocessor with a contorted architecture may need so many special instructions to get data past bottlenecks that the number and power of the remaining instructions is quite low. The survey tables indicate the provision of specific binary digit manipulation and decimal arithmetic instructions, and the number of conditional jump instructions.

Addressing Schemes

Several basic forms of addressing were defined in an earlier article[1], and these are taken a little further in Table 5 (and other literature[6]). The limitations of instruction codes and the need for efficiency in short-word machines dictate that most microprocessors offer some, but not all possible, modes. Many offer a poor selection of modes which are not uniformly implemented for all combinations of operand and operation. In some microprocessors, the modes are named inappropriately, giving a false impression of the power of the instruction set. For example, the "auto-indexed register-indirect" mode is called "indexed" by one manufacturer. Unfortunately, for any microprocessor, there is no easy way of analysing the instruction set and addressing modes except by examining them in detail, though a gallant attempt has been made by some workers in the field to do so graphically[7]. In Tables 1–3, the information given on the main modes of operand addressing represents the author's opinion of the modes which are implemented in a reasonably regular way for a high proportion of the arithmetic and logic operations.

A useful choice of addressing modes should allow a balance to be struck between the range of locations from which the operand may be drawn at random and the number of address binary digits carried in the instruction. The abbreviated modes allow some range to be traded for reduced space by taking advantage of the fact that most references to the memory are local rather than global in scope. Microprocessors which offer register-indirect addressing and little else are handicapped for general tasks. Tables 1–3 show how the choice of addressing modes affects the execution times. Whether an increase in the program size and execution time is important depends upon the effect these have on the chip count and development time, and hence, ultimately, on cost.

Also shown in Tables 1–3 are the addressing modes provided for JUMP and similar operations. A full-range jump, which can transfer the program to anywhere in the memory, and relative jumps, which have a restricted range but shorter instructions, are both useful. Also shown is the depth of the subroutine or interrupt stack; that is, the degree to which subroutines may be nested. An on-chip stack may impose a definite limit, whereas a stack in the main memory requires that an external random-access memory (RAM) be used. The tables show the size of the stack pointer where this method is used.

TABLE 5
Common Modes of Addressing

Addressing Mode	Remarks
Direct	Full explicit address in memory given in instruction
Register	Instruction carries short address of local register
Implied	Instruction has no address part: operation tied to particular source or destination
Immediate	Data is in location immediately following instruction.
Indirect	Location indicated contains address of data instead of data itself. Other mode (including indirect) may be used to indicate the indirect location
Register-indirect	Address of the data is contained in a pointer-register
Indexed	Effective address formed by adding address in instruction to address in index register. Addresses may be abbreviated
Auto-indexed	Pointer or index register incremented or decremented automatically before or after use
Page	Partial address in instruction laminated with partial address in page-address register (PAR) to form effective address. PAR usually contains high-order address
Page zero/ current page	As for page addressing, but zero assumed for high-order address (Page 0) or upper part of program counter used (current page)
Abbreviated	Addresses used to form effective address are not full-length
PC-relative	Address in instruction added to address in program counter: used for relative jumps

Note: These are not rigorous definitions, but act as a guide to microprocessor practice; there are many compound forms which defy description

† An operand is a quantity on which an operation is to be performed

Register Complement

Many microprocessors provide several general-purpose registers on the CPU chip, or privileged access to small sections of the main memory. These are the most precious resources which the programmer can allocate. The instructions which use them are generally more compact, execute faster, and can be used for a greater variety of operations than those which use the main memory. Their usefulness depends upon the extent to which they can receive results from the arithmetic unit, as well as supplying operands to it. Arithmetic registers (or accumulators) can do both, and are shown separately in the survey tables from the general-purpose (or working) registers, which generally only supply operands.

Some microprocessors have no hierarchy of storage, and treat all memory and input/output resources alike in a single address range. If the bus structure is well-designed, this can make the device easier to program without apparent reduction in the performance. For the versatile microprocessors, the basic architecture is classified as register-based or single-address in Tables 2 and 3. There arc variations for those microprocessors which can carry out operations between registers or between memory locations.

The speed of operation and program efficiency advantages tend to be lost if so many registers are provided that more than a few binary digits in the instruction are needed to address them. Indirect addressing may then have to be provided, possibly without more-direct modes. The incentive for including these larger RAM stores is a reduction in the chip count if no other RAM storage is needed. The larger RAM stores arc thus common in the compact microprocessors, and are listed separately from working registers in Table 1 if they arc less accessible (that is, have less-direct addressing modes or fewer operations) than the working registers. The compact microprocessors, in particular, often have completely separate address ranges and instructions for program memory, external RAM, internal RAM, working registers and input/output ports.

Registers (in addition to the program counter and stack pointer) arc often provided which can accept and manipulate address information as well as, or instead of, data. They usually work as pointers for register-indirect addressing, or as index registers. Details of these registers are included in the survey tables if they can be used to affect the address accessed in the main memory during a storage addressing cycle.

HARDWARE DESIGN

Power Supplies

Some microprocessors require 2 or 3 different voltage rails, which of course adds to the cost of using them. The present trend is towards the use of a single power rail, often 5 V.

Microprocessors implemented in CMOS technology have very small power dissipations, and are tolerant of variations in the supply voltage, which makes them suitable for operation on batteries. They also have excellent immunity to electrical noise.

Clock Requirements

Many microprocessors require external clock signals, sometimes at high power levels and in several phases, with critical specifications for the overlap of one phase with another. Most manufacturers also supply clock generators designed specifically for their microprocessor, though these clock generators can be costly. The trend is towards having the clock generators on-chip, requiring the provision externally of only a quartz crystal or a resistor-and-capacitor network.

Input/Output Ports

The compact microprocessors usually have direct input/output (I/O) ports on-chip, which have latches to maintain the states of the output lines; some devices also offer individual I/O lines. Table 1 indicates the number of I/O lines on the minimum system, and the maximum number of these which may be used respectively as input or output ports. Special chips, combining memory and I/O ports, may also be available for compact and versatile microprocessors; otherwise, plain latch circuits and buffers are usually used for I/O ports.

A few microprocessors also have full word-serial I/O, backed up by a shift-register accessed in parallel within the microprocessor.

Package Size

Most microprocessor parts are available in standard 40-pin packages; some can be obtained, with reduced capabilities, in smaller packages. Some manufacturers use special packages which offer particular advantages, such as quad-in-line, which arc of small size without sacrificing the number of pins. Ceramic packages are preferred to plastic packages for most telecommunications applications.

Memory Speed

The access time required of the program memory for a processor to run at full speed is shown in Tables 2 and 3. However, most microprocessors are designed so that they can be used with slower memories, if required. The performance of modern MOS memory components is such that only the bipolar microprocessors arc likely to benefit from memory devices faster than standard MOS memories.

SYSTEM CONSIDERATIONS

Parts Family

Many manufacturers offer a whole range of special parts to go with their microprocessors. Use of such parts can considerably reduce the number of components needed to implement a working system. Some microprocessors require the use of special memory parts because of the design of their bus structure. Some manufacturers of microprocessors that will work with standard parts also make special combination parts (for example, ROM with I/O ports) to reduce the chip count of smaller systems.

Interrupts

Interrupt facilities are normally needed if a microprocessor is required to give a quick response to external events which are not under its control, particularly if there are many possible sources. The design of the interrupt structure, and of a program which successfully copes with interrupts, is a difficult task, and beyond the scope of discussion here. The new address (vector) may be provided as a fixed value internally (usually to a store location which contains a jump to a service routine) or, on some microprocessors, by an external device. Sometimes, multiple levels of interrupt arc implemented on the CPU chip, where an interrupt of a higher level can interrupt one of a lower level (or *priority*), but not one of a higher level. Often the ability to mask out some levels is provided; the extent of the interrupt facility is indicated in Tables 2 and 3.

Other Facilities

Other microprocessor features that might be required include: direct memory access (that is, the ability to disconnect the microprocessor from the bus); microprogramming; advanced arithmetic modes (for example, multiplication and division); programmable timers; serial I/O; low power consumption; and wide temperature range. The facilities required depend upon the user application, but the facilities offered depend largely on where a manufacturer intends his microprocessor to be used.

SELECTING A MICROPROCESSOR

It is appropriate now to consider how a microprocessor might be chosen. There are several methods which are commonly used; for the most part, they harbour traps for the unwary.

Selection by Features

In principle, the features required in an application might be used to decide on the most suitable microprocessor. In practice, rejection of the devices not possessing these features is not likely to single out the best choice. The features which would be useful could be listed, and scored on a "weighted points" system, to place the contenders in rank order. The relative value of, say, a larger on-chip ROM and larger on-chip RAM would be hard to decide. Since the weights would have to be chosen arbitrarily, the conclusion would not be likely to be sound. For example, whether a 2 kword ROM is better than a 1 kword ROM clearly depends upon the program. The 2 kword device is not twice as good as the 1 kword device, particularly if the program needs only 1 kword. The danger of using tick lists and ranking schemes lies in their false air of objectivity.

Selection by Performance

It is common practice to measure the performance of a large computer system with a suite of "benchmark" programs, which attempts to simulate the mixture of operations likely to be encountered in the real workload. This may be reasonable for such an installation which tackles a variety of tasks. The greater the performance, the more scope there is for adding more users or more tasks. A dedicated microprocessor system differs from this in that generalized computing power is not a requirement. Any excess performance cannot be used in the application, so that adequacy is the criterion of performance.

Performance has more than one meaning when applied to microprocessors. A more comprehensive benchmark test is needed which takes into account all the factors relevant to the particular application in just proportion. Fortunately, there is one medium of comparison which does so: cost.

Selection on Cost

In a commercial enterprise, the measure which relates such diverse factors as performance and maintenance liability is cost. Cost is, after all, the basis upon which the viability of a business as a whole is established. When all the costs of development, production and maintenance are considered, the best choice of microprocessor for a given application is that which gives the lowest overall cost. Any characteristic which is relevant ought to affect the cost, and so establish its own degree of importance.

Not all cost factors can be accurately or reliably forecast, but even order-of-magnitude guesses for some of the values may be sufficient to pinpoint the critical areas. When considerations which cannot have a value placed upon them are then also weighed, warranted conclusions are more likely to be reached.

Selecting for a Range of Applications

Where many applications are to be satisfied from a small range of preferred microprocessors, the selection task becomes more complex, particularly if most of the applications lie in the future. Selection on cost becomes impractical and some compromise is necessary between selection on technical merit and selection on other principles.

COMMERCIAL CONSTRAINTS

Because of the moderate quantity requirements for telecommunications applications, the exploitation of semiconductor technology requires the use of devices that are successful commercially. This is not to say that custom-designed circuits are not favoured in instances where numbers are high, or where there are special requirements, but it is not proposed to discuss custom-built circuits here, except where such a circuit is itself a microprocessor.

Custom-Built Microprocessors

In principle, a microprocessor which is specifically designed for telecommunications would have a better performance than general-purpose commercial designs. In practice, the pace of advance in microprocessor technology is such that these performance advantages would probably be rapidly eroded by improvements in the performance of commercial devices generally. Within one or two years, the bespoke design would be likely to offer no advantage. The very high development cost for designing a chip, and providing support for the users, would also make the cost of each chip extremely high, when so few would be produced. The need for a range of microprocessors to fulfil the variety of applications which exist is a further barrier to the use of a bespoke design. As long as no special requirements emerge which cannot be met by commercial designs, most administrations seem unlikely to use bespoke designs of microprocessor.

The Lifetime Problem

The important problem of maintaining equipment in 30 years' time places great emphasis on accurate predictions of component reliability, security of supply and the longevity of the component as a product.

Approval and Reliability

Reliability is of great importance, not only to ensure that the equipment gives good service, but also because failure rates higher than those assumed during the design of the equipment would invalidate calculations of maintenance savings and run down stocks of spares. Reliability starts at manufacture, and so the aim of approval schemes is to ensure that a particular manufacturer is capable of achieving and maintaining acceptably high standards of reliability in the devices produced. Manufacturers' processes, even though nominally of the same type (for example, silicon-gate n-channel MOS), are all different. Each process must be assessed and approved individually. The large number of manufacturers, and hence of processes, prevents the approval of more than a small proportion. However, the approval of any particular device requires also the validation of its design. This and the testing of production devices are problems which are not yet fully solved[8].

The accessibility of a manufacturer and his willingness to cooperate in a rigorous approval procedure is a constraint upon the choice of devices. The cost and organizational effort put into the prediction and measurement of reliability is an incentive to minimizing the variety of devices and manufacturers. The approval of a process is common to many LSI products of that manufacturer, and so interest in other LSI products (chiefly memory devices) may influence the choice of processes, and hence, indirectly, the choice of microprocessor.

Sources of Supply

For practical reasons, 2 sources of supply in the UK are preferred; this is difficult to achieve, since there are so few UK designs of microprocessor. Most successful microprocessor designs already have sources in USA, Europe and Japan, some having as many as 6 secondary sources.

The other devices in the parts family may not be so well catered for, so that the use of some convenient peripheral devices may have to be avoided for the sake of security of supply. Most manufacturers are keen to have a secondary source of supply since many of their customers require it. It is salutary to note that one technically excellent microprocessor was taken out of production only because no other manufacturer was willing (or able) to duplicate the very advanced technology which made it so powerful.

Importance of the Manufacturer

The manufacturer also comes under scrutiny when technical support is considered, since the quality of the development aids, software utilities and documentation has a significant effect upon the ease and speed with which a microprocessor can be used. More than that, though, it is a fair indicator of his commitment to the product.

Commitment to particular designs of microprocessor, and to microprocessors generally, is important when so much depends upon being able to obtain a continuing supply of devices for production and maintenance in the future. Manufacturers vary considerably in their ability to produce commercially successful designs. Such success requires inspired design, sound support services, strong selling and a proven future commitment to microprocessors.

Knowledge of the manufacturer's aspirations and track record ought, perhaps, to be more influential in making a selection than even major technical features in individual designs of microprocessor. In choosing a microprocessor, a manufacturer is also being chosen.

Costs of Proliferation

The costs of training, development aids and other fixed costs can be reduced by avoiding the use of too many different designs of microprocessor. The chief incentive, though, may not be the costs directly associated with the development itself, but the largely unpredictable penalties of failure of supplies in the future. (Microprocessors are not unique in this respect, but are particularly difficult because of their myriad characteristics, which make direct substitution practically impossible.) Securing the sources of supply and achieving confidence in the reliability of the devices are expensive procedures. These costs can be reduced by the choice of intelligent standards. The savings are, admittedly, to be offset by the penalties incurred in using devices not ideal for some applications. Yet, without the overall view taken by the use of standards, each application would seek to minimize its own costs, heedless of the costs for applications taken as a whole.

A POSSIBLE SELECTION STRATEGY

It has been suggested that cost is the ideal medium of comparison but, in practice, short cuts have to be taken. By splitting the microprocessor field into loosely defined regimes, the best and worse devices in each class can be distinguished, even without much detailed knowledge of the applications, which mostly lie in the future. Microprocessors which are obsolescent, or which it would be impractical to approve, could be discarded. When the character of the manufacturer, user preferences, and prior experience are considered, a few standards tend to emerge for each class. A concentrated effort could then be made to evaluate more thoroughly these few devices for approval, along with support circuits and memories. The standards could be changed when an important application favoured a newer device.

THE FUTURE

Many commentators do not see an end to the rapid improvements being made in the complexity of LSI circuits. The gradual inception of truly functional minicomputers-on-a-chip, the availability of programmable ROMs without a price penalty over masked ROMs, and developments in computer languages, will eventually make it economic to use a powerful minicomputer for even the most trivial tasks. This, and the fact that many alliances have taken place between manufacturers to ensure survival, makes it seem possible that the present trend towards increasing diversity in designs might be reversed. That would ease the problem of selection. However, as the useful limits of system partitioning are reached, with programming costs becoming more significant than hardware costs, developments in language may eclipse improvements in the hardware. There is no reason to suppose that standards will be spontaneous in the language field, any more than they are in hardware now. The problem may simply change to one of selecting software.

CONCLUSION

This article has considered some of the influences on the performance of microprocessors and has shown that performance is really defined only in relation to the application. The converse is also true, in that the applications themselves are not defined in a vacuum, but are strongly influenced by knowledge of the performance and features of commercially available microprocessors. In a sense, developments in each build on one another in an iterative process. A further article in this *Journal* will deal with some of the applications of microprocessors in the British Post Office (BPO) Research Department, and will show to what extent microprocessors and their applications are tailored to each other.

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References

1 TONGE, J. D., GAUNT, D. L., and KENDALL, J. P. Programmable Logic and Microprocessors. *POEEJ*, Vol. 70, p. 136, Oct. 1977.

2 BENTON, R. H., COWAN, R. A., MAYNARD, K. J., and SCARFE, R. T. Programming and Microprogramming of Microprocessors. *POEEJ*, Vol. 70, p. 234, Jan. 1978.

3 MOORE, G. E. Microprocessors and Integrated Electronic Technology. *Proc. IEEE*, Vol. 64, No. 6, p. 837, June 1976.

4 VERHOFSTADT, P. W. J. Evaluation of Technology Options for LSI Processing Elements. *Proc. IEEE*, Vol. 64, No. 6, p. 842, June 1976.

5 REYLING, G. Considerations in Choosing a Microprogrammable Bit-Sliced Architecture. *Computer*, Vol. 7, p. 26, July 1974.

6 SHOEFFLER, J. D. Microprocessor Architecture. *IEEE Trans Industrial Electronics and Control Instrumentation*, Vol. IECI 22, No. 3, p. 256, Aug. 1975.

7 ASPINALL, D., BARTON, M. H., and DAGLESS, E. L. Comparison of Microprocessors: A Graphical Approach. *Euromicro Newsletter*, Vol. 3, Part 2, p. 84, Apr. 1977.

8 LIVINGSTONE, A. W. Automatic Testing of Complex Integrated Circuits. *POEEJ*, Vol. 70, p. 161, Oct. 1977.

The Regenerator 5A—A Microelectronic Project for Strowger Exchanges

Part 1—Development of the Mark I

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Microelectronic techniques have been exploited to produce an electronic replacement for the electromechanical Strowger pulse Regenerator 1A, which has been in service for nearly 40 years. These electronic Regenerators 5A, Mark I have now been in operational use for 7 years, improving service and reducing maintenance costs. The economies provided by the electronic regenerator have resulted in it being specified for new equipment, and for complete replacement of the Regenerators 1A. The development of the Regenerator 5A provided a timely stimulus to the telecommunications industry in the early phases of the application of microelectronics to exchange equipment. Part 1 of this article deals with the origin of the project and the development of the Mark I, and Part 2 will cover the development of later designs.

INTRODUCTION

The Regenerator 5A is an electronic replacement for Strowger pulse Regenerator 1A, which is a plug-in electromechanical device standardized by the British Post Office (BPO) around 1938. The Regenerator 1A receives and stores Strowger pulses and retransmits them with their speed and break-to-make ratio corrected to closely defined limits. This permits economies and flexibility in line plant provision.

By the early-1960s, 100 000 Regenerators 1A were in service and their usage was expanding rapidly, but changing conditions made it necessary to review the position. Although its design and associated circuits had been improved, performance defects caused intermittent misroutes; these were tolerable in lightly-used equipment, but had a greater impact in an era of high intensity of telephone traffic and subscriber trunk dialling. Furthermore, substantial maintenance effort was required for fault location and adjustment of the mechanism. Feasibility studies into electronic devices indicated that, at the time, there was no alternative that would fit into the space of the Regenerator 1A or have an acceptably low power consumption. The demand for regenerators continued to increase, and by 1966 more than 200 000 were in service.

In view of advances in microelectronics, with consequent scope for size reduction, it was considered an opportune time for further studies. BPO requirements for an electronic regenerator were therefore circulated within industry, followed in 1968 by invitations to tender for development of a BPO standard design. A number of companies submitted offers but, around this time, BPO procurement policy was changing to a performance-based philosophy with minimal specification of detail. Consequently, it was decided not to proceed with a standard design but, instead, to assess a private-venture design under development by Pye TMC Ltd. and, by joint effort, bring it to early completion. The device was progressively improved and, when it was deemed to be acceptable for operational use, 10 000 items were ordered from the Company as Regenerators 5A, Mark I.

Part 1 of this article describes the background to the project, the design aims and problems, and the realization of the Mark I type in some detail; this will facilitate an appreciation and understanding of the Marks II, III and IV designs to be described in Part 2.

SERVICE REQUIREMENTS

The development was based on a plant service requirement (PSR), prepared by Telecommunications Headquarters (THQ) Service Department, to provide an electronic regenerator as a plug-in replacement for the Regenerator 1A in all existing and new applications.

The PSR target figures were set so that the new item would be cost competitive with the existing one, have a superior performance, and incur lower maintenance charges. Modification of the associated circuits, testers and routiners was to be avoided, as was the refurbishing of existing components such as relays. With the passage of time, the exacting nature of the requirements became increasingly apparent to the development engineers concerned.

ELECTROMECHANICAL REGENERATOR 1A

The Regenerator 1A is described in detail in an earlier issue of this *Journal*[1], but its salient features are described below to illustrate the requirements imposed on the electronic replacement. The ingenious mechanism brought together techniques and parts already in production, to form a complex new item. Fig. 1 shows the Regenerator 1A in a junction relay-set. Three magnets control counting, storing and retransmission of the incoming pulses; a circle of 42 code-pins, free to move axially, act as a pulse-store, and a drive-wheel and cam operate the outgoing pulsing contacts.

Circuit and Operation

Fig. 2 shows the relationship between the regenerator and the main circuit in a 1938 junction circuit application. Subsequently,

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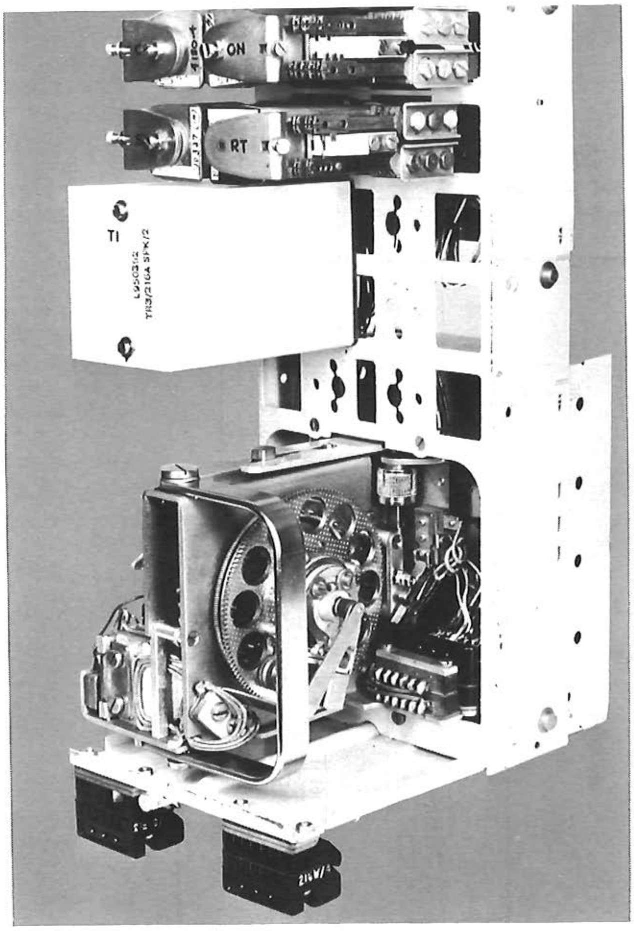


FIG. 1—Regenerator 1A in junction relay-set

many different circuits using regenerators were produced over the years to meet changing requirements.

Pulse Reception and Storage

Incoming pulses, repeated by relay A, are counted by the receiving and marking elements. Receiving magnet RM steps the storage ratchet and associated marking lever, which is lifted clear of the code-pins by marking magnet MM before stepping commences. Release of the lever indicates the end of the pulse train by moving a code-pin axially; that is, setting the pin. The interval between 2 adjacent code-pins corresponds to one pulse; thus, the number of stored pulses is indicated by the total interval between the 2 code-pins that have been set. As the ratchet rotates, energy is stored in a spring on the pulsing drive wheel.

Transmission of Stored Pulses

Transmission is effected by transmitting magnet TM, operating and releasing the resetting plunger, which rests against a set code-pin. On operation of magnet TM, the pin against which the plunger is resting is reset. The release of magnet TM then permits the pulsing drive wheel to be rotated by its spring until the plunger engages with the next set code-pin. During rotation, a cam actuates 2 pairs of pulsing springs to transmit main output pulses of 10 pulses/s and $66\frac{2}{3}$% break ratio, and auxiliary output pulses of 10 pulses/s and 50% make ratio for supplementary circuit functions if required. The receiving and transmitting elements function independently, permitting simultaneous storage and transmission. Relays in the main circuit control the inter-digital pause (IDP) between trains of outgoing pulses.

Mechanically-Operated Contacts

Mechanically-operated contacts play a part in the operation. Contacts NI are open when the regenerator is at rest. When pulses are received and the storage ratchet moves from rest,

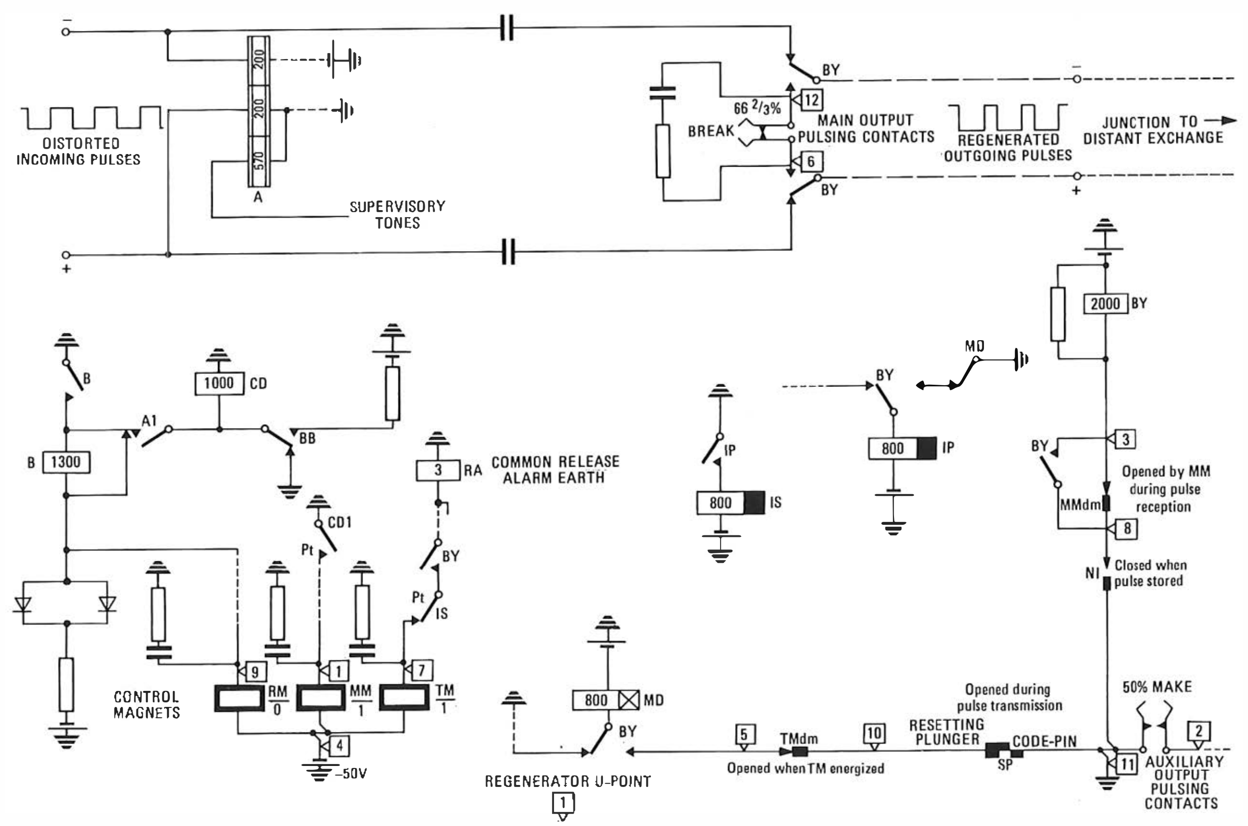


FIG. 2—Interconnexion of a Regenerator 1A and a junction relay-set

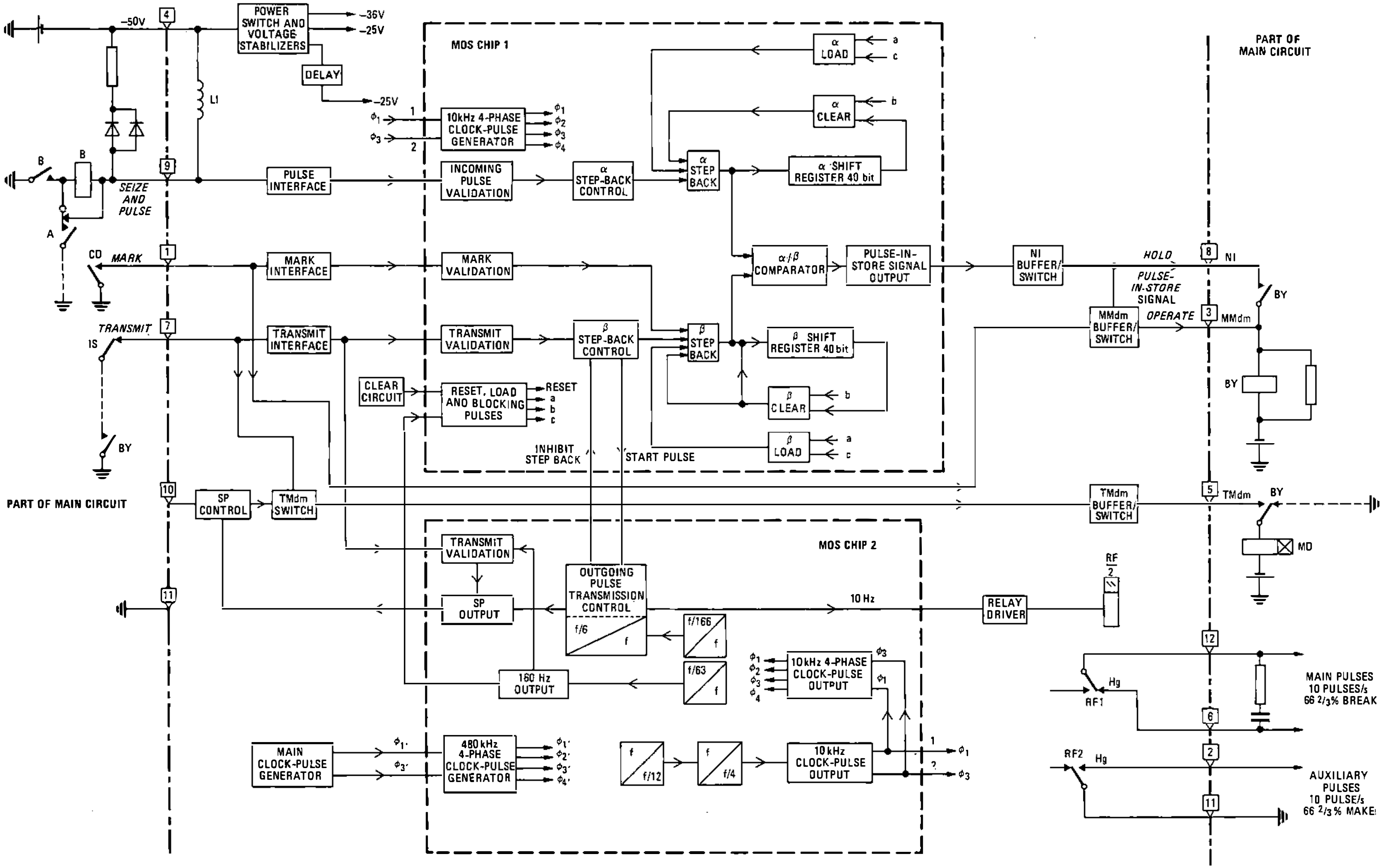


FIG. 3—Block diagram of Regenerator 5A, Mark I

contacts NI close and remain closed until all pulses have been retransmitted. Contacts MMdm open when magnet MM is operated by contact CD1 during the break period of the first pulse of each incoming train, and close when contact CD1 opens at the end of the train; this operates relay BY, which is then held until contacts NI reopen after all stored pulses have been transmitted. Contacts TMdm open on operation of magnet TM, and close on release. The resetting plunger and set code-pin form a pair of contacts, SP, used electrically in some circuits.

DEVELOPMENT SPECIFICATION FOR ELECTRONIC REGENERATOR

The development involved using a new rapidly-evolving technology in an environment electrically hostile to electronic equipment, and specification of requirements presented difficulties. BPO and British Standard Institution (BSI) specifications and quality-assessment procedures for micro-electronic components did not exist. A wholly "black-box" specification was not acceptable to the BPO. The invitations to tender for development of a standard design were therefore based on an outline specification, which was produced to guide development. The specification set out general design aims considered achievable within the *state of the art*, together with detailed mandatory functional requirements necessary to meet the PSR.

As development proceeded, the outline specification was to be updated for issue as a formal specification for competitive procurement. This approach was continued when the BPO decided not to proceed with development of a standard design but, instead, to co-operate in bringing the private-venture design to completion.

DESIGN PROBLEMS WITH ELECTRONIC REPLACEMENT REGENERATOR

Feasibility studies identified critical performance features and possible causes of malfunctioning which it was essential to cover in the design.

Spurious Input Signals

The relay contacts, which apply signals to regenerator U-points (U9, U1 and U7 for magnets RM, MM and TM respectively in Fig. 2), may give rise to spurious effects caused by mechanical bounce. Additionally, there may be incorrect re-operation of relay A during *break* periods of the incoming pulses. This can be caused by an oscillatory reverse surge through the windings of relay A, the incoming line and the resistor-capacitor quench across the remote pulsing contacts. In the Regenerator 1A, these spurious signals are absorbed by the relatively long magnet operate and release times. The problem was, how to obtain the same degree of protection in the electronic elements?

Holding of Relay B

When contact A1 is re-operated during *make* periods of incoming pulses, the induced voltage developed across magnet RM by the change in its flux accelerates the growth of current in relay B and improves its holding performance[2].

To avoid degradation of the relay's performance with the electronic version, a source of energy equivalent to magnet RM was required, and the provision of a corresponding inductor in the electronic device appeared to be the best solution.

Protection against Overvoltage Surges and Electrical Noise

These requirements are interrelated and are often dealt with as a single design aspect described as *noise*. Although features providing protection against overvoltages might eliminate some noise interference, particularly at low frequencies, and noise protection is effective to some extent against overvoltages, it is prudent to consider them separately.

Overvoltage Surges

In Strowger telephone exchanges, high transient voltages are developed in the wiring, and on the common battery-supply leads, during switching of mechanism magnets and relay coils. Furthermore, if a battery-supply fuse serving groups of equipment blows, disconnecting a distributed load of many coils, heavy surges can occur on the distribution leads, and also on the commoned battery-supply feeds isolated by the blown fuse. These phenomena have been known for many years, but have not been fully investigated as their effects were limited to audible clicks, which were dealt with by segregation of wiring and screening, and by the use of inductors as choke-coils. However, experience with diodes and transistors in early electronic applications in exchanges showed that, where they interfaced with electromechanical equipment, protection was generally essential, and limited information existed from these activities. This information was made available for design guidance, without prejudice to the requirement for the best *state of the art* protection. Essential details were:

(*a*) transient overvoltages on 50 V battery-supply leads:
250 V for 0·5 ms, 2 kV for 1 μs;
(*b*) transient overvoltages on other leads:
2 kV for 1 μs.

Noise Interference

Strowger equipment is well-known as a source of electrical interference, capable of adversely affecting electronic circuits. Interference sources and modes of propagation are described in an earlier issue of this *Journal*[3]. The main sources are induced voltages from coils, surge currents in wiring, and local high-frequency oscillations at sparking and arcing contacts. Possible causes of interference existed in the various main circuits. Sparking and arcing were known to occur at the contacts of relays MD, IP, IS, etc., due to the effects of their slow-to-release features. It was not possible to quantify the characteristics of these interference sources.

Protective Measures

Retrospective action to suppress overvoltages and interference at source, on each of the large number of relays involved, was out of the question; therefore, all protective measures for the control leads had to be within the regenerator. Similarly, protection within the regenerator was necessary for the battery-supply leads, although a single large device, common to a rack or group of circuits, had some attraction and has been used in a new design of signalling equipment.[4] For the regenerator, however, this would have departed from the aim of avoiding changes to existing equipment and would have been relatively costly to implement.

ELECTRONIC REGENERATOR 5A, MARK I

Design Philosophy

The design is based on p-channel metal-gate, metal-oxide-semiconductor (MOS) integrated-circuit technology,[5] in which the inherent capability of MOS technology of providing a large number of logic function devices in an extremely small area is exploited in 2 large-scale-integration (LSI) chips. These were designed specifically for the regenerator by the Pye TMC designers and the information prepared in a form suitable for chip production by semiconductor manufacturers. There were other possibilities: for example, the device could have been produced from a larger number of standard (catalogue) integrated circuits, suitable for diverse applications; or the detailed design of the logic and chips could have been left to the semiconductor manufacturer.

At that time, there was some controversy in industry as to the relative merits of the various options, but the subsequent course of events amply justified the decisions made for the regenerator.

Realization of Design

The final Mark I design is shown in block form in Fig. 3. It comprises, essentially, 2 MOS LSI chips using 4-phase dynamic logic,[5] together with a multivibrator clock-pulse source and peripheral interface and buffer/switch elements made up from discrete components. The construction and component arrangements are shown in Figs. 4 and 5.

MOS 4-phase Dynamic Logic

The principles of MOS devices, their use to perform logic functions, and basic application configurations have been described in earlier articles[5].

Nearly all MOS integrated circuits have been developed from simple inverter configurations, but a particular logic function may involve a number of MOS devices on the chip as can be seen in Fig. 6, which shows a dynamic 4-phase 1 bit shift-register stage, together with the clock-pulse waveforms.

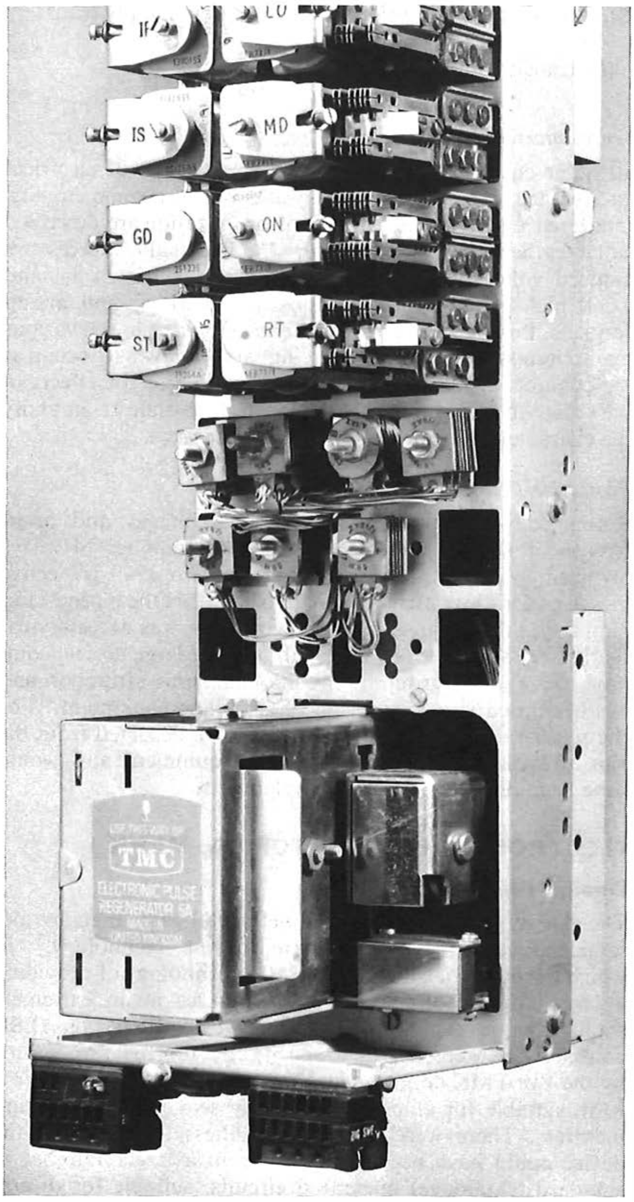


FIG. 4—Regenerator 5A, Mark I in junction relay-set

This element illustrates the principle of clocking information through logic stages. When the ϕ_1 and ϕ_2 pulses are at logical 1, semiconductors TR2 and TR3 are ON and the node capacitance, designated *C1*, is charged. If the input is also at logical 1, semiconductor TR1 is ON and, when the ϕ_1 pulse returns to logical 0 (earth) C1 is discharged via semiconductors TR1 and TR2 during the remaining ϕ_2 pulse period. If the input is at logical 0, semiconductor TR1 is not ON and C1 remains charged.

When the ϕ_3 and ϕ_4 pulses are at logical 1, semiconductors TR5 and TR6 are both ON and, if capacitance C1 has remained charged, semiconductor TR4 is also ON. The node capacitance (C2) of semiconductor TR7 in the next stage, is charged via semiconductor TR6. When the ϕ_3 pulse returns to earth, semiconductor TR6 is turned OFF and, during the remaining ϕ_4 pulse period, C2 is discharged via semiconductors TR4 and TR5, indicating logical 0 to the next stage. If C1 has been discharged, semiconductor TR4 is OFF and C2 remains charged, indicating logical 1 to the next stage. Thus, logical 0 or 1 applied to the input appears at the output one clock period later.

In the stage, 2 successive inversions, each of a half clock period duration, are carried out as the input signal is clocked through. Clock pulse control, in this manner, requires the outputs and inputs of successive gates to be compatible.

Internal Power

On seizure, the -50 V supply to the regenerator is extended to stabilizer circuits, which provide -36 V to MOS chip 2, and -25 V to MOS chip 1 and peripheral circuits. To avoid spurious output pulses during seizure, the -25 V supply to the output buffers is delayed until the chip-seizure is complete.

Clock Pulse Supplies

The multivibrator provides 2 basic pulse supplies, ϕ_1' and ϕ_3' at 480 kHz, which are expanded to 4-phases ϕ_1'–ϕ_4' in MOS chip 2, to control divide-by-12 and divide-by-4 counter stages, producing ϕ_1 and ϕ_3 pulses at 10 kHz. These are expanded to 4-phase, ϕ_1–ϕ_4, for dynamic control of logic functions, including divide-down elements producing the 10 pulses/s outgoing pulse supply, and also a 160 pulses/s supply for the transmit validation element. The 10 kHz ϕ_1 and ϕ_3 pulses are extended to MOS chip 1, and expanded to 4-phases for dynamic control of the registers and other logic functions in that chip. Similarly, the 160 pulses/s supply is extended to MOS chip 1 for the register clearing and resetting elements.

Input Interfaces

The interface elements in the input paths, through which signals are received by the regenerator, translate the signals

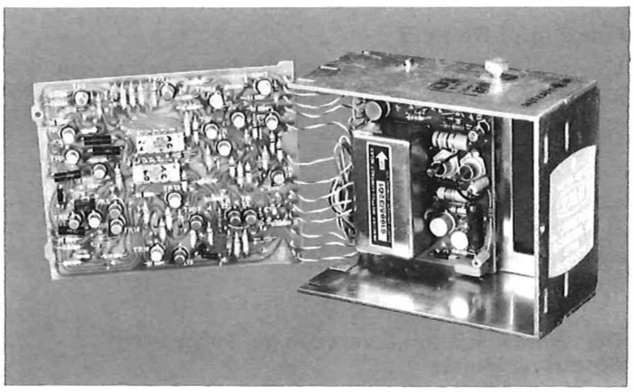


FIG. 5—Construction and component layout of Regenerator 5A, Mark I

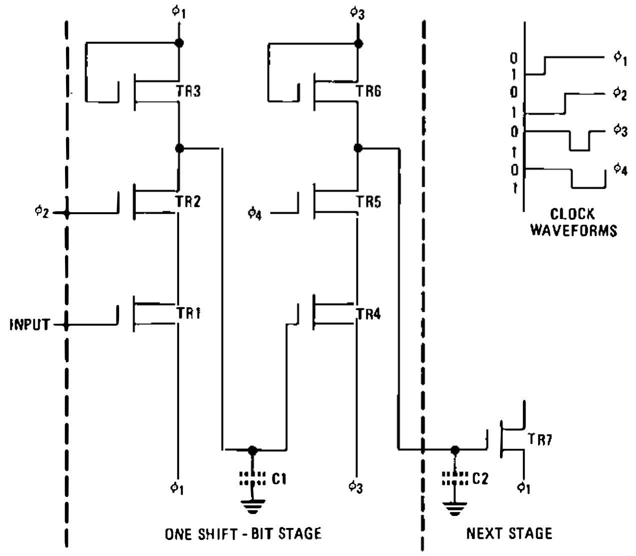


FIG. 6—Dynamic 4-phase 1 bit shift register stage

to logic levels compatible with the chip inputs. An inductor, L1, connected to the *seize and pulse* lead at U9, acts as a source of energy, equivalent to magnet RM in the electromechanical regenerator, to maintain the holding performance of relay B.

Signal Validation

Validation elements are provided in the incoming leads to prevent spurious signals, caused by irregular operation of relay contacts (such as contact bounce), being accepted by the regenerator. The validation element in the seize-and-pulse input must also deal with false pulses caused by reverse surges in the winding of relay A, described earlier. The validation elements consist of chains of bistables controlled by strobes, and signals are not passed on unless the changes-of-state are maintained for prescribed periods.

Output Buffer Switches

Because the chip outputs do not have sufficient current-carrying capability to directly control relays in the main circuit, transistor switching stages are interposed as buffers in the output paths via which signals are passed from the regenerator.

Output Pulses

The pulses from the 10 pulses/s output buffer are repeated to the main circuit via a 2-change-over mercury-wetted-relay (relay RF).

Protection

A Zener diode in the regenerator internal power switch provides protection against excessive voltage rise on the −50 V battery supply lead. Decoupling capacitors are used to protect the −36 V and the −25 V internal supplies against induced electrical noise.

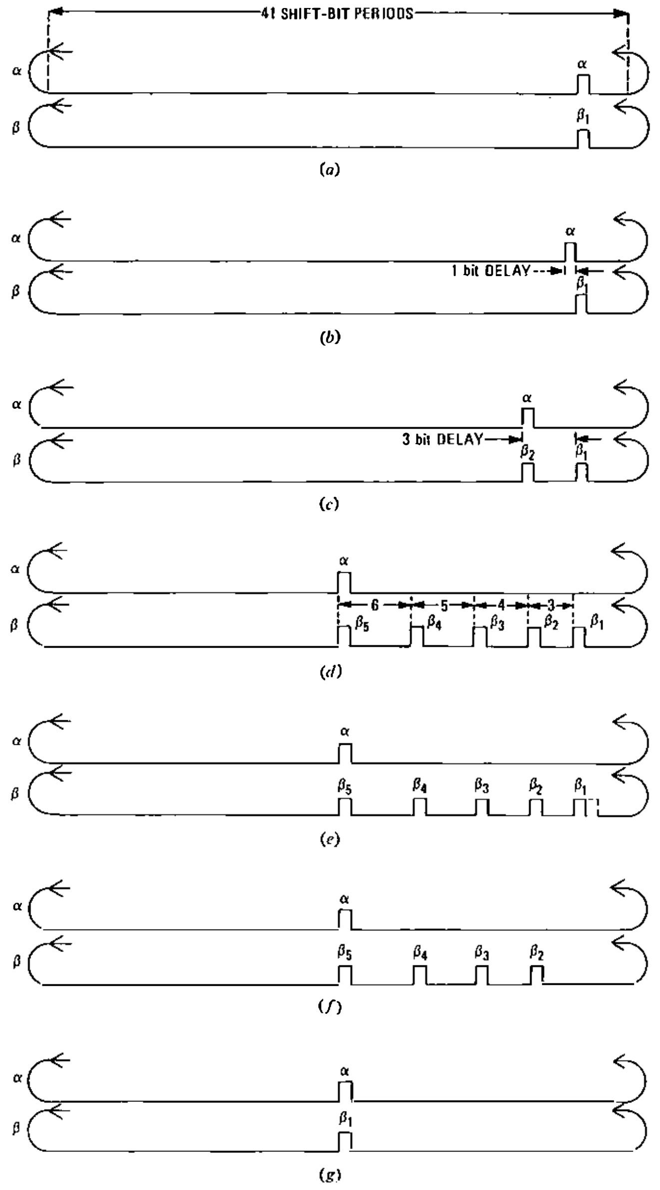
The interface elements in the input leads incorporate current limiters, which protect the MOS chips against voltage surges that originate from the main circuit. The semiconductors in the output buffer switches are protected by clamping diodes.

The chip inputs and outputs are protected by diode and resistor arrays; these are incorporated within the chip as part of the MOS manufacturing process[6].

Outline of Operation

The regenerator is seized when the main circuit is taken up by a call; internal power and clock pulse supplies are established, bistables are reset, and the α and β registers cleared. Single pulses are then loaded into both the α and β registers, and they circulate in synchronism until pulses are received from the main circuit (Fig. 7(*a*)). The incoming pulses are first validated to ensure that spurious signals are not passed on and then applied to the α step-back circuit, which precedes the α register; each valid pulse steps back (that is, delays) the circulating α pulse by one shift-bit period (Fig. 7 (*b*)). Thus, at the end of the incoming pulse train, the α pulse is delayed relative to the β pulse by a number of shift-bit periods equal to the number of pulses in the train. The end of the train is marked in the β register by the loading of another β pulse in synchronism with the delayed α pulse (Fig. 7 (*c*)).

Similarly, further pulse trains are counted in the α register and stored and marked by additional pulses in the β register. For convenience, the successive β pulses are referred to as β_1, β_2, etc. Fig. 7 (*d*) shows digits 3, 4, 5 and 6 stored.



(*a*) Seizure: α and β pulses in sychronism
(*b*) One pulse stored; α pulse stepped back 1 bit
(*c*) Three pulses stored; digit 3 marked
(*d*) Digits 3, 4, 5 and 6 stored and marked
(*e*) One outgoing pulse of digit 3 transmitted; β_1 stepped back 1 bit
(*f*) Digit 3 transmission complete
(*g*) All digits transmitted; α and β pulses back in synchronism

FIG. 7—Pulse sequences in α and β registers

The lack of synchronism between the α and β_1 pulses, following the first incoming pulse and α step-back, is detected in the α/β comparator, and a *pulse-in-store* signal is applied to the main circuit *hold* lead and switch MMdm. When a complete train of incoming pulses has been received, switch MMdm is closed; this extends the *pulse-in-store* signal to the *operate* lead to operate relay BY, thereby connecting the regenerator pulsing contacts to the outgoing path (Fig. 3).

Transmission of outgoing pulses cannot commence until a complete train of pulses has been stored. Then, a signal via the *transmit* lead causes the β step-back element to delay the β_1 pulse one shift-bit period (Fig. 7 (*e*)), while the outgoing pulse transmission control sends a 10 pulses/s output pulse to the main circuit. This step-back sequence is repeated for each stored pulse until β_1 reaches the β_2 position, indicating the end of the digit (Fig. 7 (*f*)). Transmission then ceases, the outgoing IDP follows, and a further transmit signal starts transmission of the second train, which had been received and stored in the β-register during transmission of the first train. This action is repeated until all stored pulses have been transmitted.

The remaining single β pulse and the α pulse are then back in synchronism (Fig. 7 (*g*)) and the *pulse-in-store* signal ceases, releasing relay BY, and leaving the regenerator in the seized state for the duration of the call. On cleardown, the seizing signal is removed and the internal power switched off.

TEST PROGRAMME

Close co-operation between the BPO and the Company brought the design to model form for general validation tests in THQ Telecommunications Development Department's (TDD's) Circuit Laboratory. In this phase of the project, the design progressed from a large breadboard of discrete components and small integrated circuits to an early plastic-case model, which included the 2 MOS chips, but used reed relays to repeat the signals at the inputs and outputs interfacing with the main circuit. This permitted the operation to be checked without complications from interference and over-voltage surges. The relays were then progressively replaced by semiconductor circuit elements and the case reshaped in metal sheet.

After incorporation of changes arising from the laboratory tests, 25 prototypes were purchased for a more extensive programme of laboratory checks of critical features and life tests, and for trials in the field.

The early results were considered promising enough to permit time to be saved by ordering 10 000 regenerators to the up-dated outline specification while the programme was still in progress. This order was large enough for economical production, and for BPO assessment of the performance of a significant quantity of the new form of device.

INSTALLATION AND PERFORMANCE

THQ Service Department made special arrangements for installation, investigation of faults and feedback of information. The object was to validate and establish confidence in the new regenerator under operational conditions, more rapidly than was possible with the normal distribution arrangements, by fitting quantities at a large number of exchanges embracing the full range of main circuits.

Deliveries and installation proceeded concurrently from late-1970 to late-1972. Initial supplies were fitted in 2000-type satellite exchange discriminators; these are short-holding-time equipments, in which each regenerator handles up to 360 000 calls per year, regenerating approximately 12-million pulses. Later, installation was extended to less-heavily-worked, long-holding-time junction circuits at large centres, and, finally, as confidence built-up, to small unattended exchanges.

Regular performance checks were made, including functional tests and measurement of the speed and break-to-make ratio of the outgoing pulses. Any regenerators considered to have failed in the field were sent to TDD's Circuit Laboratory for confirmatory tests before being returned to the manufacture for investigation and report.

The arrangements worked well: teething troubles were identified early and obscure faults in exchange equipment were revealed; these had been tolerated by the electromechanical regenerator for years, but were hazardous to the electronic replacement. Appropriate information was fed back to the manufacturer and disseminated to the Regions and Areas.

The regenerators settled down well and, even before the installation programme had been completed, the performance trends were considered very favourable and authority was given to expand the project.

SUMMARY

This part of the article has described the development of the Regenerator 5A, Mark I. Part 2 will describe later designs.

ACKNOWLEDGEMENT

The success of the Regenerator 5A, Mark I project, which was the first device in the world to employ LSI MOS techniques successfully in Strowger electromechanical equipment, is a tribute to the enterprise and expertise of Pye TMC Ltd. and their team of development engineers.

References

1 EDWARDS, A. G., and LAWRENCE, J. A. The Mechanical Impulse Regenerator. *POEEJ*, Vol. 30, p. 261, Jan. 1938.

2 ATKINSON, J. Telephony, Vol. 2, p. 161. New Era Publishing Co., London, 1950.

3 SCOTT, W. L. Electrical Contacts in Telephone Exchanges: Contact Opening and Closing Phenomena and Quenching Techniques, Part 1. *POEEJ*, Vol. 61, p. 263, Jan. 1969.

4 HILL, R. A., and GEE, J. H. A Miniaturized Version of Signalling System AC No. 9. *POEEJ*, Vol. 65, p. 216, Jan. 1973.

5 REYNOLDS, F. H., and MORTON, W. D. Metal-Oxide-Semiconductor (MOS) Integrated Circuits, Parts 1 and 2. *POEEJ*, Vol. 62, p. 168, Oct. 1969, and Vol. 63, p. 105, July 1970.

6 MULLARD. MOS Integrated Circuits and their Applications, 2nd. Edition. Mullard Ltd., London, 1973.

A New Method of Tone Identification

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UDC 621.395.667: 621.295.63: 6381.31

This article describes the method of tone identification that has been adopted for the measurement and analysis centre project.[1] *The method could be applied in other computer-based systems.*

INTRODUCTION

A tone-identification system is a feature that is commonly incorporated within telecommunications equipment. The recent development of measurement and analysis centres (MACs) included the requirement for a method of tone identification. The MAC system is a minicomputer-based network-surveillance system that provides quality-of-service performance statistics, and also acts as a maintenance aid by identifying faulty equipment. In the MAC system, test calls are generated to terminate at distant telephone exchanges on test-number relay-sets, which are designed to return a special test tone if the call is established correctly. The MAC system is required to identify this tone and all the other supervisory tones which can result in the event of failure of the test call. The availability of the MAC minicomputer and its high timing-accuracy, enabled a considerably more sophisticated algorithm to be used in the tone-identification process than in previous relay-operated equipment. This article describes the design of the tone-identification system adopted for the MAC system; it is likely that this design could be applied in telecommunications systems that incorporate a computer and which have a requirement for a tone-identification system.

MAC TONE-IDENTIFICATION REQUIREMENTS

The MAC system is required to identify the following conditions:

(*a*) MAC test-number tone (TNT1),
(*b*) multi-metering test-number tone (TNT2),
(*c*) busy tone (BT),
(*d*) equipment-engaged tone (EET),
(*e*) ringing tone (RT),
(*f*) number-unobtainable (NU) tone,
(*g*) no tone (NT),
(*h*) congestion announcement (CA) tone, and
(*i*) pay tone (PT).

The identification is based on the tone cadences detected in 2 frequency ranges: the first being centred on 400 Hz and the second being centred on 1004 Hz. All the tones can be considered to contain 400 Hz only (except the MAC test tone TNT1, which consists of alternate bursts of 400 Hz and 1004 Hz). The frequency of 1004 Hz was chosen because it is used to perform transmission level measurements and it is a CCITT* requirement that these be performed at 1 kHz. It is also required to identify all the tones over very wide tolerances of time and frequency variation. The design of the tone-identification system also accommodates some degree of corruption to the ideal tone cadences in the form of superimposed false ON and OFF signals.

The MAC system is able to originate test calls on a maximum of 30 test circuits simultaneously, therefore the tone-identification system had to be capable of identifying tones received on 30 circuits in parallel; this requirement dictated that the algorithm be capable of quick and efficient evaluation.

MAC TONE-RECOGNITION HARDWARE

Each MAC test circuit incorporates 2 tone receivers that are used to detect 400 Hz and 1004 Hz signals. The 400 Hz receiver operates to frequencies within the range 340–500 Hz. The power levels of the received signals at the MAC access equipment interface with the telecommunications network lie in the range 0–35 dBm. The 1004 Hz receiver is required to operate in the range 994–1014 Hz at power levels in the range 0–35 dBm.

The first level of noise rejection and the accommodation of corruption is performed by the tone-recognition hardware. This is achieved as follows:

(*a*) Tones received at the interface are passed through an analogue filtering system and the output is digitally integrated. The purpose of the digital-integration phase is to provide rejection of very short ON or OFF pulses.‡

(*b*) The effect of the digital integrator is that a 400 Hz input signal must be received for 60 ms before a 400 Hz ON output is given by the integrator and a 400 Hz OFF condition must be sustained for 16 ms before it is recognized as a valid signal state; that is, ON pulses of 400 Hz of duration less than 60 ms and 400 Hz OFF pulses of duration less than 16 ms, are rejected. Similarly, ON and OFF pulses of 1004 Hz of duration less than 60 ms are rejected.

EFFECTS OF DIFFERENT METHODS OF DIGITAL INTEGRATION

In the design of the 400 Hz detectors, much consideration was given to the question of whether to bias the output towards the OFF state. Two approaches to digital integration of the analogue input signal were considered; the difference between the 2 approaches was the time taken for a 400 Hz ON output to revert to a 400 Hz OFF output when the 400 Hz input is discontinued. Under the first method, referred to as the *long method*, this time was 60 ms: under the alternative method, referred to as the *short method*, the time was reduced to 16 ms. One reason why the short method was adopted eventually was that, since OFF was the normal state and because signal information and noise are presented as the ON state, it was considered to be good engineering practice to bias the output towards the OFF state. In addition, the short

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* CCITT: International Telegraph and Telephone Consultative Committee

‡ For convenience of description, the term OFF pulse in this article is used to indicate the absence of signalling information

method was found to have definite advantages in the identification of RT and CA tone, as explained later in the article.

The delay time for the receiver output to revert to the OFF state is limited to a minimum of 15 ms. This limit was set because RT can be so heavily modulated that it can appear to have breaks approaching this time in the envelope at the threshold level of detection.

Investigation Procedure

The effects of the 2 methods on noise rejection and the identification of tones were compared using the following procedure.

The hardware was modified to include both methods of integration, and the inclusion of a single switch allowed fast change-over between the 2 methods of operation.

An output was made available from the hardware, consisting of the digital representation of the input signal after digital integration. This facility was particularly valuable as it became a simple matter to determine when the input signals were of sufficient duration and amplitude to give rise to output signals. In addition, by observing frequency-indicator light-emitting diodes in the tone-recognition hardware, it was possible to determine immediately when output pulses were generated.

An audio-frequency trace recorder was used to make visual recordings of the tones received from the telecommunications network and of the corresponding digital representation output from the hardware.

Two high-accuracy frequency generators were used so that various time-bases could be superimposed on the analogue and digital traces of the tone for calibration purposes.

A technique applied widely throughout the investigation was to record a number of cycles of a tone using the long-method of digital integration, and then change to the short-method and record further cycles of the same test call.

Noise Immunity

During the evaluation of the effect of the 2 methods, it became apparent that the most common form of noise experienced during OFF periods was large numbers of closely-spaced short-duration low-amplitude pulses. When these pulses were of greater amplitude than the threshold level of detection, they were seen as a series of peaks. It is possible for a series of short-duration pulses to be integrated into a continuous 400 Hz ON output by the digital-integration circuit. Under the long-method of digital integration this effect occurs when the noise is present for greater than 50% of the time whereas, under the short method, this effect occurs only when the noise is present for greater than 79% of the time. Thus, the probability of noise giving rise to spurious pulses is greatly reduced when using the short-method of digital integration.

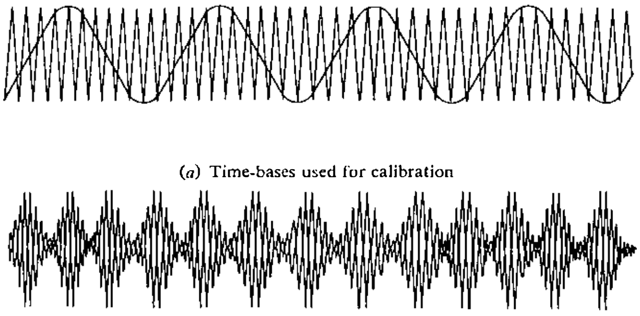
It was found, in practice, that the proportion of test calls which contained any noise above the threshold of detection was very low. Also, it was shown that the problem of spurious OFF periods during ON periods is not significant.

Effect on Ringing Tone and Congestion Announcement

Except for CA tone and RT, little difficulty was experienced in identifying any of the supervisory conditions with either method of digital integration and the low noise conditions experienced. A minor problem was experienced in identifying RT, and more severe problems were encountered in identifying CA tone, which is notoriously difficult to identifiy.

Ringing Tone

During the tests, it was shown that an occasional feature of RT is the presence of low-level background RT during the nominally silent portion of the tone cadence. Because RT is a heavily modulated signal, the appearance of the background RT is a series of peaks which may be above the threshold level of detection, depending on the amplitude. Fig. 1 shows the form of typical RT modulation, recorded on the network, also shown is the superimposed time-bases used for calibration purposes.


(*a*) Time-bases used for calibration

(*b*) RT modulation

FIG. 1—Typical RT modulation waveform

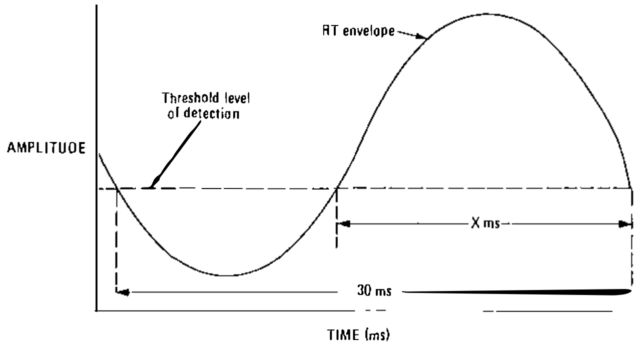


FIG. 2—RT modulation cycle

The cycle time of RT modulation is approximately 30 ms, as shown in Fig. 2. The period for which the tone is above the threshold level of detection depends upon the amplitude of the RT envelope, and is equivalent to x milliseconds in Fig. 2; the duration of the OFF pulse is therefore $(30-x)$ milliseconds. It is theorectically possible for the series of peaks generated by background RT to be integrated into a spurious 400 Hz ON output by the digital-integration circuit. The probability of this occurring is lower using the short method, as is indicated below.

(*a*) Long-integration method

If $15 < x \leqslant 30$, then continuous 400 Hz ON output.

If $0 \leqslant x \leqslant 15$, then continuous 400 Hz OFF output.

(*b*) Short-integration method

If $23 \cdot 7 \leqslant x \leqslant 30$, then continuous 400 Hz ON output.

If $0 \leqslant x < 23 \cdot 7$, then continuous 400 Hz OFF output.

The effect of background RT breakthrough was investigated and it was established that, using the long-method of digital integration, on some occasions, genuine ON pulses were elongated by background RT and spurious ON pulses were generated. The spurious pulses were always successfully removed by the short-method of digital integration.

Congestion Announcement Tone

Using the short method, the effect on the digital representation of a speech input was predicted to be that the proportion of 400 Hz OFF signals would be increased at the expense of the 400 Hz ON signals. During the evaluation tests of the 2 methods this effect was clearly demonstrated. Under the long method, the speech portion of the CA cycle was seen to give rise to prolonged periods of the 400 Hz ON state, sometimes as long as 770 ms. The presence of long-duration pulses makes the identification of CA tone more difficult and, in some instances, impossible. To establish a unique definition of some CA tones, it is essential that not more than one pulse of duration greater than 285 ms is present in the cycle because, if this requirement cannot be met, the definitions of CA tone and RT overlap. The effect of the short method of digital integration was that the long-duration pulses were consistently broken up. In addition, some short-duration pulses seen using the long method were removed completely by the short method. However, the total number of discrete pulses seen during each cycle was approximately the same using either method of digital integration. Using the short-method of digital integration there was rarely more than one pulse of duration greater than 285 ms detected during a cycle of CA.

TONE-IDENTIFICATION SOFTWARE

The algorithm used is based on threshold duration counters, and is an extensive development of one that has been applied successfully, over many years, in the British Post Office (BPO) TRT 119 call sender. Early techniques of tone identification, similar to that used in the TRT 119 call sender, have been described in previous issues of this *Journal*.[2, 3] The principle of operation is to isolate unique features of the tone, rather than perform a rigorous cadence comparison.

It was not possible to adopt the TRT 119 call sender algorithm in its entirety for a number of reasons: its most serious inadequacy is that the process of tone identification is too slow (to distinguish between CA tone, EET, BT and RT requires a 12 s sample); secondly, it does not positively identify CA tone because the category used for all unrecognizable tones is CA (for example, recorded announcements and double switching giving rise to speech). It was required that a MAC should positively identify CA tone and have an additional category for unrecognizable tones, known as *unrecognizable result* (UR).

MAC TONE-RECOGNITION PROGRAM

The tone-recognition program is based on the use of threshold duration counters; these count the number of pulses received which exceed certain durations during a fixed sample period. For each pulse received, all of the threshold duration counters of less than the duration of the pulse are incremented. A significant improvement of the algorithm (compared with the TRT 119 algorithm) is its ability to look at OFF pulses as well as ON pulses. The success of the program is dependent upon the choice of values for the key parameters. These parameters are:

(*a*) the number of threshold duration counters,
(*b*) the threshold durations of ON and OFF pulses,
(*c*) the fixed sample period, and
(*d*) the minimum and maximum permissible values for each tone.

Considerable development effort was required to establish the correct parameters, and the values selected are shown in Table 1. Many thousands of test calls were made to obtain examples of supervisory tones generated at many centres in

TABLE 1

Definition of Tones at a MAC and Test-Circuit Interface

		400 Hz ON Counters										400 Hz OFF Counters						1004 Hz ON	
Threshold Durations		60 ms		170 ms		285 ms		1000 ms		2500 ms		265 ms		800 ms		1100 ms		900 ms	
Tone Number	Tone Name	Min	Max	Min	Max	Min	Max	Min	Max	Min	Max	Min	Max	Min	Max	Min	Max	Min	Max
1	BT	7	11	7	11	7	11	0	0	0	0	7	11	0	0	0	0	0	0
2	EET	7	11	7	11	3	6	0	0	0	0	7	11	0	1	0	1	0	0
3	RT	3	6	3	6	2	6	0	0	0	0	2	3	2	3	2	3	0	0
4	NU	1	2	1	2	1	2	1	2	1	2	0	1	0	0	0	0	0	0
5	CA1	7	12	2	6	2	6	0	0	0	0	2	11	1	7	1	5	0	0
6	CA2	4	11	1	6	0	1	0	0	0	0	2	10	1	7	1	5	0	0
7	CA3	2	4	1	4	0	0	0	0	0	0	2	5	1	5	1	5	0	0
8	TNT1	1	3	1	2	1	2	1	2	0	0	1	4	1	3	1	3	1	2
9	TNT2	1	5	1	4	1	4	1	3	0	0	1	4	1	4	0	3	0	0
10	NT	0	1	0	0	0	0	0	0	0	0	1	2	1	2	1	2	0	0

Min: Minimum Max: Maximum
Note: Sample period of all tones is 6 s

the UK. The cadences of large numbers of operational tones were measured accurately, and were recorded using the MAC equipment. However, it is known that there is scope for variation of these parameters and it is considered that, while the values chosen are satisfactory for successful operation, there is room for further refinement. Since the program is controlled by a table of data, it is a relatively simple process to adjust parameters and to add new, as yet undefined, tones within the limits of storage available.

The fixed sample period chosen was 6 s, this represents an integral multiple of the nominal cycle time of the majority of supervisory tones.

The threshold durations adopted were as shown in Table 2.

TABLE 2

Threshold Durations

400 Hz ON	400 Hz OFF	1004 Hz ON
60 ms 170 ms 285 ms 1000 ms 2500 ms	265 ms 800 ms 1100 ms	900 ms

There were many factors involved in the choice of the threshold durations, and some of these may be deduced by studying the various tone cadences shown in Table 3. A brief summary of the reasons for selection of these threshold durations follows.

(*a*) 60 ms ON

This was a natural choice because this threshold duration is provided automatically by the hardware and provides a basic count of discrete pulses.

(*b*) 170 ms ON

The primary purpose of this counter is to record the shorter pulse of EET, and the burst of tone in the CA cycle.

(*c*) 285 ms ON

This counter is primarily required to record all the pulses of BT while avoiding the shorter pulse of EET; thus, the discrimination between the similar tones BT and EET is provided by this counter.

(*d*) 1000 ms ON

This counter is primarily required to record the ON pulses of TNT2 and to distinguish between RT and TNT.

(*e*) 2500 ms ON

This counter is required to record the identification of NU tone. Any burst of 400 Hz of duration longer than 2500 ms is considered to be NU tone.

(*f*) 265 ms OFF

This counter is required to record the shorter OFF periods of EET.

(*g*) 800 ms OFF

This counter is required to record the OFF pulses of TNT2.

(*h*) 1100 ms OFF

This counter is required to assist in the identification of CA tone. (A characteristic of CA tone is a period of silence of at least 1100 ms.)

(*i*) 900 ms ON (1004 Hz)

This counter is required to record the 1004 Hz pulses in the TNT1 cycle.

TABLE 3

Tone Cadences

Tone	Cadence	
	Nominal	Tolerance
BT	0·375 ON 0·375 OFF	0·285–0·435 0·285–0·435
EET	0·400 ON 0·350 OFF 0·225 ON 0·525 OFF	0·305–0·460 0·265–0·405 0·170–0·260 0·400–0·605
RT	0·400 ON 0·200 OFF 0·400 ON 2·000 OFF	0·305–0·460 0·150–0·230 0·305–0·460 1·520–2·310
NU	Continuous	
PT	0·125 ON 0·125 OFF	0·080–0·167 0·080–0·167
CA	0·200 ON 3·800 Speech 1·000 Silence	0·190–0·210 3·000–4·000 0·800–1·800
TNT1	1·992 ON 0·996 OFF 1·992 ON (1004 Hz) 0·996 OFF	±0·005
TNT2	1·500 ON 1·500 OFF	1·000–2·200 0·800–2·000

Note: All frequencies are nominally 400 Hz except where stated for TNT1.

The range of permissible values for each of the counters shown in Table 1 is necessary to cater for all the normal variation of cycle times and pulse durations shown in the tolerance column of Table 3. This range of values enables a tone which starts out with a cadence close to the ideal cadence to be corrupted on the network and yet remain within the definition given in Table 1. The single definition of a tone holds for many ways in which the ideal cadence may be corrupted. Obviously, if a tone is generated at the outset with a borderline cadence then the degree of further corruption that can be accommodated is much less. Another advantage of this method of representation is that a single definition of a tone holds for all possible points of entry into the cycle.

Other points of interest concerning the choice of values in Table 1 are as follows.

(*a*) It is possible for NU tone to include a break in continuity of up to 800 ms.

(*b*) It is possible for NT to include a single burst of tone of up to 170 ms.

(*c*) Three definitions of CA tone are required to cater for the different amplitudes at which it may be received. It is common for CA tone to be received at such a low level that only the (nominally) 200 ms burst of tone every 5 s is transmitted to the hardware, and all of the speech portion is lost.

(*d*) The only additional threshold duration counter required specifically for TNT1 is 1004 Hz ON. The duration of the ON pulse is nominally 1992 ms but 800 ms ON was chosen to accommodate a spurious break of up to 190 ms in the middle of the pulse. This relaxed definition of TNT1 is possible because the receipt of a prolonged burst of a 1 kHz signal from any other source is unlikely, and 800 ms provides an adequate time for a positive identification of TNT1.

The operation of the tone-recognition program is very simple. It consists of scanning sequentially the threshold duration counters, and evaluating the results according to the list of tone definitions shown in Table 1 until a match is found for which the set of threshold duration counter values all lie within their permissible ranges for a given tone.

PERIPHERAL TONE-IDENTIFICATION SOFTWARE

The way in which the tone-recognition program was integrated into the system design was of great importance. It was necessary for the program to be supported by a certain amount of peripheral software. A flowchart showing the functions of this external software is given in Fig. 3. The main purposes of the supporting software are as follows.

It was required to supplement, by software, the level of noise rejection provided by the hardware. Before the presence of a tone on the line is acknowledged it must be present for at least 170 ms, including the initial 60 ms delay contributed by the hardware. Bursts of tone of shorter duration than this are ignored. When a tone has been present for 170 ms, the software waits a further 1830 ms before examining it further; this delay is introduced because a test-number relay-set can take up to 2 s to return TNT1, and the intervening period of RT must be ignored.

It is usual for NU tone, NT and PT to be identified externally to the program. In fact, NU tone and NT are only handled by the program when they contain a level of corruption. After the 2 s waiting period, the program is started from the next ON or OFF transition received. However, under some circumstances, no further transitions will be received and the program will never be entered and, therefore, certain maximum waiting periods are set. If 400 Hz is received continuously for 3 s, then this signal is identified as NU tone. Similarly, if no 400 Hz or 1004 Hz signal is received for 5·5 s the absence of tone is identified as NT, provided that the NT timeout has expired. It is necessary to test for NT for at least 5 s because a period of silence of this duration is possible in a CA tone cycle. However, before NT is finally confirmed, a check must be made that the NT timeout has expired; this period is 20 s for own exchange and local calls and 30 s for STD type calls. Lastly, PT has its own special test because it is required to identify this tone in a period of 1 s. The test for the presence of PT is based on the number of pulses received in 1 s that are not more than 285 ms duration.

The software also performs the function of converting to pulses those frequency transitions detected on the line; these pulses are stored for retrospective analysis. The software then has to process these pulses to compensate for the distorting effect of the hardware digital-integration; to accomplish this it is necessary to extend the apparent duration of 400 Hz ON pulses by 44 ms and reduce the apparent duration of 400 Hz OFF pulses by 44 ms.

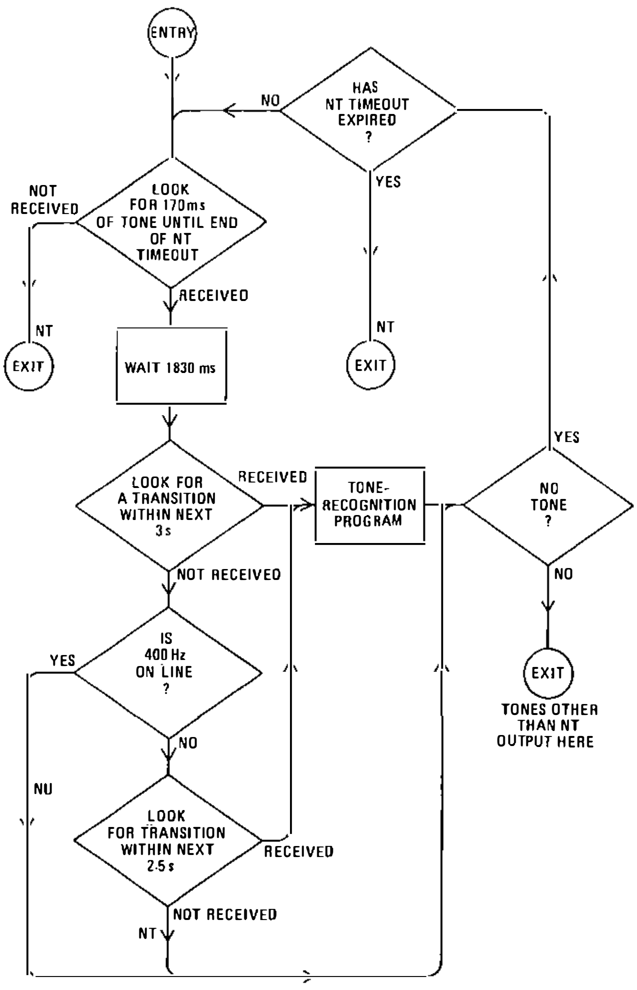


FIG. 3—Peripheral tone-identification software

CONCLUSION

The BPO is to provide a MAC in each of the 61 Telephone Areas in the UK. A few installations have been completed and are in service, and the method of tone identification described in this article has worked with great success in the field. It is considered that the method is sufficiently well developed to be applied directly in other telecommunications systems that are computer based and for which tone identification is required.

The decision as to when a tone is sufficiently far from the ideal cadence that it should be rejected is always subjective. The human ear has an amazing ability to cope with an incredible degree of variance from the ideal, and with very high levels of noise corruption. However, the tolerance ranges shown in Table 1 are generous and, in some cases, are as wide as possible before the representation of 2 tones overlap.

References

1 SAPSFORD, B. G. Measurement and Analysis Centres, Part 1—System Concept and Equipment Description. *POEEJ*, Vol. 70, p. 243, Jan. 1978, and Part 2—Processor Programming and Operation. *POEEJ*, Vol. 71, p. 28, Apr. 1978.

2 FRANCOIS, R. A., and HOLLIGON, E. R. Measuring Quality of Service on Trunk Routes. *POEEJ*, Vol. 60, p. 227, Oct. 1967.

3 MACKIE, R. H. An Electronic Automatic Call Sender for Testing New Telephone Exchange Installations. *POEEJ*, Vol. 61, p. 165, Oct. 1968.

Techniques for Measuring the Transmission Properties of Optical-Fibre Cables

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UDC 621.391.63: 677.521

This article describes the techniques used by the British Post Office to measure the transmission properties of optical-fibre cables, both in the laboratory and in the field. The measurements include the determination of loss, bandwidth and refractive-index profile. Fault location by the observation and measurement of backscattered light in the fibre is also described.

INTRODUCTION

This article is concerned with the techniques and test equipment involved in making precision measurements of the transmission properties of multimode optical fibres in both the laboratory and the field. One of the characteristic features of multimode fibre as a transmission medium is that its loss and bandwidth depend on the distribution of power among the many possible modes of propagation; a situation which contrasts with the coaxial-cable case and which requires that care must be taken to obtain measurements of practical relevance.

Measurements are required to determine the acceptability of fibres and cable after production; measurements are also required following installation in the field and, subsequently, for maintenance purposes. In the factory, the measurements include those to ensure that the desired refractive-index profile (and hence numerical aperture and core diameter) has been achieved, the minimum bandwidth requirement has been met and that the maximum loss limit has not been exceeded. In the field, loss measurements are the most useful in determining transmission performance, but the capability of measuring bandwidth may be required, particularly if there is a need to assess the possibility of increasing the traffic-carrying capacity of a link.

Test equipment, closely related to that used for bandwidth measurements in the time domain but employing pulse-echo principles, can be used to detect non-continuity of a fibre, to measure fibre length, and to observe backscattered light. The latter facility is a valuable technique for recording the losses and reflections within a fibre link by a non-destructive method which requires access to one end of the link only.

In describing the test equipment and measurement techniques used, this article is illustrated by reference to the British Post Office (BPO) Martlesham–Kesgrave–Ipswich optical-fibre cable installation[1,2,3] but, for the sake of completeness, is not confined to it.

LOSS MEASUREMENTS

The optical power launched into a fibre is attenuated due to 2 distinct loss-mechanisms, known as *absorption loss* and *radiative loss*.

Absorption loss occurs when optical power is absorbed by impurities within a fibre, notably, water and transition metal ions, and dissipated as heat. The temperature rise is very small (approximately 10^{-6} °C) and can be detected only by careful measurement[4].

Radiative loss, or scattering loss, occurs when optical power is scattered out of the fibre. There are 3 reasons for this:

(*a*) Inhomogeneities in the fibre which are small compared to the wavelength of light produce Rayleigh scattering both forwards and back.

(*b*) Optical power, particularly in the high-order modes, can tunnel through the cladding into the radiation field.

(*c*) Major imperfections in a fibre, or bends in a fibre, can cause coupling between the modes and, consequently, power is lost to the radiation field, either directly or indirectly, through the high-order modes.

To make a measurement of radiative loss alone is difficult: but, for system purposes, it is sufficient to know the *total loss*; that is, the sum of the absorption and radiative losses. Total loss is much easier to measure, and the measurement techniques are discussed in this article.

Total Loss Measurement

For ease of description, it is convenient to consider total loss measurements under 2 basic categories: those in which the loss is measured as a function of optical wavelength, typically, from the visible region of the spectrum (approximately 500 nm) into the infra-red region (1000–1500 nm); and those made at a single wavelength, which is usually chosen because it is the wavelength at which a system will operate.

Some measurement techniques and some components of test equipment are common to both types of measurement and, indeed, to many other measurements made on optical fibres. One important requirement is to determine the amount of power that has been launched into a fibre under test. The value of this power cannot easily be calculated because it is affected by the individual wavelength responses of all the components that go to make up the test equipment, by the quality of the end face through which power enters the fibre and, also, by the optical characteristics of the particular fibre under test.

After an initial measurement has been made, one way of eliminating most of the uncertainties relating to the determination of launched power is by cutting back the fibre under test to a length of about 1 m and repeating the tests. In this way, the test equipment response can be taken into account because the same end-face is used in both tests and, thus, the launched power is constant. A flat, perpendicular end-face can be prepared using a specially designed tool[6], although small irregularities are not important if the cut-back technique is used.

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Care must also be taken at the receiver to ensure that the measured power originates only in the core and is not due to ambient light or due to power propagating in the fibre cladding.

Immediately in front of the detector element of the receiver, the fibre under test passes through a liquid whose refractive index is greater than that of the fibre cladding. This removes the cladding modes which are bound by the air-cladding interface, whereas the true guided-modes, which are bound by the core-cladding interface, are unaffected. The immersion may be accomplished by laying the fibre across a plate whose surface is smeared with the index-matching liquid; alternatively, the fibre end may be inserted into a glass cell of the liquid which is placed immediately in front of the detector. The latter technique has the added advantage that the liquid in the cell has a refractive index much closer to that of the fibre core than air and this reduces any deviation or scattering of the light due to a slightly imperfect fibre end.

It has been stated in the introduction to this article that both loss and bandwidth are not single-valued parameters of a fibre, each being dependent on the mode distribution launched. It is therefore sensible to launch a distribution which corresponds closely to the stable mode distribution; that is, the mode distribution which is reached after a long length ($\geqslant$1 km) of fibre. The measurement will then be representative of the loss of a fibre when it forms part of a long jointed link and, hence, the losses measured on the individual fibres that make up the link may be added together to predict the total loss. This aspect is important because, if all modes are excited equally at the input, the loss measured on a 1 km length of fibre is typically 0·5–1·5 dB† greater than its loss when measured as part of a longer link. Considering a 10 km link, the predicted loss will be 5–15 dB greater than its measured loss. On step-index fibres, this problem has been solved by the use of a mode scrambler situated immediately after the launch point[7]. There are several types of mode scrambler; in all cases, they attempt to introduce a large amount of mode coupling (caused by microbending) in a few centimetres of fibre, thereby creating a mode distribution characteristic of a long length of fibre. The techniques that have been used include wrapping the fibre tightly around a drum and sandwiching the fibre between a pair of rough surfaces.

The effect of mode scramblers on graded-index fibres has been found to be less reproducible, therefore an alternative approach has been adopted that restricts the amount of power launched into the high-order modes. This is achieved by restricting the numerical aperture[8] (NA) at launch by using an aperture stop to ensure that the launched power is concentrated around the core axis; this is achieved by choosing suitable lenses to produce a demagnified image of the source at the core (see Fig. 1). The image is positioned on axis with the aid of an infra-red television camera, which views the

† Throughout this article, decibels (dB) refer to optical powers, and are defined as $10\log_{10}(V_1/V_2)$. The reason for using the factor 10 (rather than the factor 20) is that the voltage output from a photo-receiver is directly proportional to the incident optical power

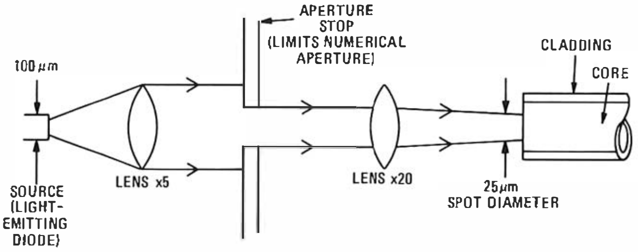
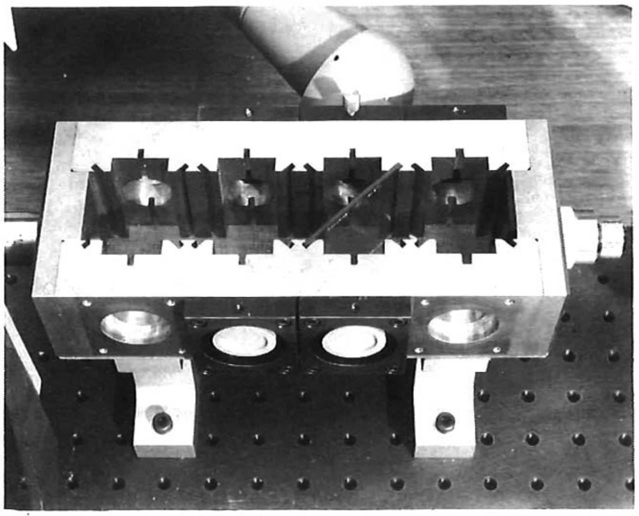


FIG. 1—The restricted launch conditions used to simulate stable mode distribution in graded-index fibres



Note: Microscope lenses screw into accurately aligned holes at each end of the block. The straight slots are used for filters, and the angled slots for the positioning of beamsplitters. The side ports can accommodate lenses, eyepiece units, sources and detectors for many different applications.

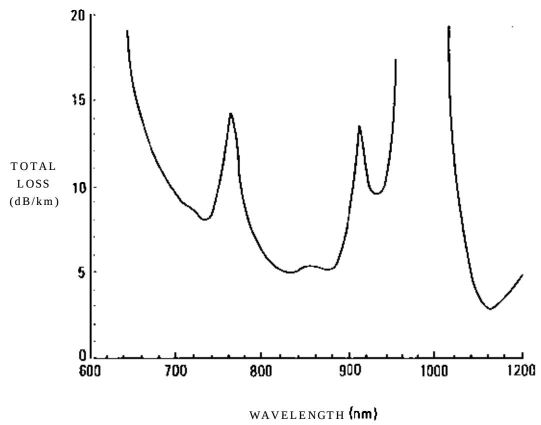
FIG. 2—A specially designed mount for housing optical test equipment

reflection (amounting to 4% of the incident light) from the fibre end via a beam splitter. The television camera is also used as an aid to ensure that fibre ends have been well prepared.

During a loss measurement, the power launched into a fibre must remain constant. This requires that a stable optical-mount be used to hold the lenses and beamsplitters, and the use of rigid micromanipulator stacks to adjust the positions of the source and the fibre under test. An example of this type of mount is shown in Fig. 2. Alternatively, a much simplified launching technique may be used which is more suitable for field measurements. This uses a source whose output is transmitted through a short fibre tail which is specially manufactured to have an NA and a core size smaller than the fibre type being measured, but having an identical cladding diameter. The fibre tail is butted in a groove against the fibre to be measured and a small drop of refractive-index-matching liquid is added at the interface. This is, effectively, a temporary joint which, when clamped, provides a very stable restricted launch without the complications of imaging optics and television monitoring.

Spectral Dependence of Fibre Loss

A measurement to determine the spectral dependence of fibre loss should be necessary only at the fibre production stage; this measurement is a useful means of identifying fibre impurities and other loss mechanisms. An example of such a measurement result, taken on a graded-index fibre made by a chemical-vapour-deposition (CVD) process, is shown in Fig. 3. The CVD process produces a fibre that is contaminated with a relatively large amount of water, the effect of which is indicated in Fig. 3 by the 3 peaks at 760, 910 and 980 nm. The peaks at 760 nm and 980 nm are overtones of the fundamental stretching vibration of the oxygen-hydrogen ion at 2700 nm, and that at 910 nm is a combination band with silica. The gradual decrease in loss with increasing wavelength is due to the λ^{-4} dependence of the Rayleigh scatter[9].



Note: This measurement result was obtained from one of the fibres used in the Martlesham–Kesgrave link.

FIG. 3—The loss spectrum of a silica fibre made by a CVD process

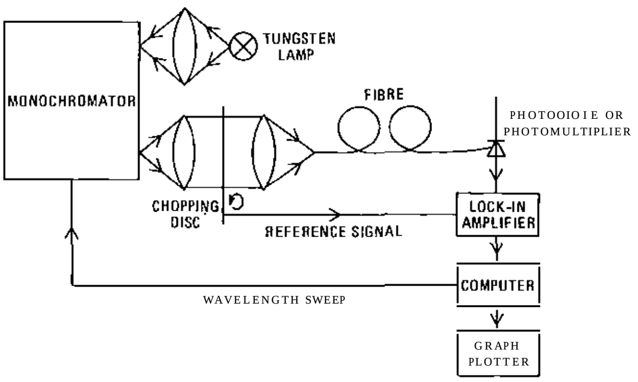


FIG. 4—Arrangement of test equipment used to obtain the loss spectrum of a fibre

A schematic diagram showing the equipment arrangement used to obtain the above loss spectrum is given in Fig. 4. Further details may be found elsewhere[10]. The input power source is a tungsten lamp which is imaged onto the entrance slit of a grating monochromator. The wavelength resolution is set at a few nanometres, thereby selecting a wavelength range narrower than any absorption features in the spectrum. The output is chopped mechanically, at a few hundred hertz by a rotating sector disc, and then launched into the fibre by means of an optical imaging system. The object of the chopping is to distinguish between the alternating received signal and any stray light at the detector. The modulation frequency signal emerging from the end of the fibre can be detected with either a photodiode or with a cooled photo-multiplier and rectified with a lock-in amplifier. The rectified signal is then recorded by a minicomputer, and the wavelength is stepped on by a motor drive connected to the diffraction grating. By this means, the optical output is recorded throughout the wavelength range of interest.

If L is the length of a fibre and $\alpha(\lambda)$ is its extinction coefficient at wavelength λ, the power transmitted is given by

$$P_\lambda(L) = P_\lambda(0) \exp\{-\alpha(\lambda)L\},$$

and the loss is given by

$$\frac{10}{L}\log_{10}\frac{P_\lambda(0)}{P_\lambda(L)} = 10\,\alpha(\lambda)\log_{10} e$$

$$= 4{\cdot}34\,\alpha(\lambda) \text{ decibels/kilometre.}$$

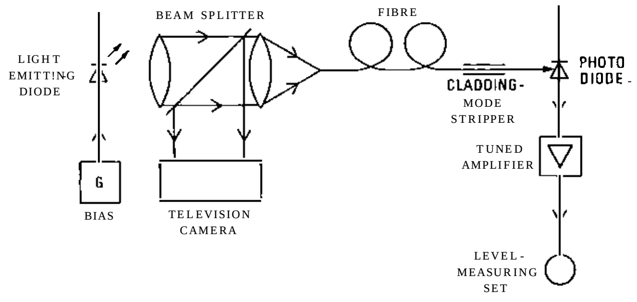


FIG. 5—Arrangement of test equipment used to make point loss measurements

Both the long-length and short-length measurement results, $P_\lambda(L)$ and $P_\lambda(0)$ respectively, are stored in the minicomputer throughout the wavelength range of interest and are used to calculate the extinction coefficient ($\alpha(\lambda)$) and to plot a graph of this parameter against wavelength.

Single-Wavelength Loss Measurements

At present, the semiconducting devices that are used as sources in optical communications systems operate in a small band of wavelengths (800–900 nm), but it is possible that, in the future, wavelengths of 1060–1300 nm may be used. Consequently, at this time, a design engineer is interested only in the loss at these wavelengths, and a more straightforward measuring technique may be used, as shown in Fig. 5. The source is usually a semiconducting device, either a light-emitting diode (LED), or a laser, whose output is modulated at a low frequency, typically 10 kHz. It is possible, but often less convenient, to use a tungsten lamp with wavelength selection provided by an optical filter but, in this case, a mechanical light-chopper is needed. An imaging system is used to launch power into a fibre and, at the receiver, detection is by a large area photodiode, followed by a tuned amplifier (for example, a level-measuring set) or a lock-in amplifier.

Loss Measurements in the Field

To simplify fibre loss measurements in the field, a small portable test set, which can be mains or battery powered, has been constructed. It has 3 main components: a transmitter, a jointing-jig and a receiver. The stabilized, modulated source contains a laser or an LED coupled to a fibre tail, which is chosen to generate a suitable mode distribution, as mentioned earlier. This fibre tail is jointed temporarily, using the jig, to the test fibre. The use of a modulated source, combined with a sharply-tuned receiver, makes it possible to measure fibre losses of up to 90 dB, corresponding to a received power of only 1 pW.

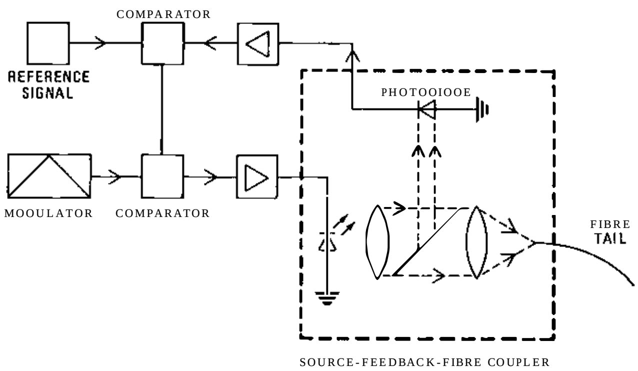


FIG. 6—Schematic diagram of the transmitter unit used for loss measurements made in the field

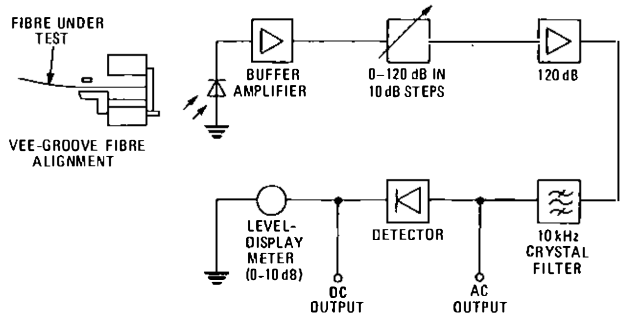


FIG. 7—Schematic diagram of the receiver used for loss measurements made in the field

TABLE 1

Measured and Predicted Losses of an Optical-Fibre Cable Route

Route Length (km)	Cumulative Loss (dB)	Predicted Loss (dB)
1·00	4·7	4·7
1·75	8·1	8·2
2·75	12·4	12·6
3·75	16·0	16·3
4·75	20·6	20·6
5·75	24·8	24·8

The transmitter contains a modulator and a feedback network, which controls the output power by comparing the feedback signal with a reference voltage (see Fig. 6). The optical coupler to the fibre tail contains a beamsplitter which redirects a small percentage of the light onto a photodiode in the feedback circuit. The overall stability is governed by the temperature stability of a photodiode (0·1% per °C) and the crystal oscillator frequency.

The receiver uses a large area photodiode, followed by a low-noise buffer amplifier (see Fig. 7). To enable subsequent stages to operate within a limited dynamic range, an attenuator is inserted before the main amplifier and the narrow-band (1 Hz) crystal filter. After detection, the signal is displayed on a meter, the power being the sum of the meter readings and the attenuator settings. Stability is limited by the photodiode to 0·1% per °C, as before.

For field measurements, where remote working is necessary, the lock-in amplifier used in the laboratory is unsuitable as a detector because of its need for a reference signal. The test method also implies the use of 2 receivers, one for the test fibre and one for the short-length measurement. To overcome the problem of different sensitivities, a stable and portable source has been used to calibrate the test equipment at each end of a link immediately prior to taking a measurement. However, during the testing of the Martlesham–Kesgrave–Ipswich optical-fibre transmission system, it was found that the need for a calibration source could be avoided by making measurements in both directions. The operators at each end of the cable transmit and receive alternately on the same fibre, and each performs a short-length measurement. It may be shown that, provided the fibre loss is the same in both directions and the test equipment is stable, the calibration errors cancel out, thus permitting the loss to be calculated.

When attempting to assess the accuracy of loss measurements, it is important to remember that the result depends on the conditions of measurement and that 2 laboratories will not, except by chance, reach the same conclusion about the loss of the same fibre unless they use identical launch conditions. Thus, reproducibility of results is a very important consideration since, if repeated measurement of the same fibre always leads to the same answer, that answer can be considered accurate for the launch condition in use. In the laboratory, where both ends of the fibre are available in one place, the reproducibility of loss measurements is within ±0·1 dB. In the field, with all the complications of remote working, the spread is ±0·2 dB, using either the calibration source or the 2-way method.

The effectiveness of the restricted launch technique is illustrated in Table 1, which shows the measured losses at each kilometre stage of a jointed 6 km link and a running total of the individual measurements. In Table 1, the cumulative loss is the measured total loss for the length specified, and the predicted loss is the sum of the measured individual losses of the fibres forming the link up to that point. The measurements were obtained on channel 1 of the 140 Mbit/s Martlesham–Kesgrave cable. Similar agreement was obtained on the other channels. Bearing in mind that the summation would have been 3–9 dB larger than the measured loss at the 6 km stage if care had not been taken over the launching conditions, the agreement is very satisfactory.

MEASUREMENT OF BANDWIDTH

The bandwidth of a multimode optical fibre is limited principally by 2 factors: mode dispersion and material dispersion. Unlike coaxial cables or metallic waveguides, multimode fibres support 500–1000 modes. Each mode has a different propagation velocity and, if power is launched into many modes of a fibre in the form of a short pulse, the output pulse will be spread in time. Also, in contrast with coaxial transmission systems, the carrier signal from the light source does not have a single characteristic wavelength, but covers a small range which greatly exceeds the modulation bandwidth. This range is normally termed the *linewidth* of the source, and is most conveniently specified as the width, in nanometres, at half peak amplitude. Thus, for a semiconductor laser, the width can be 2–4 nm, whereas for an LED it can be 40 nm. This source linewidth leads to the second factor which limits bandwidth, namely material dispersion. The velocity of light in a medium depends on the refractive index, which is a function of wavelength. Thus, the different wavelengths emitted by the source travel at different velocities, causing additional pulse-broadening in proportion to the length of a fibre. This contribution can be calculated if the material-dispersion coefficient of the fibre and the linewidth of the source are known.

The total pulse-broadening due to mode and material dispersion can be calculated by summing the squares:

$$\sigma_{\text{TOTAL}} = \sqrt{(\sigma^2_{\text{MODE}} + \sigma^2_{\text{MATERIAL}})},$$

where σ is the RMS broadening of a very narrow input pulse. The measurement of bandwidth can be carried out either in the time domain (that is, using pulses), or in the frequency domain (that is, continuous wave).

Frequency-Domain Measurements

In the frequency domain, an LED or laser light source is modulated sinusoidally and the resulting signal is launched into the fibre under test. The received level is recorded for a range of frequencies, thus giving the combined frequency response of the fibre and the test equipment. To remove the test-equipment response, the fibre under test is cut back or replaced by a short length and the measurement is repeated. This second response is subtracted (on a decibel scale) from the first, thus producing a result that is dependent only on the characteristics of the fibre under investigation. In bandwidth measurements, the fibre response is quoted relative to zero frequency.

In general, both amplitude and phase are a function of base-

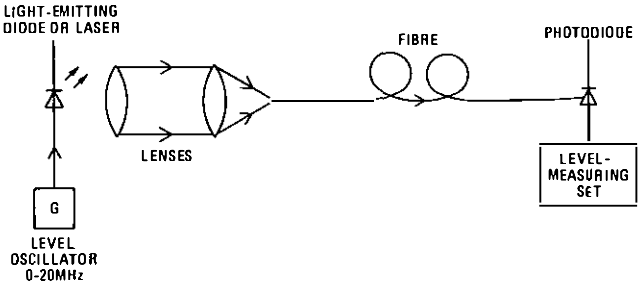


FIG. 8—Arrangement of test equipment used for bandwidth measurements in the frequency domain

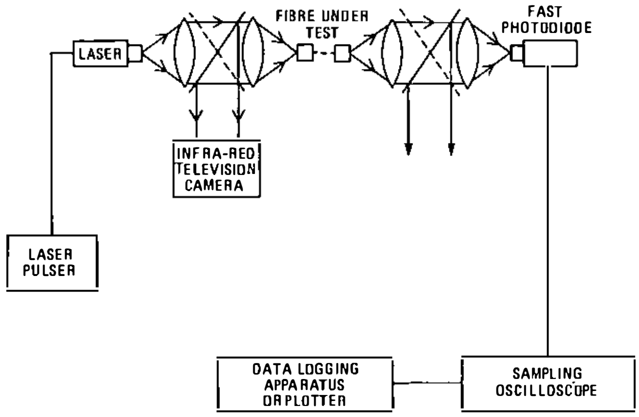
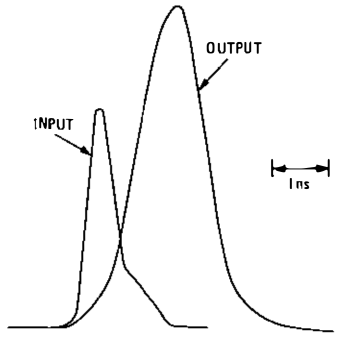


FIG. 9—Arrangement of test equipment used for bandwidth measurements in the time domain



Note: The amplitude scales for the pulses are arbitrary and different

FIG. 10—A typical input pulse (generated by the test equipment shown in Fig. 9) and the corresponding output pulse at the end of 2 km of fibre

band frequency so that, to define fully the transmission properties of a fibre, both characteristics need to be measured. However, in the frequency domain, the measurement of phase response is difficult because of the large phase difference between the signals at the ends of the fibre, therefore the test equipment used at present is restricted to the measurement of amplitude/frequency response only.

An earlier section of this article described the nature of the launched power distribution needed to obtain meaningful loss measurements applied to jointed sections. The bandwidth of a fibre is also dependent on the launched power distribution. Consequently, it is important to ensure that conditions are repeatable so that a valid comparison can be made between different fibres or, for example, the performance of one fibre can be monitored through various stages of a cable installation process.

The arrangement of test equipment required for bandwidth measurements in the frequency domain is shown in Fig. 8. The test equipment includes an LED which is coupled to the fibre under test by a lens and modulated by a signal from an oscillator. The transmitted light is received by an avalanche photodiode whose gain is stabilized against temperature effects. The signal power is detected by a level measuring set (scale calibrated in decibels). Using this arrangement, the bandwidth that can be measured is typically 20 MHz, and a loss of 40 dB can be tolerated. These limits can be extended by using a laser source in conjunction with a spectrum analyser, possibly up to 1 GHz, but, in general, frequency-domain measurements are little used at present.

Time-Domain Measurements

To measure the bandwidth of a fibre in the time domain, a short light pulse is launched into the fibre under test and a broadened pulse is detected at the far end. The test fibre is then cut back or replaced by a short length and a transmit pulse is recorded. This pulse may be regarded as the input to the test fibre because the effect of the receiving equipment on the width of the pulse is the same for both measurements. Hence, any further broadening is attributable to the test fibre alone. In the time domain, the results are usually presented in terms of the impulse response function (IRF) which can be measured directly only by using an input pulse of vanishingly small width, known as a *delta function*. The IRF can be computed by deconvolving the measured input pulse from the output pulse of the fibre on test. To do this, the amplitude/frequency and phase/frequency responses are calculated for the input and output pulses by Fourier transformation, the former is divided into the latter, and the resulting response is inversely transformed to yield the IRF. Both presentations, the amplitude/frequency response and IRF are useful. One important advantage of the time-domain method is that the phase information is obtained using a simple design of test-equipment; this is possible because the receiver is triggered by the incoming pulses, not by the original signal. Consequently, the equipment is not required to absorb the relatively-long transit time of the pulse which is equivalent to the large phase difference in the frequency domain.

The test equipment used to make pulse measurements is shown in Fig. 9. The source is a single heterostructure gallium arsenide laser, which produces a peak power of 2 W in a pulse duration of less than 300 ps. This output is coupled into the fibre under test by a pair of microscope objectives between which is a beamsplitter. This allows the light reflected from the fibre end to be observed with an infra-red television camera so that the laser junction (a stripe) can be focused consistently on the fibre diameter. The output pulse is received by a photodiode and displayed on a sampling oscilloscope from which it is transferred to punched tape to facilitate computation. This arrangement can be used on fibres having losses up to 65 dB and bandwidths up to 1 GHz. Typical input and output pulses are shown in Fig. 10, and the corresponding frequency, phase and impulse responses are shown in Fig. 11.

A straightforward extension of the time-domain method is the shuttle-pulse technique, which is useful for investigating the bandwidth of short lengths of fibre where the pulse dispersion is small. In this method, a semi-transparent mirror is placed at each end of the fibre and the signal is launched through one mirror and allowed to travel backwards and forwards by successive reflections at the ends. A photodiode, which, with suitable optics, may be at either end of the fibre under test, detects the fraction of the pulse power passing through the mirror each time it reaches the end. By sampling at the correct time, the broadened pulse may be examined after the desired number of transits, thus effectively increasing

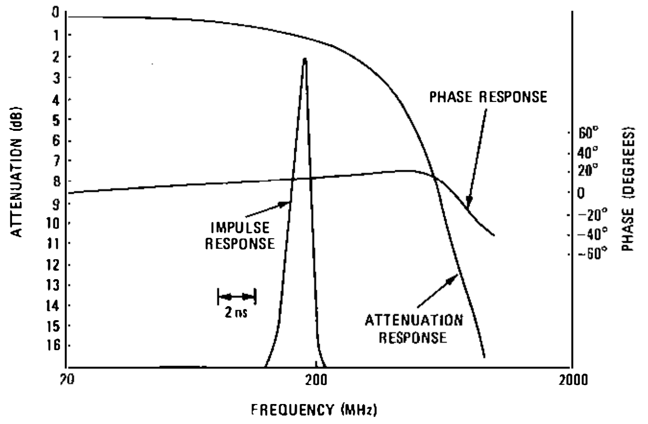
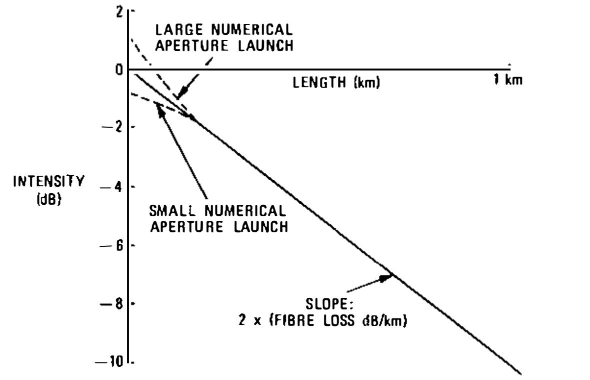


FIG. 11—Frequency, phase and impulse responses (computed from the pulses shown in Fig. 10)



Note: When the launched power distribution differs from the stable mode-distribution, the intensity approaches the stable mode case over a distance which is dependent on the strength of mode coupling in the fibre

FIG. 12—Plot illustrating the intensity of backscattered light as a function of length for various launching conditions

the length of the fibre. In one experiment, 9 passes along a 1 km length of fibre were observed. Care must be taken that the mirrors do not disturb the propagating modes.

BACKSCATTER MEASUREMENTS IN OPTICAL FIBRES

Rayleigh scattering in optical fibres has already been mentioned as a source of loss. It is caused by microscopic variations in the refractive index of the glass, and its local intensity is a characteristic of the material rather than the fibre. However, when the intensity of the scattered light is observed from the end of a fibre it is affected by the fibre properties. The observation point can be the source end since scattering occurs in both directions.

At each point in the fibre, the amount of backscattered light is proportional to the signal power at that point. Therefore, because the signal power diminishes along the fibre, the backscattered light intensity also diminishes and its measurement gives a guide to the fibre loss.

To determine the amplitude of the backscattered light from a particular point along the fibre, the signal is launched in the form of pulses, and the distance to the scattering point can be deduced from the time delay. For a fibre having a constant loss, the signal power decreases exponentially with length. Hence, a plot of the logarithm of the backscattered power as a function of length (or time) is a straight line, the slope of which gives the fibre loss in decibels per kilometre. In the section of this article that dealt with the loss characteristics of fibres, the concept of a stable mode distribution was discussed. If the launched power distribution is different from this stable mode distribution, the loss of the first section of the fibre can be either lower or higher than the steady-state result. Fig. 12 shows the effect of different NA launch conditions on the backscatter result and indicates that the steady-state condition is approached as the fibre length increases. This means that, when using the backscatter technique for determining the loss of a fibre, it is not necessary to launch the stable mode distribution because it is evident when this condition exists, provided the fibre is of sufficient length.

Backscatter Measuring Equipment and Test Results

One of the problems associated with the measurement of backscatter is that the reflection from the launch end of a fibre is considerably greater (typically 15–20 dB) than the backscattered light. This reflection may saturate either the photodiode or the following amplifiers and limit the dynamic range over which the system can be operated. This problem can be solved either by gating-out the unwanted reflection[11], or by eliminating it. The latter approach has been adopted, and the measuring arrangement is shown in Fig. 13. The test equipment consists of a single heterostructure laser, driven at 20 A for 100 ns at a repetition rate of 8 kHz. The amplitude of the received backscatter is proportional to the product of the pulse amplitude and the pulse duration. However, the resolution in distance is dependent on the pulse duration, and a compromise must be made between these 2 factors. A 100 ns pulse width produces a backscatter signal of about 30 dB below the incident power while still giving a resolution of about 10 m. The pulse is launched through 2 lenses and a beam splitter into the fibre, which is held in a special launching cell. The cell is filled with a liquid having the same refractive index as that of the fibre core. The pulse of light refracts into the fibre, which is now index matched, and the unwanted signal is reflected harmlessly away. The backscattered light is reflected by the beamsplitter through a lens and onto the photodiode. The resulting signal is amplified and detected by a boxcar receiver, which can recover the signal from its accompanying noise, and is then displayed on an X-Y chart recorder. As mentioned earlier, it is an advantage to have a logarithmic plot of the result and a logarithmic amplifier can be included to achieve this. Using this test equipment, a dynamic range of approximately 50 dB can be achieved. Since the light must travel in both directions, the permissible fibre

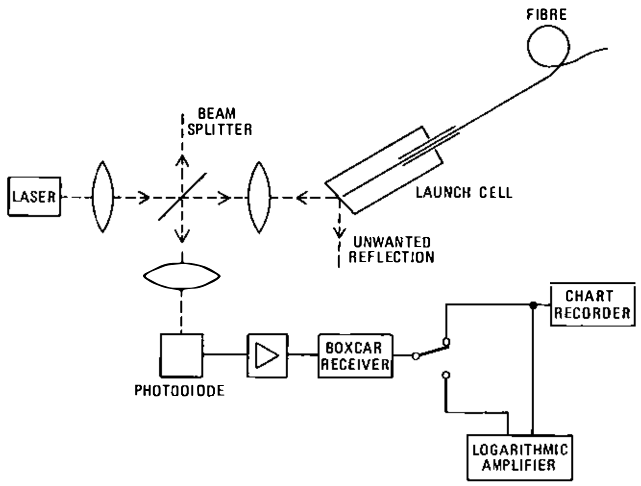
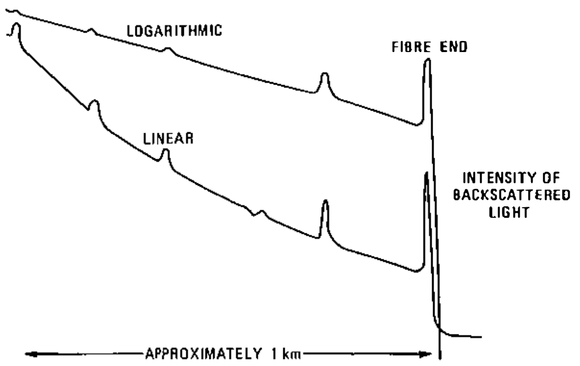
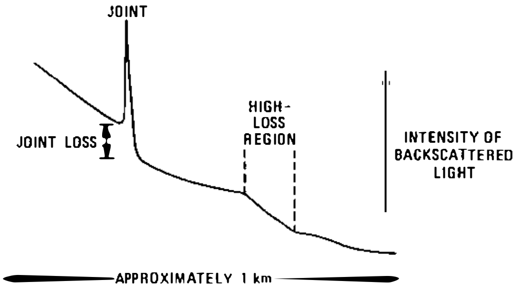


FIG. 13—Arrangement of test equipment used for measurement of backscatter



Note: The peaks are reflections caused by discrete scattering centres

FIG. 14—Linear and logarithmic plots of the intensity of backscattered light measured on 1 km of fibre on-drum



Note: Associated with this joint is a reflection, shown by the peak, and a loss shown by the sharp drop

FIG. 15—The backscattered light intensity observed from 2 fibres connected by a poor joint

loss is about 25 dB, which corresponds to a range of 5–6 km in present designs of fibre.

The backscatter measurement result of 1 km of fibre is shown in Fig. 14. The logarithmic plot is almost a straight line, thus indicating that the power distribution launched was not too different from the stable mode distribution and, also, that the fibre has a constant loss along its length. The peaks in both the plots are due to discrete scattering centres which, in the case illustrated, give rise to negligible loss. In this example, the fibre end was placed in index-matching liquid, which considerably reduced the amplitude of the reflection from this point. Another example of backscatter measurement, obtained from 2 fibres connected by a poor joint, is given in Fig. 15; the position of the joint is revealed by a reflection and subsequent loss of signal. The fibre following the joint has an interesting characteristic in that part of its length has a higher loss than the other, and this is shown by an increase in the slope of the graph.

BREAK DETECTION

The position of a break in an optical fibre can, in principle, be determined by measuring the time taken for a pulse of light, reflected from the discontinuity, to return to the transmitter. However, should the fibre end happen to be immersed in a liquid with a similar refractive index, the amplitude of the reflection will be greatly reduced. A further reduction can be expected in the likely event of the broken end not being flat. However, from Fig. 13, it can be seen that, even if the reflection is eliminated, the break can still be located by the cessation of backscattered light from the body of the fibre.

REFRACTIVE-INDEX PROFILE MEASUREMENT: GRADED-INDEX FIBRES

The mode dispersion in a fibre is critically dependent upon the refractive-index profile of the graded core. Consequently, detailed knowledge of this shape is necessary and a great deal of effort has been devoted to its determination. The measurement is difficult because the core is small (a typical radius is 30 μm) and the maximum refractive-index difference to be measured is about 0·01. In this case, most of the conventional techniques for refractive-index measurement are unsatisfactory, and new approaches are necessary.

One technique depends on measuring the power accepted by a fibre: either by measuring the power trapped inside the fibre[12], or by measuring the power refracted outside the fibre[13]. Both methods involve scanning a focused spot of laser light across the core of a fibre. In the first method, the power output from a short length of fibre is measured as a function of the input spot position. It can be shown[14] that the power $P(r)$ accepted at distance r from the centre of a multimode fibre is given by

$$\frac{P(r)}{P(0)} = \frac{n^2(r) - n^2(a)}{n^2(0) \quad n^2(a)},$$

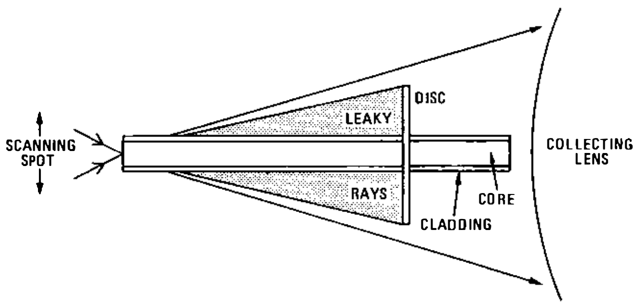
where a is the radius, n the refractive index, $n(0)$ is the axial index and $n(a)$ is the cladding index. For the small index differences exhibited by optical fibres, this expression reduces to

$$P(r) \propto \{n(r) - n(a)\},$$

showing that the power accepted at radius r is proportional to the local index-difference. This relationship only holds if all modes are excited equally. Unfortunately, this cannot be achieved without exciting a class of modes described as *leaky.* These modes gradually leave the fibre over distances which may range from millimetres up to a kilometre or more, and necessitate a correction which makes an otherwise simple technique difficult and time consuming.

This problem is avoided in the second technique shown in Fig. 16. Here, the scanned spot is arranged to overfill the NA of the fibre, which is immersed in a liquid of slightly higher refractive index than that of the cladding. This results in the presence of a hollow cone of refracted rays outside the fibre.

The leaky rays are largely confined to the inner part of the cone and can easily be removed by a disc placed on the axis. The higher-angle rays in the outer part of the cone are collected by a lens and focused onto the detector. It can be shown that this power is proportional to $n^2(r)$ and the detector output gives a direct measure of the refractive-index distribution, without the need for a leaky-mode correction. A typical refractive-index profile is shown in Fig. 17. The central



Note: In practice, the end of the fibre near the collecting lens is turned away so that the transmitted light is not observed

FIG. 16—Arrangement of test equipment used to determine refractive-index profile

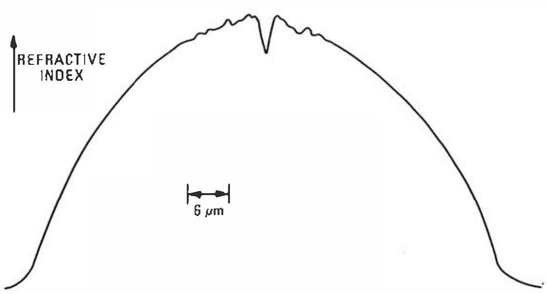


FIG. 17—A typical refractive-index profile of a silica CVD fibre

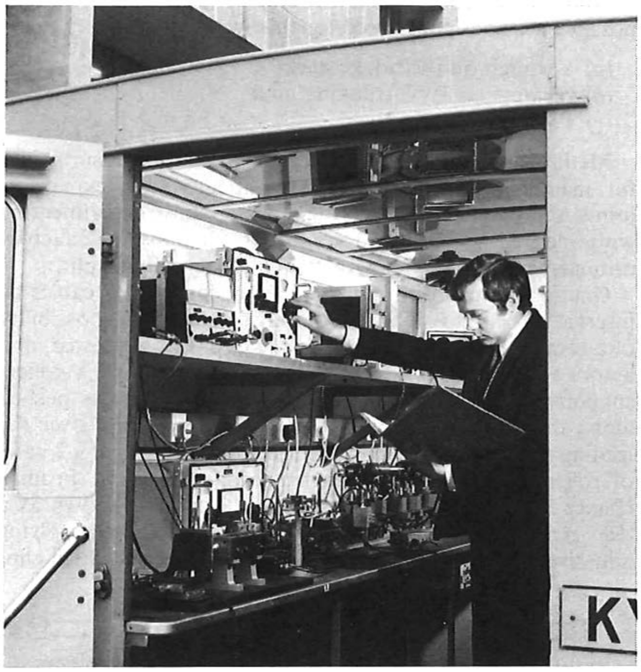


FIG. 18—View through the rear door of one of the test vehicles

dip may vary in depth from fibre to fibre. This is a feature of fibres made by the CVD process, and is due to evaporation of dopant while the preform is being collapsed.

At present, these measurements are made periodically in the laboratory for diagnostic purposes. With production fibres, such measurements will probably be confined to quality assurance laboratory tests.

TEST VEHICLES

Many of the measurements described in this article have been made along the roadside as well as in the BPO Research Centre. Principal among these are the determination of fibre loss and bandwidth. Two test vehicles have been equipped as mobile laboratories suitable for carrying out detailed investigations into the transmission properties of installed optical-fibre cables.

The vehicles, one based on a forward-control Land Rover chassis and the other on a Leyland 30 cwt utility, are fitted out with benches, heating and lighting. Power is provided by 2 kW petrol-driven generators carried in purpose-built trailers. Because, at the time of fitting out, no ruggedized optical equipment was available, the apparatus used was almost identical with the laboratory equivalent except that, where a choice existed, the unit with lower power consumption was preferred. Most of the equipment is bolted to the benches so that it is ready for work to begin shortly after arrival on site. Fig. 18 shows a view through the rear door of one of the vehicles.

Between Martlesham and Ipswich, a 20-pair copper cable was drawn in along the duct route of the fibre cable and fitted with connectors at each fibre jointing point. Loudspeaking telephones were used for communications between the vehicles and to the terminals. A direct exchange line was also available. In other circumstances, radios are used but these are much less convenient because they occupy at least one hand of the test operator.

After nearly 2 years and an appreciable mileage, the test equipment has proved surprisingly reliable and robust. No serious failure has occurred. It is expected that further equipment will be fitted in due course to provide more sophisticated diagnostic facilities. At the same time, portable, battery-operated equipment is being developed for routine measurement work and use outside Research Department.

CONCLUSION

This article has described methods for determining the most important transmission properties of optical fibres. Most of the test equipment used has been of a laboratory standard but, with the experience gained during many measurements in the field, work is proceeding on the prototypes of simple and reliable test-sets for use in future optical-fibre cable installations.

ACKNOWLEDGEMENTS

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References

[1] BRACE, D. J., and RAVENSCROFT, I. A. Optical-Fibre Transmission Systems: The 8 Mbit/s Feasibility Trial. *POEEJ*, Vol. 70, p. 146, Oct. 1977.

[2] BERRY, R. W., and RAVENSCROFT, I. A. Optical-Fibre Transmission Systems: The 140 Mbit/s Feasibility Trial. *POEEJ*, Vol. 70, p. 261, Jan. 1978.

[3] BRACE, D. J., and HENSEL, P. C. Optical-Fibre Transmission Systems: The Martlesham-Kesgrave-Ipswich Optical-Fibre Cable Installation. *POEEJ*, Vol. 71, p. 54, Apr. 1978.

[4] WHITE, K. I. A Calorimetric Method for the Measurement of Low Optical Absorption Losses in Optical Communication Fibres. *Optical and Quantum Electronics*, Vol. 8, p. 73, 1976.

[5] REEVE, M. H., BRIERLEY, M. C., MIDWINTER, J. E., and WHITE, K. I. *Optical and Quantum Electronics*, Vol. 8, p. 39, 1976.

[6] HENSEL, P. C. A Simplified Optical Fibre Breaking Machine. *Electronics Letters*, Vol. 11, p. 581, 1975.

[7] EVE, M., HILL, A. M., MALYON, D. J., MIDWINTER, J. E., NELSON, B. P., STERN, J. R., and WRIGHT, J. V. Launching Independent Measurements of Multimode Fibres. Proceedings 2nd European Conference on Optical Fibre Communication, p. 143, Paris 1976.

[8] DYOTT, R. B. The Optical Fibre as a Transmission Line. *POEEJ*, Vol. 67, p. 164, Oct. 1974.

[9] BEALES, K. J., MIDWINTER, J. E., NEWNS, G. R., and DAY, C. R. Materials and Fibre for Optical Transmission Systems. *POEEJ*, Vol. 67, p. 80, July 1974.

[10] BRAIN, M. C. Techniques for Measuring Total Loss of Optical Glass Fibre Waveguides. BPO Research Department Report No. 572, 1976.

[11] PERSONICK, S. D. Photon Probe: An Optical Fibre Time Domain Reflectometer. *Bell System Technical Journal*, p. 355, Mar. 1977.

[12] SUMNER, G. T. A New Technique for the Refractive-Index Profile Measurement in Multimode Optical Fibres. *Optical and Quantum Electronics*, Vol. 9, p. 79, 1977.

[13] STEWART, W. J. A New Technique for Measuring the Refractive-Index Profiles of Graded Optical Fibres. Integrated Optics and Optical Communications, Japan, 1977.

[14] GLOGE, D., and MARCATILLI, E. A. J. Multimode Theory of Graded Core Fibres. *Bell System Technical Journal*, Vol. 52, p. 1563, 1973.

Fitting PVC Ducts on In-situ Cables

D. F. PARSELL and T. E. MILLS†

When Norfolk County Council decided to renew the deck of Roudham Heath Bridge, which carries the A11 trunk road over the Ely–Norwich railway line, they started a chain of events which could produce consequences of great importance when the diversion of telecommunications plant is necessary because of road works. The bridge carries four 92 mm bore earthenware ducts: part of a 38 m section between 2 surface jointing boxes.

The Council planned, originally, to construct a new bridge and approximately 1·7 km of new carriageway adjacent to the existing bridge and carriageway. On completion of the new bridge, the existing carriageway was to be closed in order to reconstruct the deck of the old bridge, the structure of which was to be retained. It was agreed with the Council that the 4 ducts could be suspended for the 18 m between the bridge abutments while the redecking was carried out. The new section of carriageway and the new bridge were constructed according to plan and the road traffic diverted.

About 2 months before redecking was to commence, the Council decided that it would be necessary to renew completely the old bridge. This meant that the duct crossing the bridge would need to be suspended over a span of some 60 m, which included the whole 38 m section of 4-way duct and approximately 11 m of 3-way multiple duct and single duct of the 2 adjacent sections. The Council time scales did not allow work to be deferred to give time to divert the telecommunications plant to the new bridge. At a meeting between interested parties, it was decided that a suitable supporting structure could be erected to support safely the plant over the whole 60 m span.

When the ducts over the bridge were uncovered, they were found to be encased in concrete. Since the concrete surround could not be supported, the concrete and duct were broken away carefully, leaving 7 main and junction cables exposed. The 2 jointing chambers were demolished and the 11 m of duct of the adjacent sections were strapped to a supporting member. A wooden box, lined with expanded polystyrene 75–150 mm thick, was constructed around the whole length of exposed cables and supported by constructing a framework of L-angle iron and scaffold poles around it. This was then suspended on a catenary between 2 supporting towers, enabling the contractor to proceed with the task of demolishing the old bridge and constructing the new one.

A decision had to be taken on how to restore protection to the cables once bridge rebuilding was complete. The possible methods considered for this were:

(*a*) a trough on the bridge deck,
(*b*) rectangular PVC trunking, and
(*c*) PVC duct cut in 2.

Methods (*a*) and (*b*) were rejected because of unsuitability for maintenance and because of difficulty with expansion joints and bending at the bridge exits. Whilst experimenting with method (*c*) it was found that the most satisfactory arrangement was a duct with a single longitudinal split.

One further problem remained; how could the 7 cables be inserted into split PVC duct without damage? The possibility of a tool to insert the cables into the duct was suggested, and a tool was designed which consisted of 2 sets of V-rollers supported horizontally at each end of a bar. It was pushed along the split in the duct, with the cable passing over the leading pair of rollers, down into the duct between the 2 pairs of rollers and under the trailing pair (see Fig. 1). From a chance remark during testing, the tool became known as a *duct zip*. It had been intended to seal the split by applying adhesive and fitting securing clips at 1 m intervals. Such clips

† Cambridge Telephone Area.

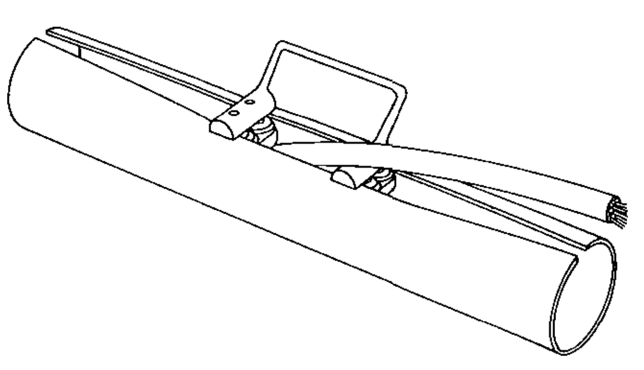


FIG. 1—The duct zip, showing cable insertion

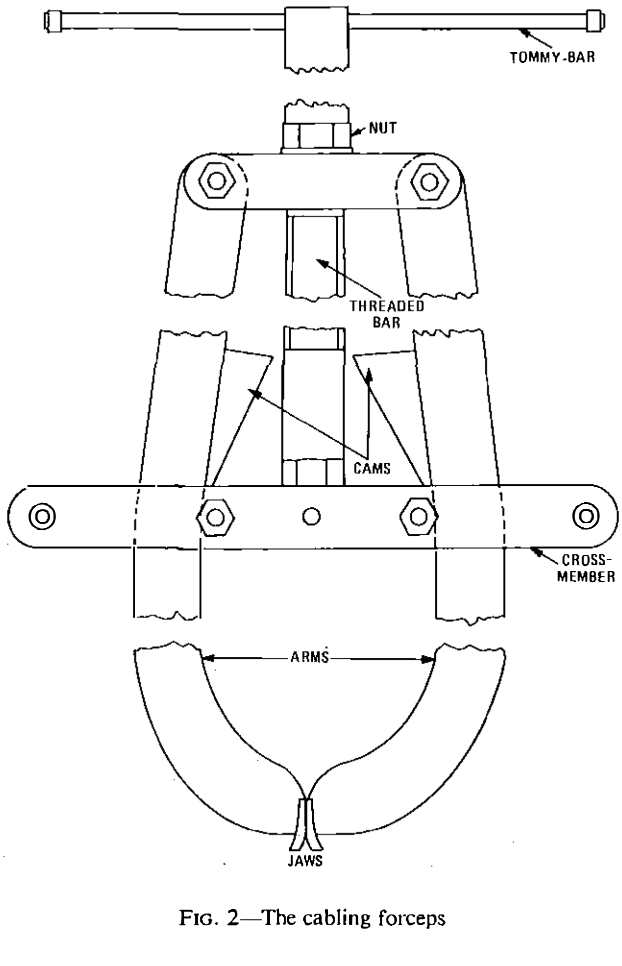


FIG. 2—The cabling forceps

arc made by cutting a longitudinal strip about 50 mm wide out of a length of the same type of duct and then cutting the duct into 60 mm lengths. However, following a suggestion by Telecommunications Headquarters, it was decided that the sealing should be carried out with a specially-produced extruded H-section PVC strip, which proved a much easier and cleaner method. The operation at Roudham Heath, using these methods, proved a complete success.

There was another bridge renewal which, if the method used at Roudham Heath could be employed, would show considerable saving over conventional methods by avoiding changing over some 10 400 cable pairs. However, the cables involved were 72 mm in diameter, compared with 41 mm at Roudham Heath, the latter being the maximum which could be accommodated by the duct zip. A second tool was therefore developed: the *cabling forceps* (see Fig. 2).

Three or four cabling forceps are required to insert a large cable into a split PVC duct and the method of use is shown in Fig. 3. The cable is supported over the duct with the split uppermost; the first pair of forceps are opened by hand (Fig. 3 (*a*)), placed over the cable, closed (Fig. 3 (*b*)), and the jaws on the ends of the arms are inserted into the split. When the tommy-bar is rotated, the screw action raises the cross-member up the cams, which separate the jaws (Fig. 3 (*c*)). The procedure is repeated with a second pair of forceps approximately 450 mm along the duct in the direction of insertion. A third and, if necessary, a fourth pair, of forceps is placed along the duct at further 450 mm spacings until the cable drops into the duct at the position of the first forceps. The forceps are then moved along the duct in a leap-frogging manner until the cable is completely inserted into the duct (Fig. 3 (*d*)).

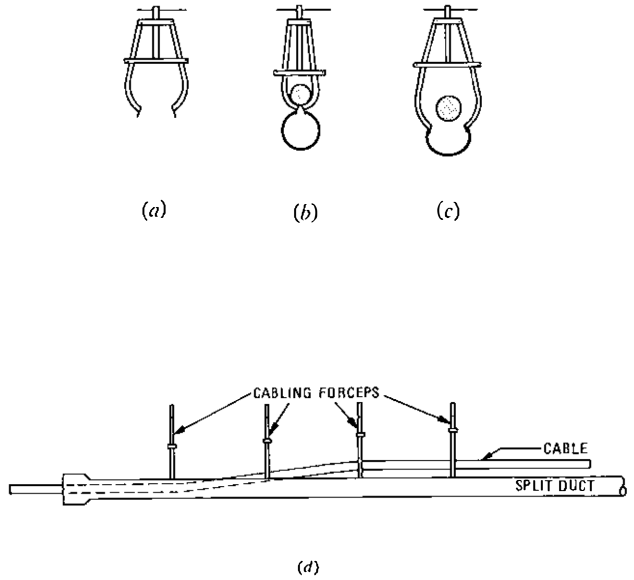


FIG. 3—Cable insertion using forceps.

The principle of both tools is that the cable passes into the duct, in the slot that traverses its length, under the action of the tools. The duct zip is fast in operation but is limited to smaller cables; the cabling forceps are slower to use but can deal with any cable size. Both tools have been used many times, and development is proceeding to make them available for general use.

Book Review

Beginners' Guide to Audio. I. R. Sinclair. Newnes-Butterworths. 184 pp. 105 ills. £2·75.

The book is presented in 7 essentially self-contained chapters, each of which encompasses one broad aspect of sound recording and reproduction suitably divided under sub-headings. After a brief introduction to simple acoustics, Chapter 1 describes the mechanisms involved in recording sound, and includes the principles of microphones (with some examples), and the principles of disc and magnetic recording. Chapter 2 deals with the first stage of reproduction, and covers turntables, pick-up cartridges, cassette replay, and broadcast reception. This chapter also has a useful résumé of the problems associated with reproduced sound, especially noise and distortion. A short account of impedances and matching is included. The next 3 chapters are concerned with the amplifying chain: voltage amplifiers, equalization (tone controls and filters), and output stages. Chapter 6 examines the problems and limitations of loudspeakers and, with the help of several photographic examples, describes the various types of loudspeaker systems in use. The final chapter summarizes the various systems in use: 2-channel-disc, tape-and-radio, discrete and matrix 4-channel working, Hafler, and a description of Dolby and dbx noise-reduction methods. A somewhat sketchy index completes the book.

The description of voltage amplifiers in Chapter 3 contains too much unrelated detail and unnecessary mathematics, and is unsuccessful in a book of this type. It would have been better written in a similar manner to the chapter on output stages, in which the descriptive text is well supported by skeleton circuit diagrams used as examples.

With the exception of Chapter 3, the book provides a sound descriptive background on most aspects of sound reproduction, and almost lives up to its title of a "beginners' guide" which requires no previous knowledge of electronics or audio; the scope of the subject is perhaps a little too ambitious to be entirely successful in this attempt. The accuracy of the text is generally good, although there are a few errors (for example, the description of the need for pick-up side-thrust correction is incorrect), and some inconsistencies (for example, while one chapter rejects integrated circuits for high-quality audio systems because they are noisy, another chapter does so because of problems of consistent production, and nowhere are other problems, such as transit time, mentioned). The text is not always well served by the accompanying diagrams, some of which are quite confusing; the photographic illustrations, however, are very useful.

The author avoids the pitfall of being influenced by present fashion, while providing at the same time a précis of most current trends, although the omission of any mention of digital methods is disappointing.

I. H. B.

National Synchronization Reference Clock

W. BUNN, B.SC.†

UDC 621.395.74: 621.395.345: 621.395.66

In an integrated digital transmission and switching network, the timing of the multiplexing and switching functions must be controlled to a great accuracy. The British Post Office will adopt a synchronized method of control for its proposed network of digital transmission and switching systems. This article describes the operation of the reference clock that will be used as an absolute reference standard for synchronization of the UK integrated digital network.

INTRODUCTION

In the British Post Office (BPO) proposed integrated network of digital transmission and switching systems, the transmission and switching rates will be governed by the frequency of quartz crystal oscillators, contained in the secured timing units (TUs) of the digital-switching exchanges. To ensure that the mean frequency of the exchange TUs is the same, it is proposed to synchronize them via the 2·048 Mbit/s digital line links, as described in a previous article.[1] Unfortunately, the natural frequency of a quartz oscillator drifts away from its nominal value as the oscillator ages; thus, the frequency of a network of synchronized oscillators will drift unless the network is locked to a frequency source that has an acceptable margin of accuracy and reliability.

The CCITT* recommendations[2] on plesiochronous working over digital international routes require that the absolute frequency of a national network shall be maintained within ± 1 part in 10^{11}; if recommendations are made with regard to international synchronization, a high-accuracy frequency reference will be required at a digital international switching exchange in each country.

To meet the CCITT recommendations, the BPO has decided that its reference frequency source will be provided by a caesium-beam primary-frequency standard. This can be achieved by locking the network to one or more caesium standards located at main switching centres, or by using the off-air frequency standards provided by the National Physics Laboratory (transmitted from the BPO radio station at Rugby).

Among the disadvantages of off-air transmissions are that the short-term stability is poor (particularly at sunrise and sunset) and that the transmissions are not continuous. The main disadvantages of a caesium-beam frequency standard are poor reliability and a long mean-time-to-repair (the latter because the caesium standards are of foreign manufacture and their complexity of design requires them to be returned to the manufacturer for repair). However, it was decided to provide the UK network with caesium-beam frequency standards and to overcome their poor reliability by replication.

OPERATION OF A CAESIUM-BEAM FREQUENCY STANDARD

All clocks work on the basis of counting a regular periodic phenomenon. In the case of atomic frequency standards, certain atoms undergo transitions between well-defined energy states and, as a result, they emit or absorb energy in the form of an electromagnetic wave whose oscillation frequency depends upon the structure of the atom and the difference in energy between the 2 states concerned. The element commonly used is caesium, which is a soft yellow element with a melting point of 28·4°C.

In 1967, the Thirteenth General Conference of Weights and Measures related this emission or absorption to absolute time, and so adopted the atomic time scale. The second was defined as, *the duration of 9 192 631 770 periods of the radiation corresponding to the transition between the 2 hyperfine levels of the ground state of the caesium 133 atom.*

The basic unit of a frequency standard of this type is the caesium-beam tube, which selects the desired transition and, at the same time, minimizes any external environmental effects on the selection mechanism[3,4]. A block diagram of a typical caesium-beam tube is shown in Fig. 1.

A small reservoir of caesium, heated to about 100°C, forms a jet of atoms which is focused into the air gap of a selecting magnet. The atoms behave like tiny magnets, and their orientation is dependent on their energy state. The state-selector magnet is positioned to deflect the atoms that are in the required energy state into a cavity, where they are excited by an electromagnetic field created by an injection frequency. The electromagnetic field induces some atoms into other energy states and a second state-selector magnet deflects into a detector those atoms that have undergone the required transition. The detector output current is proportional to the number of atoms that have undergone the required transition in energy, and is at a maximum when the injection frequency corresponds exactly to the desired transition of the caesium 133 atom. To separate the various possible transitions between the magnetic hyperfine levels, the cavity is placed in a uniform magnetic field (which is screened from the influence of external fields, either terrestrial or parasitic).

The output current from the caesium-beam tube is used in a control loop to control the frequency of a high-quality quartz oscillator from which the injection frequency is derived. A block diagram of the control loop is shown in Fig. 2. In addition to providing a 5 MHz output signal, the quartz oscillator also drives a synthesizer and a modulator. The synthesizer produces a 12·63 MHz signal from the

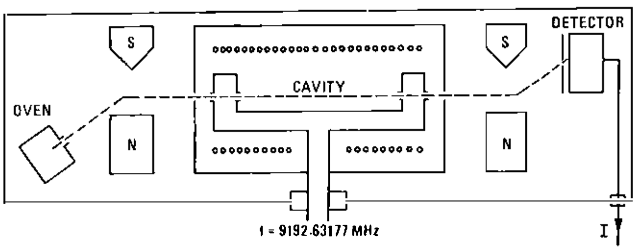


FIG. 1—Block diagram of caesium-beam tube

† Research Department, Telecommunications Headquarters

* CCITT—International Telegraph and Telephone Consultative Committee

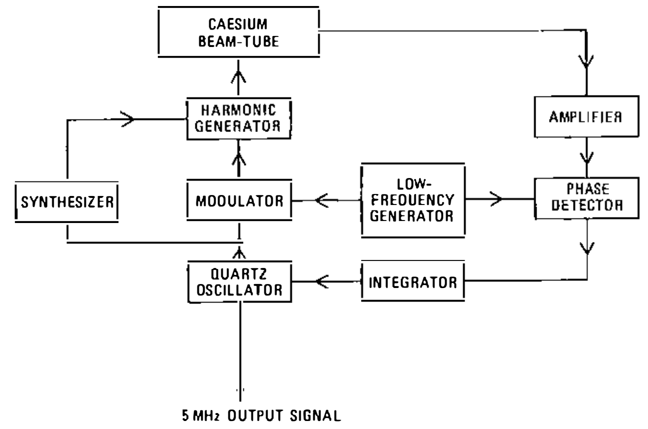
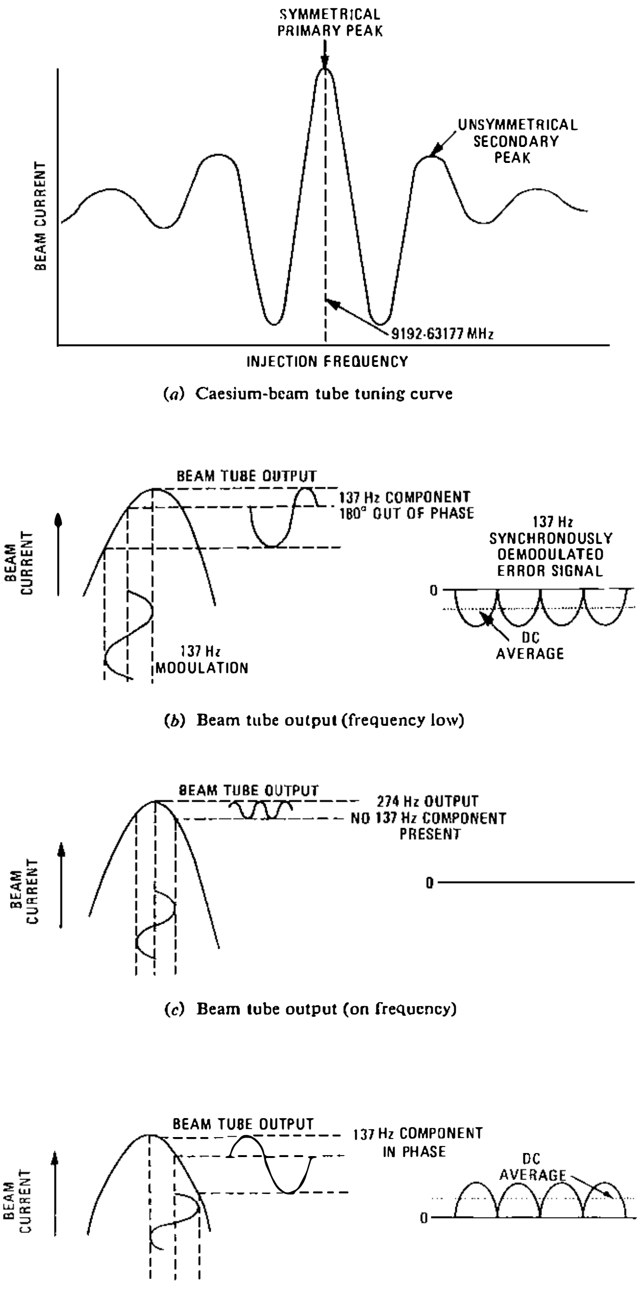


FIG. 2—Block diagram of a caesium-beam frequency standard



(*a*) Caesium-beam tube tuning curve

(*b*) Beam tube output (frequency low)

(*c*) Beam tube output (on frequency)

(*d*) Beam tube output (frequency high)

FIG. 3—Relationship between caesium-beam current and injection frequency

5 MHz source. The modulator adds 137 Hz frequency modulation to the 5 MHz source. The outputs from the synthesizer and modulator are multiplied and mixed to give the required 9192·63177 MHz caesium-beam tube injection-frequency, which is frequency modulated at 137 Hz. The characteristics of the caesium-beam tube are such that the beam current has maximum and minimum values depending upon the applied injection-frequency, and these variations are shown in Fig. 3(*a*).

When the caesium-beam tube input signal is exactly at the caesium reference frequency, the 137 Hz modulation on the input signal produces a second-harmonic component of 274 Hz on the caesium-beam tube output current, as shown in Fig. 3(*c*). If the injection frequency is above or below the peak caesium resonance frequency, the beam current output contains 137 Hz components, as shown in Figs. 3(*b*) and 3(*d*). The amplitudes and phases of these components are used to determine the correction needed to control the original 5 MHz oscillator to zero frequency error. The beam current is filtered into its 2 components and amplified. The amplified 137 Hz component is synchronously demodulated in the phase detector to produce an average DC signal, whose amplitude is proportional to the amount of frequency error and whose polarity indicates whether the frequency is high or low. This error voltage is applied to an integrator circuit, which produces a control voltage to maintain the 5 MHz quartz oscillator precisely on frequency.

The caesium-beam tube output-current shown in Fig. 3(*a*) has a symmetrical primary peak with unsymmetrical secondary peaks on either side. The output current is monitored and, if the control loop is operating on a secondary peak, the DC balance is destroyed and an alarm signal raised. Additional monitoring circuits check continuously for correct operation of the control loop, and virtually all single faults can be detected and alarmed.

SECURITY

Having established a requirement to maintain the absolute mean frequency accuracy of the BPO synchronized network to 1 part in 10^{11} by using caesium standards as a reference source, the problem of reliability arises. The operational departments of the BPO have stated that the mean-time-between-failures (MTBF) of the frequency reference should be in excess of 100 years. However, performance data from the manufacturers of caesium standards indicate that the MTBF of a single caesium-standard is of the order of 5 years. Therefore, more than one standard unit will be required to achieve the desired reliability. A further problem arises in that, since a caesium-beam frequency standard is a highly sophisticated piece of equipment and because the quantities required within the BPO network are small, it is probable that they will be sent back to the manufacturer for repair. This will result in a long mean-time-to-repair (MTTR) because both manufacturers are foreign and have repair facilities in Switzerland for UK instruments. Including the reliability testing period required by the manufacturer after the repair, a MTTR of 4 weeks or more could result.

These considerations indicate that at least 3 caesium standards will be required in the network. There are several ways in which the caesium standards could be configured in the network:

(*a*) Each caesium standard could be located in a different digital switching centre serving a particular region of the BPO network. The network could then be split into 3 synchronous regions connected by plesiochronous links. In the event of any one caesium standard failing, synchronization links in the region served by that standard would need to be reconfigured to synchronize to either of the other regions. This arrangement would require fairly-complex automatic, or manual, change-over facilities.

(*b*) The outputs of the caesium standards could be synchronized via digital links to form a completely synchronous network; this would require the development of special synchronization equipment and reliability would be dependent on the continuity of the links.

(*c*) All 3 caesium standards could be sited in the same exchange and secured locally. The network would then have only one reference, which would have a high MTBF. The disadvantage of this method is that there is no protection against catastrophic loss of the reference.

The BPO has decided to adopt the method whereby all 3 caesium standards will be situated in the same exchange, and a standby reference will be provided elsewhere (for use in the rare event of the main reference being destroyed). This method avoids the development and provision of a special synchronization equipment, or equipment necessary to reconfigure the network. The 3 caesium standards are to be built into a secured reference-clock, which will provide an output to maintain the TU of the parent exchange at the correct frequency and thus provide an absolute frequency reference for the national network. The caesium standards will be used in a *work* and *stand-by* mode, with triplicated change-over switches feeding triplicated links into the synchronization utility (SU) and the TUs of the digital exchange. This method of working will ensure that the reference clock has a MTBF well in excess of 100 years.

REFERENCE CLOCK

The basic elements of the reference clock are shown in Fig. 4. The caesium-beam frequency standard provides the primary reference frequency of 5 MHz. However, the synchronization system requires a reference frequency at the fundamental digital line rate of 2·048 MHz, or a multiple thereof. A frequency synthesizer is therefore used to generate an 8·192 MHz square wave, which is locked to the 5 MHz output from the caesium-beam frequency standard using a phase-locked loop. The outputs from the 3 frequency synthesizers are connected to each of 3 change-over switches; each change-over switch can connect one of the working inputs to a 32-channel frame generator and a link-control unit. The change-over switches monitor their own inputs and, if the selected input fails, they automatically switch to the next working input. Change-over can also be initiated by alarm circuits within the caesium-beam frequency standard, the frequency synthesizer, or the frequency comparator. The inputs to the change-over switch are not phase locked and, at change-over, a maximum phase discontinuity of 61 ns can occur in the outputs. The 8·192 MHz signal is divided down to 2·048 MHz; the effect of a phase discontinuity error on the divided waveform is that the next transition after change-over can be up to 61 ns early or late with respect to where it would have been had a change-over not been necessary. Each change-over switch monitors the other 2 change-over switches and, if it is not working on the same output as the others, its output is inhibited. A manually-operated switch is provided to enable the output from each change-over switch to be selected manually. If a change-over switch fails, this manual facility allows a replacement change-over switch to be brought into step with the 2 other change-over switches.

Because 3 change-over switch outputs are each connected to a 32-channel frame generator and a standard link-control unit, it is not necessary for the outputs to be phase-locked, although they normally are. The frame generator uses the 2·048 MHz signal to generate a pulse-code modulation (PCM) signal that is in accordance with the CCITT recommended

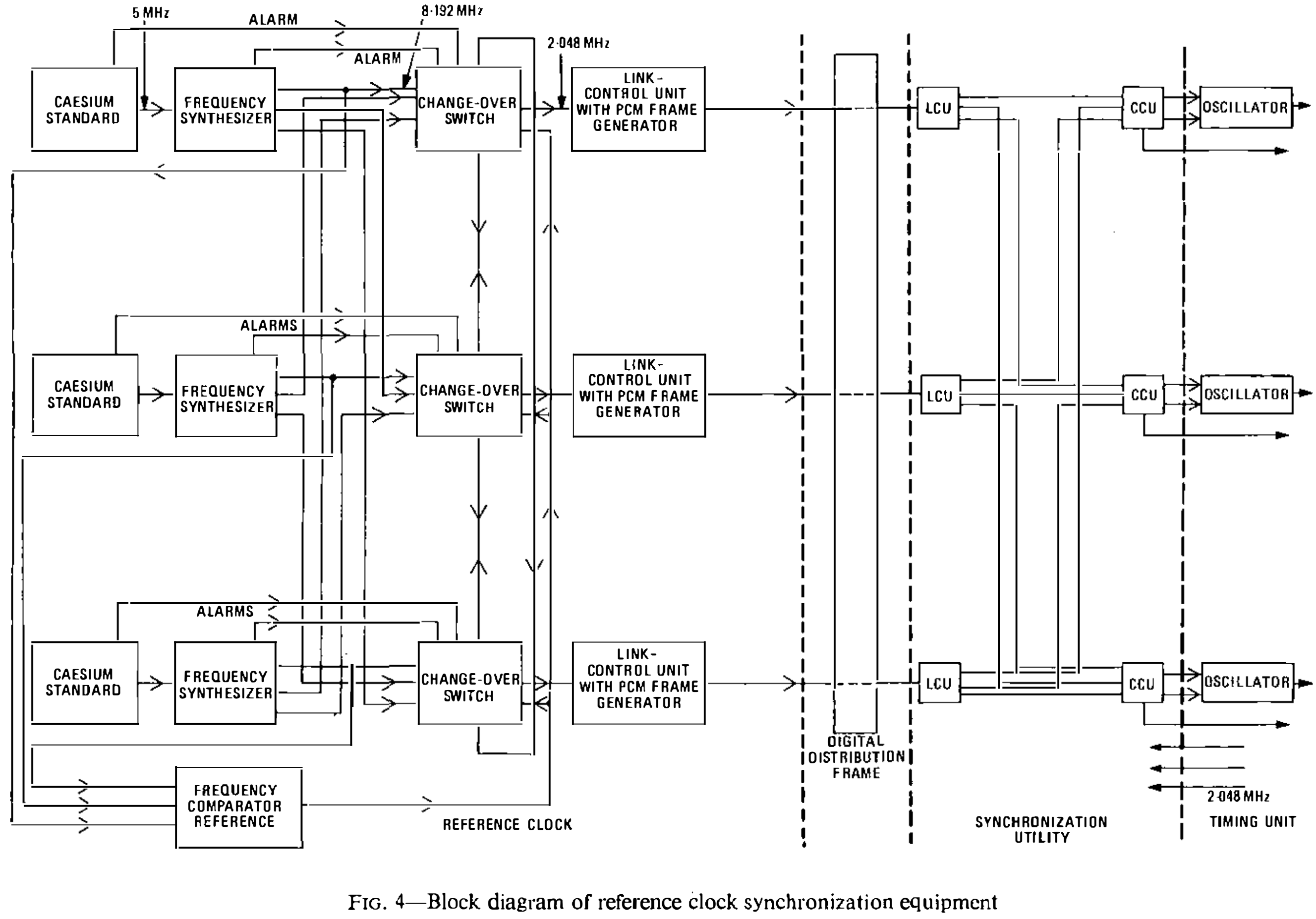


FIG. 4—Block diagram of reference clock synchronization equipment

format. This signal is applied to a standard link-control unit, replacing the signal that would have come from a digital switch in a non-reference exchange. This link-control unit interworks with an identical unit associated with the exchange SU (see Fig. 4). At the SU, the only difference between the links from the reference clock and those normally going to the SU are that they do not carry traffic and, therefore, they do not pass to the digital switch. Instead, they are connected to a digital-switch simulator in the SU, which generates the 2·048 Mbit/s PCM format necessary to provide the framing signal. The synchronization control links between the reference link-control units are unilateral and are the only effective links controlling the exchange TU. A detailed description of how the link-control units interwork to derive control signals for a TU has been given in a previous article[1] in this *Journal*.

The reference-frequency comparator compares the outputs from the 3 frequency synthesizers. An alarm signal to the change-over switches is raised if the frequency difference between any one output and the other 2 outputs exceeds one part in 10^{11}; this provides a further check to that carried out within each caesium standard.

THE EFFECT OF REFERENCE CLOCK FAILURE

Failure of the reference clock will have little effect upon the synchronization of the network other than to cause a small phase change in the TU signal output. However, assuming that the frequency of TU oscillators will vary with age and that such variations are randomly distributed, in the worst case when the reference-exchange TU is scheduled to be retuned (that is, when it is 0·6 Hz off frequency), a total reference failure could cause 50% of exchanges to experience an oscillator change-over. If all 3 oscillators of a particular TU age in the opposite direction to that at the reference exchange, then 2 consecutive change-overs will be experienced because none of the oscillators will be within 0·6 Hz of the new reference. However, the exchange should not lose synchronism because the SU has a control range of 2 Hz, but all oscillators will be indicated as faulty and a prompt alarm will be raised. The effect of multiple change-overs can be reduced if a third caesium-standard failure is anticipated by retuning the reference-exchange TU oscillators more often once 2 caesium-standard failures have occurred. In this case, when all 3 caesium oscillators have failed and the TU becomes the reference for the network, it will be nearer the absolute frequency.

Following a reference clock failure, a second disturbance in the system occurs when the reference clock is restored; the effect on the synchronization of the network will depend upon the length of time that the reference clock has been out of commission. The time taken for the frequency of TU oscillators to drift 0·6 Hz is about 15 months. Assuming a random distribution of the ageing effect of TU oscillators, then in 4 weeks (which is the time it is anticipated it will take to repair a caesium standard), even if all the TU oscillators, age in the same direction, 7% of the TUs will experience a change-over when the reference is reinstated and approximately 2% will have activated a prompt alarm, since all oscillators will require retuning. But, again, synchronism should be maintained. The effect of reinstating the reference clock can be reduced considerably if, as has been suggested, the network is reconfigured to be synchronized to a standby reference when it is known that the main reference will be out of service for an extended period.

For international working, the maximum recommended slip† rate is 1 slip per 70 d. On failure of the reference clock, the slip rate may increase to 1 slip per 6·4 min (when the reference exchange TU is 0·6 Hz off frequency); this represents a maximum frequency inaccuracy of 3 parts in 10^7. This inaccuracy will increase while the reference clock is out of commission, due to the ageing of the TU oscillators.

† Slip—An aberration of the timing processes associated with the transmission or switching of a digital signal resulting in an unwanted loss or gain of a time slot or of a set of consecutive time slots

PERFORMANCE OF INTERNATIONAL CIRCUITS

International links will operate in a plesiochronous manner with a frequency inaccuracy not greater than 1 in 10^{11} and slips will occur at international boundaries, even when the national networks are synchronous. The maximum rate at which these slips will occur over any one link is once every 70 d.

It has been suggested[5] that the effect of slips on customer service is tolerable provided that the slip rates do not exceed 300 slips per hour for telephony circuits and 3·5 slips per hour for data applications. It is obvious that there are several orders of magnitude in hand before any intolerable effect would be noticed. Even under rare fault-conditions when the reference exchange in the UK drops back to an inaccuracy of 3 parts in 10^7, slips occur only every 6·4 min.

When the effects of these slips are compared with other transmission impairments which cause errors, they appear insignificant as they contribute an error rate better than 1 in 10^{10} over an hypothetical reference circuit, compared with a design objective of 1 in 10^5.

Despite these comparisons, it is necessary to ensure that the design of the reference clock will permit the introduction of international synchronization at a later date in the eventuality of a significant improvement of the other parameters affecting service to customers. At the CCITT, both the German administration and the BPO have put forward proposals for mutually synchronizing the reference clock frequencies based upon their respective national proposals, should the requirement arise.

Since it is not desirable to control the frequency of the caesium standards themselves, the BPO proposals include the use of a phase microstepper between the outputs of the caesium standards and the frequency synthesizers; this would enable the phase of the 5 MHz output to be advanced or retarded at regular intervals under the control of an international SU. Suitable maximum parameters would be a phase change of 2 ns every 100 s to control a possible inaccuracy of 1 part in 10^{11} in the reference frequency.

SUMMARY

A description of the proposal to refer the national synchronous network back to a reference clock containing caesium atomic standards has been given. This article has also included a brief description of the caesium atomic standard, the reference clock and associated security considerations. The effects caused by the failure of the reference clock have been discussed both with regard to the national and international networks.

References

1 BOULTER, R. A., and BUNN, W. Network Synchronization, *POEEJ*, Vol. 70, p. 21, Apr. 1977.
2 CCITT Orange Book, Vol. III, Recommendation G811.
3 Caesium-Beam Frequency Standard Model 3200—Operating and Instruction Manual, Oscilloquartz, SA.
4 Caesium-Beam Frequency Standard Model 5061A—Operating and Service Manual, Hewlett Packard Ltd.
5 CCITT Contribution—Effects of Synchronization Slips, COM IV, No. 102-E, Annex 3.

The Associate Section National Committee Report

ANNUAL CONFERENCE

The annual conference took place at the Technical Training College, Stone, on Saturday, 20 May. We were very happy to have in attendance as an observer Mr. R. Farr, the new secretary of the Institution.

NATIONAL TECHNICAL QUIZ FINAL

The 1977–78 National Technical Quiz final took place at the Institution of Electrical Engineers on 28 April. The finalists this year were the Technical Training College (TTC), Stone, and London Telecommunications Region, East Telephone Area. After an exciting battle, the TTC drew away and ran out winners by 39½ points to 31½. The Bray Trophy was presented to the captain of the winning team, Mike Haynes, by Mr. J. S. Whyte, Senior Director of Development of the British Post Office and President of the IPOEE.

NATIONAL PROJECT COMPETITION

Because of the lack of entries for the proposed Project '78 Competition, *A Novel Use for Alternative Energy Sources in the Post Office*, the General Purposes and Finance Committee decided that a more realistic project should be put forward. Therefore, the new project is: *Design a Telephone Combining the Maximum Number of Facilities for Use by Disabled Customers*. The instrument must be recognizable as a telephone and should need no add-on parts. Anyone interested in entering the competition should contact the project organizer as soon as possible, with a letter of intent, to ascertain whether he is still able to accept entries. The final date for models of instruments to be available for adjudication is 30 November 1978.

PRESENTATIONS OF AWARDS

On the occasion of the National Quiz final, presentations were made of the Cotswold Trophy, awarded annually to the centre adjudged to have best furthered the aims of the Associate Section of the Institution, to the Colwyn Bay Centre and of the E. W. Fudge Trophy, awarded annually to the winners of the project competition, to the Sheffield Centre for their film *Make Someone Happy*. The film was then shown and thoroughly enjoyed by the audience.

M. E. DIBDEN
General Secretary

Associate Section Notes

AYR CENTRE

The Centre was re-established last year by volunteers who accepted the various offices on a temporary basis until an annual general meeting could be convened. An enthusiastic recruiting campaign was quickly undertaken to put the Centre on a firm footing and membership has trebled during the year. Following the annual general meeting, when officers were elected on a more permanent basis, the Centre affiliated to the National Committee and is maintaining close contact with the other Scottish centres.

Centre activities have included a lecture on *Video Tape-Recording* by a representative of Phillips Electrical and an ambitious programme of visits to such diverse establishments as Killoch colliery where, at 620 m below the surface, the method of long-wall working was shown, and Hunterston nuclear power generating station. A tour of a container ship at the Greenock terminal and another of the Volvo truck assembly plant at Irvine also proved most interesting to members.

R. GLEN

BOURNEMOUTH CENTRE

During the 1977–78 session, the response to some of the visits and activities has not been as great as may have been hoped. Some of the interesting new venues visited included the new ambulance centre at Winchester and, also at Winchester, the Hampshire police gave us a most interesting and enjoyable evening when we visited their headquarters.

A visit to the Motor Insurance Repair and Research Centre at Thatcham included an interesting talk about motor insurance and what is being done to achieve the more economical repair of vehicles and greater efficiency in carrying out this work. This was followed by a tour of one of the most up-to-date vehicle workshops in the country and, at a later date, by some film material from the Government's Road Research Centre. A visit to Plessey, at Poole, was of interest to all sections of the membership, with products for telecommunications and postal mechanization being of special interest.

The quiz team were successful up to the Regional final where, after a close and interesting contest, a half-point difference kept the Regional trophy from us.

At the annual general meeting, the following officers and committee were elected for 1978–79.

Chairman: J. Dymott.
Vice-Chairman: B. Fielder.
Secretary: B. Smith.
Treasurer: J. Hancock.
Assistant Secretary: M. Newman.
Committee: J. V. Woodland, D. Y. Kent-Smith, S. Jeffrey, M. B. Smithers, R. H. Ough, J. Kelsall and T. R. Keyts.

B. SMITH

COLWYN BAY CENTRE

The twenty-fifth anniversary meeting of the Centre was held on 14 April at the Royal British Legion Club, Llandudno. This was the most convenient date, although the actual founding date was 19 January 1953. The subject chosen was the aircraft *Concorde*. The lecture was given by British Airways' staff: Mr. R. A. R. Wilson on the first 2 operational years of the aircraft, with aspects of navigation and communication, and Mr. D. V. Pettit presented films, including a fine new production, *Supersonic Achievement*, which handled the progression of development of aircraft up to Concorde very well. The meeting was attended by 140 members, from both the Senior and Associate Sections, and guests. The vote of thanks was given by the President, Mr. K. E. Stotesbury. A focal point of the meeting was an exhibition called *ITT and Art*, being a selection of art works commissioned by the International Telephone and Telegraph Corporation, since art is communication in its highest form. A commemorative booklet was produced for the occasion in which there were contributions from several previous speakers of note. Concorde was the obvious choice for a cover symbol, with the hope, too, that the Centre might have a bright future.

Last year, the Centre was adjudged to have been the centre best furthering the aims of the Associate Section. As a consequence, I was honoured to receive, on behalf of the Centre, the Cotswold Trophy. A fitting climax to our first 25 years.

E. DOYLERUSH

EDINBURGH CENTRE

Our November meeting was a talk from Mr. S. Duncan, Technical Officer, Edinburgh Telephone Area, on *Stored Program Control* (SPC). Using a working exchange, we were given a practical demonstration of the facilities of SPC, and the members present maintained a steady stream of questions. Only 6 members and one visitor attended our December visit to Anderson's Woollen Mill, at Galashiels. We were treated to a colourful display of the construction of our famous Scottish tartans and were shown how the labour force has been reduced since new type looms were introduced.

Our first lecture of 1978 was a combined talk from Mr. F. M. Willison, Head of Maintenance (Field and Exchange) Division, and Mr. J. Brien, Head of External Planning Division, both from the Edinburgh Telephone Area. They described their own duties and responsibilities, and explained many aspects of financial control. This was an excellent talk and 38 members were present. Requests have already been received for further talks of a similar nature.

Seventeen members were treated to a very descriptive talk on *Metal Oxide Silicon Transistor (MOST) Type Translators* by Mr. G. W. Applebe from Scottish Telecommunications Board headquarters. With the aid of slides, our speaker explained the facilities of MOST translators, and his obvious enthusiasm for the equipment was passed on to all present. This talk was well enjoyed and was highly commended.

J. L. M. ALEXANDER

GLASGOW CENTRE

The 1977–78 session in now behind us, and we enjoyed a full programme.

October 1977 *External Planning and Works* by Mr. A. Scott, Head of External Planning Division, and Mr. A. McNab, Head of External Works Division, both from the Glasgow Telephone Area.

November 1977 *The Changing Pattern of the Post Office* by Mr. A. Kerr, Deputy General Manager, Glasgow Telephone Area.

December 1977 *The History and Work of the Ordnance Survey Department* by Mr. D. Toft, Ordnance Survey Department, Edinburgh.

January 1978 Visit to Black and White Whisky bottling plant, Stepps.

February 1978 *Post Office Datel Services* by Mr. A Cruikshank, Technical Officer, Glasgow Telephone Area.

April 1978 *Science, Electrons and the Postman* by Dr. Magnus Pyke, O.B.E.

We have been involved, along with other Centres in Scotland and members of the Senior Section, in restoring items for the proposed Scottish IPOEE museum at Morningside old telephone exchange, Edinburgh. To gain some experience in preparing and setting-out a museum, it was decided, by the museum committee, to attempt an exhibition. This was embarked upon and resulted in a 3-week *Telephone Exhibition* during the Edinburgh Festival which was attended by some 1300 members of the public. We are indebted to Mr. Tapper and his staff from the Telecommunication Headquarters exhibitions group for their assistance and the air of professionalism that their stands and items added to our exhibition.

For the first time we managed to enter a team in the National Quiz Competition where we were successful against Dundee in the first round but beaten by Edinburgh in the second round.

Our youngest Centre in Scotland (Stirling) has been in existence for some 8 years. During that time, John Hannah has been their secretary and was also one of the first delegates from Scotland to the Associate Section National Committee. In 1975, John was nominated for the chairmanship of the National Committee and continued in that office until 1977. It was with great pleasure that, on behalf of the National Committee, I was asked to present, to John, a certificate of honorary membership of that committee. Our congratulations go to John on receiving this recognition of his services, along with our thanks and good wishes for the future.

R. I. TOMLINSON

NOTTINGHAM CENTRE

For the third year running, our quiz team reached the Midland Region final; this year by defeating Derby in the first round and Technical Training College, Stone, in the semi-final. The latter competition was played via landline and resulted in a 38 to 37½ points win.

In January, members were entertained with a talk on *Home Winemaking* given by Domini Gregory. A lot of interesting information on the bureaucratic hurdles which have to be surmounted when producing wine commercially was passed on, together with samples of the wines.

One of our members gave an illustrated talk, in February, entitled *The Re-birth of the Great Central Railway*. At the end of the evening, those present were left with a greater insight into the problems to be faced when preserving a working steam railway.

The final visit of the year was in April, to the East Midlands Airport. After looking at the customs area and the fire appliances we had a talk on the airport and, finally, a walk through a Viscount aircraft.

M. RUSH

SALISBURY CENTRE

The Centre's recent programme has consisted principally of rounds in the National Quiz. We were victorious in the Regional competition by defeating Exeter, Gloucester and Bournemouth and received the Centenary Trophy from the Regional Director, Mr. Trevor Urben. The team represented South West Region in the first round of of the National Quiz where it was defeated by Ballymena.

The petrol shortage and inclement weather unfortunately prevented many members from joining the trip to the Daily Mail in London; however, more members were able to visit Oldbury-on-Severn nuclear power station.

D. J. TODD

Notes and Comments

CORRECTIONS

In *Notes and Comments*, published in the January 1978 issue of the *POEEJ*, the graphical symbol shown on p. 63 for the NOT function was in fact the symbol for the NOR function. To clarify the position the symbols for both functions are reproduced below.

In the *IPOEE* Institution notices on p. 64 of the April 1978 issue, under the sub-heading *Institution Ties*, the telephone number of Mr. Foster was incorrect. The correct number is 01-272 1422. The editors apologise for any inconvenience caused. In the list of retired members Mr. J. Chesbrough should read Mr. J. Chisbrough and Mr. M. G. Gray should read Mr. H. G. Gray.

NOT NOR

PUBLICATION OF CORRESPONDENCE

The Board of Editors would like to publish correspondence on engineering, technical or other aspects of articles published in the *Journal*, or on related topics. Letters of sufficient interest will be published under Notes and Comments.

Letters intended for publication should be sent to the Managing Editor, *The Post Office Electrical Engineers' Journal*, NP 10.1.4, Room S 08, River Plate House, Finsbury Circus, London EC2M 7LY.

Institution of Post Office Electrical Engineers

CHANGE OF SECRETARY

The Council of the Institution has approved the appointment of Mr. R. E. Farr as Secretary in succession to Mr. A. B. Wherry.

Mr. Wherry was appointed Secretary in 1967 and, during his period of office, rose from Senior Executive Engineer to his present position as Chairman of the North West Telecommunications Board. It is a measure of his interest in, and devotion to, Institution affairs that he has carried out the duties of Secretary for so long in the face of increasingly onerous official responsibilities.

Mr. Farr (not to be confused with Mr. B. Farr, Secretary-Treasurer of this *Journal*) is in Network Planning Department at Telecommunications Headquarters. His address is: NP 8.3.2, Room S 04, River Plate House, Finsbury Circus, London, EC2M 7LY; Tel.: 01-432 1954.

INSTITUTION FIELD MEDAL AWARDS, 1976-77

Field Medals may be awarded annually for the best papers on field subjects, primarily of Regional interest, which have been read at meetings of the Institution during the preceding session. Field Medals have been awarded for the 1976–77 session to Mr. G. F. Alton for his paper *Security Aspects of Telecommunications Buildings in Northern Ireland*, and to Mr. N. F. Hatton for his paper *Development of the Q-Type Standard Building*.

ASSOCIATE-SECTION PAPER AWARDS, 1976-77

Although prizes may be awarded for the best papers presented by members of the Associate Section of the Institution during the preceding session, there has been a very limited response in recent years. Only one paper, *Air Disaster* by Mr. N. V. Clark, Technical Officer, London Centre, was submitted for consideration for the 1976–77 session awards, and Mr. Clark has been awarded a prize of £10 and an Institution Certificate.

HONORARY MEMBERSHIP

Messrs. E. W. Fudge and C. T. Lamping have been elected to Honorary Membership of the IPOEE in recognition of their past services to the Institution.

R. E. FARR
Secretary

IPOEE CENTRAL LIBRARY

The following books have been added to the IPOEE Library since the publication of the 1974 Library Catalogue. Any member who does not have a copy of the catalogue can obtain one from the Librarian, IPOEE, 2–12 Gresham Street, London EC2V 7AG. Library requisition forms are also available from the Librarian, from Honorary local secretaries, and from Associate Section local-centre secretaries and representatives.

5240 *Beginner's Guide to Integrated Circuits.* I. R. Sinclair (1977)

A basic book which covers the principles, construction and uses of integrated circuits.

5241 *Beginner's Guide to Domestic Plumbing.* E. Hall (1977)

Aimed at the practical householder, this book covers its subject in simple and straightforward language with numerous illustrations.

5242 *Communicate: Parkinson's Formula for Business Survival.* C. N. Parkinson and N. Rowe (1977)

The innovators of *Parkinson's Law* investigate communication between the various groups involved in business, and find it sadly lacking. Their forthright views on the subject are outlined.

5243 *Electronic Meters: Techniques and Troubleshooting.* M. Ritter-Sanders (USA, 1977)

A practical guide to locating trouble in modern electronic equipment, using electronic meters as an aid.

5244 *A Practical Introduction to Electronic Circuits.* M. Jones (1977)

Written for the reader with a knowledge of simple electric circuits, the book progresses from basic concepts to the latest building blocks. Copiously illustrated, it contains sufficient information to enable those so inclined to construct and test their own circuits.

5245 *Power-transistor and TTL Integrated Circuit Applications.* Editor: B. Norris (USA, 1977)

This book covers a wide spectrum of semiconductor-device theory and application, divided into 4 sections: triacs, thyristors and unijunction transistors; power transistor applications; audio circuits; the use of TTL integrated circuits.

5246 *Television Video Transmission Measurements.* L. E. Weaver (1977)

Operational techniques for measuring the various distortions on video signals are given. Attention is focused on the most basic measurements, with particular reference to the handling of colour signals.

5247 *The White Noise Book.* M. J. Taunt (1974).

This book covers multi-channel communication systems and white-noise testing. It summarizes the history of frequency-division-multiplex and describes the various transmission techniques in both cable and radio systems. Special reference is made to the measuring units found in the telecommunications industry.

5248 *Sir Charles Wheatstone.* B. Bowers (1977)

Principally known as the inventor of the Wheatstone bridge, Sir Charles Wheatstone was a pioneer in electrical science and telegraphy. This biography reveals a man of great ingenuity and innovation who, because of his quiet approach, was largely overshadowed by his more volatile contemporaries.

2937/2 *Electrical Technology (Fifth Edition).* E. Hughes (1977).

The latest edition (in SI units) of this well-established and popular work. It covers the electrical engineering syllabus of the second and third-year courses for ONC and the first-year course leading to a degree in engineering.

5254 *Radio Circuits Explained.* G. J. King (1977)

This book covers the various circuits used in contemporary radio receivers, from the simple transistor set to the specialized high-fidelity receiver.

5255 *50 Simple LED Circuits*. R. N. Soar (1977)

Fifty practical circuits using inexpensive and freely available components are described.

5256 *Questions and Answers: Electric Motors*. A. J. Coker (1977)

Reviewed in *POEEJ*, Vol. 70, p. 184, Oct. 1977.

5257 *50 Field Effect Transistor (FET) Projects*. F. G. Rayer (1977).

FETs find application on a wide variety of circuits. The 50 projects cover radio-frequency amplifiers and converters, test equipment and receiver aids, tuners, receivers, mixers and tone controls, as well as miscellaneous devices which are useful in the home.

5258 *Managing Human Resources*. Inbucon Consultants (1976)

A book on one of today's most important themes: the development of the human potential in organizations. Written for practising managers, it covers a wide range of topics including the latest developments in management techniques.

R. CROSS
Librarian

CGLI & TEC: GUIDANCE FOR STUDENTS

Back numbers of the *POEEJ* are available, complete with *Supplements* containing model answers to past examinations of the City and Guilds of London Institute Telecommunication Technicians' Course, and tables of references to those answers that most closely match the syllabi of the Technician Education Council's Certificate Programme in Telecommunications.

The price of back numbers is 80p each, including the *Supplement* and postage and packaging. (*Supplements* are not sold separately.)

July 1970	TpB^{Q4-10}, TTA, CB, LPPB, TgB, MC, TPC^{Q1-5} (1969)	☐
Oct. 1970	TPC^{Q6-10}, TgC, TpC, LTC, LPPC, BMCC, CRC^{Q1-7} (1969)	☐
Apr. 1972	ES^{Q3-10}, TgB, TpB, MA, CA, CB^{Q1-8} (1971)	☐
Apr. 1973	PM^{Q3-10}, ETP, ES, TPA, TpB, TPB^{Q1-6} (1972)	☐
July 1973	TPB^{Q7-10}, TgB, LPPA, MA, CA, TPC^{Q1-7} (1972)	☐
Oct. 1973	TPC^{Q8-10}, CRC, TgC, MB, CB, BMCC, TpC^{Q1-3} (1972)	☐
Jan. 1974	TpC^{Q4-10}, MC, LPPB, RLTB, TTA, LTCT: (1972)	☐
Apr. 1974	LTC^{Q2-10}, LPPC (1972); ETP, PM, ES, RLTA, TPA (1973)	☐
July 1974	MA, TTA, LPPA, CA, MB, TpB, TgB^{Q1-6} (1973)	☐
Oct. 1974	TgB^{Q7-10}, LPPB, CB, MC, CRC, LTC, TpC^{Q1-9} (1973)	☐
Jan. 1975	TpC^{Q10}, TPC, TgC, LPPC, BMCC (1973); PM, ES, MA^{Q1-9} (1974)	☐
Apr. 1975	MA^{Q10}, RLTA, ETP, TPA, TTA, LPPA, MB, CA (1974); TPB, RLTB (1973)	☐
July 1975	TgB, TpB, LPPB, CB^{Q1-9} (1974)	☐
Oct. 1975	CB^{Q10}, RLTB, TPC, CRC, LTC, TpC, TgC, MC, LPPC^{Q1-7} (1974)	☐
Jan. 1976	LPPC^{Q8-10}, BMCC (1974); ETP, PM, ES (1975)	☐
Apr. 1976	TPB (1974); TPA, MA, RLTA, LPPA, CA, TPB, MB (1975)	☐
July 1976	TTA, RLTB, TpB, TgB (1975)	☐
Oct. 1976	LPPB, CB, TPC, MC, CRC, TgC, TpC, LTC (1975)	☐
Jan. 1977	LPPC, BMCC (1975); PM, RLTA (1976)	☐
Apr. 1977	ES, ETP, TPA, MA, LPPA, CA, MB, TPB (1976)	☐
July 1977	TpB, TgB, CB, LPPB (1976)	☐
Oct. 1977	TTA, TPC, MC, TpC, TgC, BMCC, LPPC, LTC (1976)	☐
Jan. 1978	[M1, PS1, TS1]; CRC (1976); PM, ES, RLTA (1977)	☐
Apr. 1978	[LCA1, M2, EP2, E2]; ETP, TPA, MA, LPPA, CA, MB, RLTB, CB (1977)	☐
July 1978	[TSS2, XS2, RS2]; RLTB (1976): TTA, TpB, MC (1977)	☐

CGLI

BMCC: Basic Microwave Communication C
CA, CB: Computers A, B
CRC: Communication Radio C
ES: Engineering Science
ETP: Elementary Telecommunication Practice
LPPA, LPPB, LPPC: Line Plant Practice A, B, C
LTC: Line Transmission C
MA, MB, MC: Mathematics A, B, C
PM: Practical Mathematics
RLTA, RLTB: Radio and Line Transmission A, B
TgB, TgC: Telegraphy B, C
TpB, TpC: Telephony B, C
TPA, TPB, TPC: Telecommunication Principles A, B, C
TTA: Telephony and Telegraphy A

TEC

[ ] Denotes tables of references
M1, M2: Mathematics 1, 2
PS1: Physical Science 1
LCA1: Line and Customer Apparatus 1
TS1: Telecommunication Systems 1
EP2: Electrical Principles 2
E2: Electronics 2
TSS2: Telephone Switching Systems 2
XS2: Transmission Systems 2
RS2: Radio Systems 2

Model-Answer Books A series of model-answer books is also available, in which selected answers from past examinations have been collected together to cover the following subjects.

Elementary Telecommunication Practice	60p	☐	Radio and Line Transmission A	85p	☐
Telecommunication Principles A	60p	☐	Telephony and Telegraphy A	85p	☐
Line Plant Practice A	60p	☐	Telecommunication Principles B	85p	☐

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The Post Office Electrical Engineers' Journal

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Communications

With the exceptions indicated above, all communications should be addressed to the Managing Editor, *The Post Office Electrical Engineers' Journal*, NP 10.1.4, Room S 08, River Plate House, Finsbury Circus, London EC2M 7LY (telephone 01-432 4840).

Model Answer Books

Books of model answers to certain of the City and Guilds of London Institute examinations in telecommunications are published by the Board of Editors. Details of the books available are given on p. 139 of this *Journal*. Copies of the syllabi and question papers are not sold by *The Post Office Electrical Engineers' Journal*, but may be purchased from the Department of Technology, City and Guilds of London Institute, 76 Portland Place, London W1N 4AA.

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