The Journal of the

BRITISH INSTITUTION OF RADIO ENGINEERS

FOUNDED 1925 INCORPORATED BY ROYAL CHARTER 1961

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

VOLUME 24 OCTOBER 1962 NUMBER 4

"THE RADIO AND ELECTRONIC ENGINEER"—

will be the title of the Institution's *Journal* as from January 1963.

The Council of the Institution has decided to make this change in order to emphasize the professional activity of members which, as described in the Charter of Incorporation, is to "promote the general advancement of and to facilitate the exchange of information and ideas on radio science and engineering including the theory, science, practice and engineering of electronics".

The first *Journal* of the Institution was published in October 1926, when it was referred to as *The Proceedings*. In 1939 it was retitled *The Journal of the British Institution of Radio Engineers* and the abbreviation *J.Brit.I.R.E.* has become a familiar reference in the world-wide abstracts which are made of papers which have appeared in the *Journal*. This abbreviation will continue to be used in referring to papers published in the *Journal* and this will be made clear by placing the longer title in full on the redesigned front cover of future issues.

The desire to emphasize electronics is not to be regarded in any way as being at the expense of the primary science of *radio* engineering for which the Institution was founded and which is underlined in the Royal Charter, whereby a Corporate Member "may . . . describe himself as a Chartered Radio Engineer". The order of mentioning these two important branches of engineering also makes this point.

Change of a name is always a serious matter and one not to be embarked upon without due deliberation, particularly where the familiar designation of a widely circulated publication is involved. It is believed, however, that members generally will welcome the new name for the reasons already stated. The idea implicit in *The Radio and Electronic Engineer* will, of course, lead to speculation as to whether the Council will take a final decision on the often-mooted proposal to petition the Privy Council for permission to change the name of the Institution itself. The reactions of members to this change in the name of the *Journal* will provide a useful indication of their approval of this other possible change.

As stated in the last Annual Report of the Institution, published in the September issue of the *Journal*, the circulation is now almost 10,000 per month, having nearly doubled since 1953. This is a noteworthy achievement, bearing in mind that the figure includes a considerable paid circulation to libraries, research establishments and many large electronic manufacturing organizations throughout the world. Quite apart from its value to the corporate membership of the Institution, the *Journal* has a large external readership which is growing year by year. The new title underlines the fact that it is now an important part of the literature of radio *and electronic* engineering.

G. D. C.

INSTITUTION NOTICES

The Bye-Laws of the Institution

Advice has now been received from Her Majesty's Privy Council that the Bye-Laws of the Royal Charter of Incorporation have been allowed. Reference was made in the August *Journal* (page 116) to the amendments which had been called for on the original draft of the Bye-Laws.

Dinner of Council and Committees

The 1962 Dinner of the Institution's Council and its Committees will be held at the Savoy Hotel, London, on Tuesday, 27th November, at 7 for 7.30 p.m. under the presidency of Admiral of the Fleet the Earl Mountbatten of Burma, K.G.

All members who are serving or have served on Institution Standing, Group or Local Section Committees may apply for tickets for themselves and their ladies. The cost of tickets is £2 10s. each.

List of Members 1962

The tenth edition of the List of Members of the Institution has now been published and copies have been sent to all Members, Associate Members, Companions, Associates and Graduates whose names appear therein. Registered Students may obtain copies, price 5s., from the Publications Department.

Graduateship Examination

The Graduateship Examination will be held at centres in Great Britain and overseas on 21st and 22nd November next. This will be the last examination in Section A under the current syllabus. The new syllabus will apply from May 1963 for this Section, while the new syllabus for Section B will take effect from November 1963. A new edition of the "Bye-Laws Governing Elections to Membership and Examination Syllabuses and Regulations" has now been published which contains the new syllabuses for both parts of the Examination and members may obtain copies on request.

Symposium on Masers and Lasers

The Programme and Paper Committee has arranged a one day symposium on Masers and Lasers on Wednesday, 2nd January at the London School of Hygiene and Tropical Medicine, Gower Street, London, W.C.1. Papers to be presented will include the following:

- "Optical Masers"—J. H. Sanders (Clarendon Laboratory, Oxford).
- "Characteristics of a Pulsed Ruby Laser"—R. J. R. Hayward and E. A. D. White (General Electric Company).
- "Maser Noise Measurements"—C. R. Ditchfield (Royal Radar Establishment).
- "Optical Pumping Equipment for the Ruby Laser"—R. E. W. Cook (National Physical Laboratory).
- "The Performance of an Ammonia Maser with Two Cavities in Cascade"—D. C. Laine and R. C. Srivastava (University of Keele).
- "Investigation of Relaxation Oscillations in the Output from a Ruby Laser"—R. C. Smith, D. Bhawalkar and W. A. Gambling (University of Southampton).

Registration will be necessary for this meeting and members wishing to attend are invited to apply to the Institution for the necessary forms which will be ready early in December. Further details will be given in the November *Journal*.

Radio and Electronics Research in Great Britain

The Research Committee are proposing to hold a symposium on the theme of "Radio and Electronics Research".

It is anticipated that there will be strong representation from the Universities and all members concerned with fundamental research are invited to offer contributions.

The provisional date for the Symposium is Monday, 3rd December; the proceedings will start at 2.30 p.m. in the London School of Hygiene and Tropical Medicine, Gower Street, London, W.C.1, under the chairmanship of Mr. L. H. Bedford, C.B.E., M.A., F.C.G.I. (Vice-President).

Electronics and Industrial Productivity

The title of the 1963 Convention—to be held in the University of Southampton from 16th to 20th April—has now been approved by the Council as "Electronics and Industrial Productivity". Sessions are planned under the following headings (some of the subjects which it is expected will be covered are given in brackets):

Measurement and Sensing Devices
(Transducers for quantitative measurements (e.g. pressure, temperature, position, optical factors, etc.; quality sensing; objective and subjective aspects.)

Information Transmission and Communication
(Digital and analogue data transmission and telemetering.)

Control and Information Processing
(Mathematical and logical techniques; storage; function generators; controllers; data loggers and computers.)

Output Devices and Final Control Elements
(Servomechanisms; actuators; displays.)

Industrial Applications of Electronic Systems
(Machine tool and positional control; inspection and non-destructive testing; process control systems; electronic methods in production; production control systems.)

Members are invited to submit papers for consideration for inclusion in the Convention. Offers should be accompanied initially by synopses of 200 words and these should be sent to the Institution as soon as possible.

U.D.C. 681.39

Some General Features of Digital Data Acquisition Systems

By

E. B. STUTTARD, B.A.†

Presented at the Symposium on "Recent Developments in Industrial Electronics" in London on 2nd–4th April 1962.

Summary: Digital data acquisition systems make use of a single measuring system to obtain data from many inputs and record the results for future use. Such systems usually accept analogue inputs and incorporate equipment for conversion to digital form. A typical system is described followed by details of the more novel techniques used in the major sub-units. The ways in which such systems are employed are also discussed.

1. Introduction

In almost all industries the problem of measurement is becoming more acute in terms of the quantity of information to be gathered, its analysis, the accuracy required and the resulting price which has to be paid. Research requires that enormous quantities of data be gathered, and continuous monitoring of a plant for alarm conditions is a common industrial requirement.

In many cases it is convenient and economic to meet the requirements by means of a single measuring device which is shared between many inputs. Because a single measuring device is used it is possible to provide a more sophisticated instrument than would otherwise be possible, usually including digitization and subsequent print-out.

2. System Description

2.1. *Scanning*

Figure 1 is a block diagram of a typical data logging system. It is necessary to provide a device which will present the inputs one at a time to the measuring equipment. This is the scanning unit and it consists essentially of a two-pole switch with as many positions as there are inputs. The basic switching element may take a number of forms including uniselectors where low price is the main consideration, relays for general-purpose use and, in certain applications, solid-state devices.

The scanning unit is also used to provide channel identification information to the printers, to select the levels at which an alarm will occur for each input and define the correct scale factor.

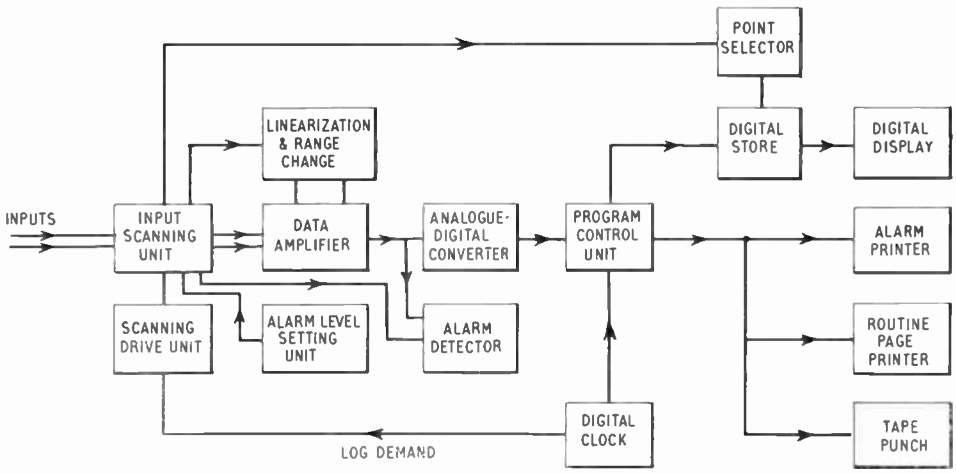

Fig. 1. Typical data logging system.

The scanning process is controlled by a scanning drive circuit which is capable of operating at various speeds. Under normal circumstances, when no record is required, the system will be scanning continuously, looking for alarms at, say, 15 inputs per second. When it is required to print results this speed must be reduced to a value dependent on the type of printer used. This is usually achieved by an inhibit circuit controlled by the printer itself which thus determines the speed.

† Blackburn Electronics Ltd., Brough, Yorkshire.

2.2. *Amplification*

Scanning is performed at low level and it is then necessary to amplify the input signals in order to bring them to an amplitude sufficient to operate the alarm detector and analogue-to-digital converter. The data amplifier is a d.c. amplifier of low noise and drift and a sufficiently good frequency response to ensure that its rise time is adequate for the scanning speed chosen. It is customary to present the output in a correctly scaled form, e.g. degrees, lb/in^2, etc., although millivolts may be more convenient for scientific work. If more than one sort of transducer is being used it may be necessary to change the gain of the amplifier and this is performed by a range change unit which varies the amplifier feedback under the control of the scanning unit. In addition linearization may be required as with thermocouples, or square root extraction for flow rate measurement. These facilities are provided by diode function generators also located in the amplifier feedback and brought into operation by the scanner on the appropriate inputs.

2.3. *Alarm Detection*

Many industrial processes require that certain parameters be kept between prescribed limits and an alarm monitoring system is then included. Alarms are detected in analogue form at the output of the data amplifier since this places the alarm detector as close as possible to the transducer.

Alarm levels are defined by voltages derived from a reference power supply by means of a resistive potentiometer chain. The upper and lower alarm reference voltages are then compared with the amplified input signals in the alarm detector which is a form of voltage comparator.

The levels themselves may be set in a number of ways such as multi-turn potentiometers or fixed resistors selected by decade switches. If complete flexibility is required a pinboard may be provided and used to select the units, tens and hundreds of the alarm level for each point separately. Grouped alarms may also be provided in which case some form of patchboard is included. The alarm detection circuits also permit certain faulty transducers to be detected in addition to high and low parameter values.

The output of the alarm detector may be used to initiate an alarm bell, to operate individual annunciator lamps and to control the alarm printer.

2.4. *Digital Equipment*

The analogue-to-digital converter produces a decimal numerical output from a voltage input, this typically takes the form of a voltage on three of thirty lines for a three-decade system. Thus a voltage of 9·99 might produce an output of 999 and *pro rata.*

The program control unit, as its name suggests, controls the operation of the whole machine. One of its functions is to ensure that the output of the analogue-to-digital converter is in the correct form for operation of the output printers. Thus a strip or line printer requires a decimal input in parallel form, all digits appearing simultaneously. In general, the output of the analogue-to-digital converter, together with identification and other information, is already in this form so it is necessary only to arrange that the levels are correct. Typewriters require a serial decimal input in which the digits are applied one at a time at a speed of about 10 per second; serialization must, therefore, be provided. If punched paper tape is to be produced there is the additional requirement of a device to produce the appropriate binary decimal code for operation of a digital computer or teleprinter.

The program control unit also performs a check on the operation of the system once per scan, a suitably attenuated voltage from a reference cell being injected at the input of the data amplifier. The resulting signal is examined in analogue form at the output of the data amplifier by means of the alarm detector and digitally at the output of the analogue-to-digital converter by means of a simple AND gate system. Any failure is used to raise an alarm, the nature of which indicates the part of the equipment which is faulty.

A digital clock is usually included in the data logging system and, in addition to providing a source of time for printing, is used to demand routine logs of all points at pre-set intervals, say once per shift.

2.5. *Printers*

It is common to provide two printers, a line or strip printer and an electric typewriter. The strip printer has the advantage of high speed with the disadvantage of an inconvenient format. It is, therefore, used to record alarm information only, the print-out including point identification number, parameter value and a symbol to denote the nature of the fault. As this printer operates only when an alarm occurs, it has little effect on the time per complete scan and its record contains no unnecessary data. It is customary to print the time at the end of any scan in which an alarm has occurred.

The typewriter is used for routine logging of all points and by using the tabulation facility it is possible to present the results in any desired format. Additionally, by using the colour change facility, points in a state of alarm are printed in red. The routine log usually starts with the time and value of the automatic system check followed by the record of parameter values. In addition an output may be provided on punched paper tape or magnetic tape.

It is sometimes necessary to provide a visual display which can be used to present the value of any selected input. As the digitized value is available only while a point is being scanned it is necessary to include a digital store in the system. This retains the value for the duration of one scan when it is cancelled and the new value inserted.

From the foregoing brief description of the overall data logging system it will be apparent that most of the requirements of modern instrumentation can be met by such an equipment. Any required measurement available in electrical form can be handled. Some of the main units of the system will be described in rather more detail.

3. Unit Description

3.1. *Input Scanning Unit* (Fig. 2)

A scanner is essentially a multipole switch, the number of "ways" being determined by the number of inputs to be logged. Two pole switching is usually employed although four poles may be needed for certain bridge-type transducers.

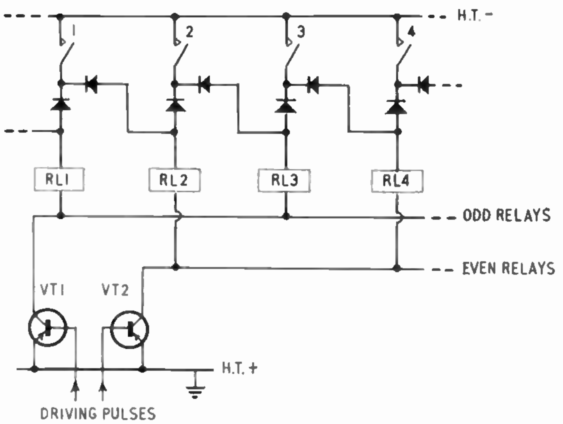

Fig. 2. Block schematic of reed relay scanning circuit.

To achieve a reliable and readily manufactured unit, use is usually made of relays, one relay being employed for each input. For the sake of convenience in manufacture and use the relays are mounted and wired in banks of ten, it being possible to omit any bank from the scanning sequence if desired.

The scan drive circuit consists of a pair of transistor switches operated by anti-phase voltages from a bi-stable. Thus one side of the switch is always conducting and the other cut off. All the even number relay-coils are connected to one side of the switch and the odd coils to the other. Each relay also has one of its own contacts in its coil circuit to hold itself while diodes prepare the following relay in the scan sequence for energization. Each time the bistable reverses its state the scanner moves forward one step.

Using this form of scanner the same scan drive circuit is used whatever size of scanner is constructed. Scanners of this type have been manufactured to sample from as few as 10 to as many as 1000 inputs.

Many types of relay have been employed in data logging systems but most current machines use dry-reed or mercury-wetted types.

The services the scanner provides to the logger are input switching, identity information and alarm level and scale factor selection.

3.2. *Data Amplifier* (Fig. 3)

After each input has been switched into the data logger it must be amplified, correctly scaled and brought to the same earth reference point as the logger. These functions are all performed by the data amplifier.

The inputs to be sampled may be considerably separated physically and electrically. The input circuit must, therefore, be completely floating and this is achieved by using a capacitor transfer technique. A capacitor is placed across the input lines, charged to the voltage between these lines and then transferred to the input of the data amplifier. In this way the input lines are never connected to the data logger in any way.

Use of this technique demands an amplifier with a very low input current. A large capacitor will allow more amplifier input current for a given error, but will involve a lower sampling rate. The size of the capacitor chosen is usually 1μF. Input currents of the required order are achieved readily by using industrial electometer type valves.

The gain of the data amplifier must be accurately known and stable. This is achieved by the use of an amplifier with very high open loop gain and overall negative feedback. To maintain the necessary high input impedance while including the input valve in the feedback loop, use is made of series feedback instead of the more usual parallel arrangement.

In order to stabilize the working point of the amplifier, drift correction must be employed. During the quiescent period between taking readings the

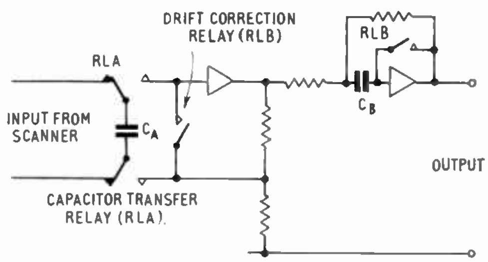

Fig. 3. Data amplifier.

The Symposium on Sonar Systems

Nearly 150 delegates from ten countries took part in the Symposium on Sonar Systems which was held at the University of Birmingham on 9th–12th July last. The Symposium which was the first open meeting of such magnitude on the subject to be held in Europe was sponsored jointly by the Electro-Acoustics and Radar Groups of the Institution, the Acoustics Group of the Institute of Physics and the Physical Society, and the Electrical Engineering Department of the University of Birmingham.

Twenty-eight of the 29 papers presented during the Symposium were preprinted and circulated to delegates a few days before the meeting. This undoubtedly stimulated the very informed discussions.

A notable feature of the papers was their diverse origins—less than 60% of the papers were from Great Britain and of the remainder, eight were from the United States, and one each from Canada, Italy, Norway and Western Germany. The authors' organizations were divided roughly equally between university or civil research establishments, industry, and naval research establishments.

The subject matter of the paper extended over virtually the whole field of sonar, excepting military applications, from theoretical studies of wave propagation to descriptions of the most recent techniques for fishing and surveying.

One of the sessions dealt with "Arrays and Signal Processing" and the papers given under this heading have implications for application to fields other than sonar, notably in radar.

A particularly useful discussion took place at the end of the Symposium on "The Future of Non-Military Sonar" and in view of its general interest a report will be given in an early issue of the *Journal*, together with abstracts of all papers. Publication of the papers themselves, with discussion where applicable, will also start with publication of two papers in the December *Journal*.

Symposium Dinner

The majority of the delegates stayed at the very pleasant hall of residence, The Manor House, Northfield, about two miles from the University. This was the scene of the Symposium Dinner on the Monday evening which was presided over by the Vice-Chancellor of the University, Sir Robert Aitken. Welcoming delegates, Sir Robert stressed the value of conferences to scientists and engineers and expressed the view that the size of the symposium was probably the optimum for encouraging informal contact between delegates. He welcomed the open discussion of a subject hitherto under a blanket of secrecy and particularly referred to the increasing number of non-military applications such as in oceanography. The Vice-Chancellor expressed especial pleasure at the initiative of Professor D. G. Tucker (*Member*) in conceiving and planning the symposium.

Sir Robert then introduced the guest of honour, Professor Sir Solly Zuckerman, Scientific Adviser to the Minister of Defence.

The importance of sonar was stressed by Sir Solly who pointed out that if sonar workers could demonstrate the solution of certain difficult problems, ample funds could be provided for development of new devices in fishing, oceanography, etc. Many of the papers in the symposium dealt with the physical laws and he recalled that the early work on sonar—then called ASDIC—in the first world war was carried out by physicists such as Langevin and Perrier in developing transducers. Sir Solly referred particularly to the use of sonar in seismic research and said that this was of important topical interest in view of the Geneva Conferences on detection of nuclear explosions.

Conclusions on the Symposium

In summing up the lessons of the Symposium, it can be stated that its particular feature was the bringing together of workers from very diverse fields of endeavour. Sonar is essentially an application of physics made practical by the use of the techniques of the electronic engineer, but its potential uses are of interest to scientists and engineers who are concerned with seismology, meteorology, oceanography, navigation, non-destructive testing and fishery research as well as those who deal with military applications. Few organizations are in the position to sponsor meetings of this kind and hence the Institution can provide a very valuable service to the wide field of science by the occasional arrangement of such symposiums.

A list of papers presented at the Symposium was published in the June 1962 issue of the *Journal* (to which should be added "The Portrayal of Body Shape by a Sonar or Radar System" by A. Freedman (A.U.W.E.). Limited stocks of reprints of some of the papers are still available and may be ordered from the Institution, price 2s. 6d. per paper.

U.D.C. 621.317.725

Automatic Scanner Loggers for Small Installations

By

J. JARDINE†

Presented at the Symposium on "Recent Developments in Industrial Electronics" in London on 2nd–4th April 1962.

Summary: Automatic scanner loggers convert analogue signals representing process variables to more concisely recorded digital data, and in many instances have replaced conventional recorders and indicators previously in use in centralized control rooms. To justify the use of a scanner logger on a small installation is fairly difficult since conventional multipoint recorders will normally meet the requirements at a greatly reduced cost. There is a need for a range of digital instruments to replace the conventional potentiometric recorders eliminating the disadvantages of these instruments while providing relevant information. This paper discusses such a range of instruments referring particularly to their advantages over conventional potentiometric recorders.

1. Introduction

In the process control field, variables such as pressure, flow and temperature are measured by suitable transducers, the outputs of which are in analogue form. Conventional multi-point indicators and potentiometric recorders currently and widely used to indicate or record these variables present fundamental design problems of instability in the servo loop and non-reliability of the dynamic sections of the instrument when faster scanning speeds are required, while limitations are imposed when individual alarm operation for multi-point installations are considered.

The basic instrument discussed in this paper aims to resolve these conventional problems by adopting "static" calibrated measuring systems of the successive comparison type and a flexible system of construction by which facilities for recording, alarm operation, modes of operation, etc., can be individually tailored to appeal to the widest circle of industrial users. A 12-point instrument with a digital range of 0–1599 units, shown in Fig. 1, is used as a model, since it illustrates in simple form the principles involved. However, it is emphasized that other instruments operating on the same principles can provide a greater or lesser number of facilities.

2. Design Aims

The following design aims were considered sufficient to retain all the facilities available on conventional instruments while improving on accuracy, simplicity and flexibility.

(a) The instrument should employ a conventional null balance technique (other than a servo balance system with slide-wire or moving coil) with an off-balance input impedance of 500 ohms.

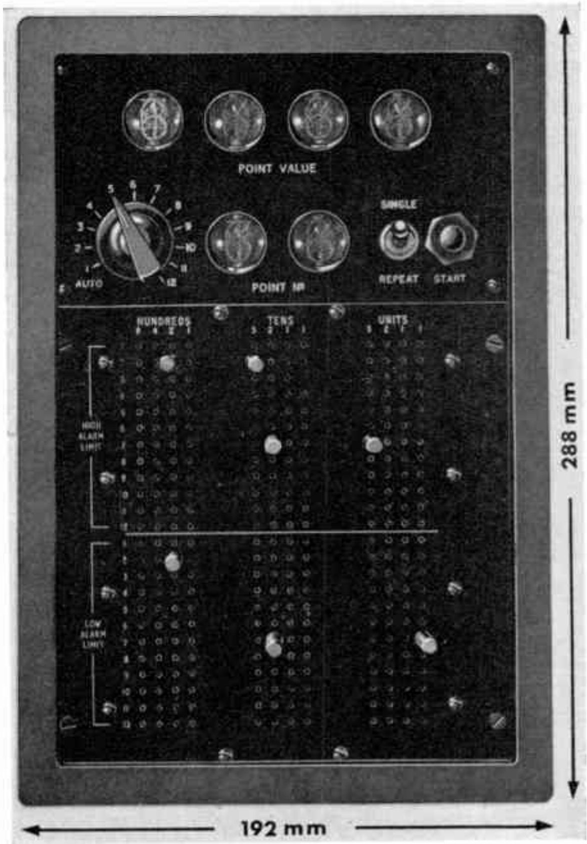

Fig. 1. 12 point digital voltmeter.

† S. Smith & Sons Ltd., Industrial Division, Glasgow, S.W.2.

(b) The accuracy should be 0·1% of reading.

(c) Presentation of the information should be in digital form.

(d) No pre-amplification of input signals should be necessary.

(e) The instrument should be small and suitable for panel mounting.

(f) Alarm setting facilities should be simple, flexible and available to the operator for reference purposes.

(g) Linearization of non-linear inputs should be built into the instrument.

(h) Facilities for individual selection of inputs should be available.

(i) The instrument should have auxiliary outputs available for chart recording, remote display of alarm conditions, remote display of input value and digital or coded print-out.

(j) The instrument should be capable of field servicing.

3. Basic Instrument

It was apparent from the design aims that the techniques already employed in digital voltmeters and data handling systems would lend themselves most easily to modification. The choice of a voltage encoder was governed by the requirement that the instrument should be adaptable for both analogue-to-digital conversion for recording, and alarm checking for monitoring purposes. This eliminated voltage encoders involving the translation of angular shaft positions to digital outputs and voltage encoders involving an intermediate conversion into a time interval, since neither of these encoders can easily be adapted to alarm checking. An encoder with no intermediate conversion, operating on a programmed control voltage principle and employing a binary coded decimal system was considered the most suitable.

3.1. *Input Selector and Analogue-to-Digital Converter Drive Units*

The input selector and analogue-to-digital converter drive units are shown in Fig. 2. These units carry out a repetitive sequence governed only by the mode of operation selected by the operator. The units contain transistors, diodes and relays, while the modes of operation are selected by two switches, SINGLE/REPEAT and AUTO/MANUAL, located on the front panel of the instrument. The AUTO/MANUAL switch is a thirteen-position switch, twelve positions being used for manually selecting any input to the instrument while the last position is the automatic position.

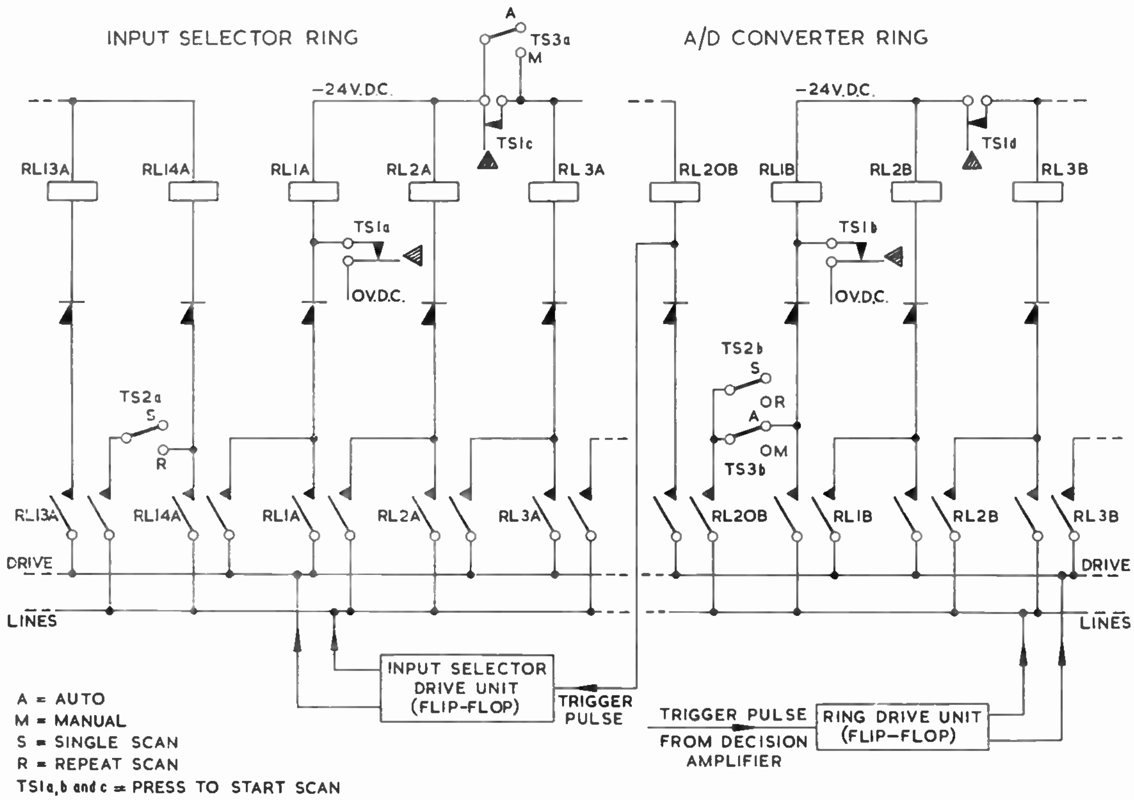

Fig. 2. Input selector and analogue-to-digital converter drive units.

The instrument has the following modes of operation:

(a) Repetitive cycles of automatic sampling of all instrument inputs at a rate of one point per second.

(b) Single cycle of automatic sampling of all instrument inputs at a rate of one point per second.

(c) Manual selection of any one input with the facility of single measurement.

(d) Manual selection of any one input with the facility of repetitive measurement at one second intervals.

An electro-mechanical relay analogue-to-digital converter drive unit was adopted in preference to a transistorized version for economy of cost and space. In addition, the multiplicity of contact arrangements possible with relays enables a variety of functions to be obtained irrespective of power levels. Servicing of the instrument is simplified since operation can be observed without recourse to special testing units and, in the event of a failure, the plug-in facility allows of rapid replacement. The relays used have a life expectancy of 10^8 operations and while this in no way compares with the transistor equivalent it will allow an expected trouble-free life of 4 years continuous operation at a measuring speed of 1 point per second.

However, advantage has been taken of the greater reliability and faster operating time of transistor switches and these have been employed elsewhere in the instrument, e.g. in the switching of the drive and decision lines which have to be operated 20 times per second.

The analogue-to-digital converter drive unit functions are outlined in Table 1. A transistorized bistable flip-flop which derives its trigger pulses from the decision amplifier is used to switch each relay RL1B to RL20B in sequence, the dwell time on each position being 40 milliseconds. A conventional method of alternately energizing two common drive lines and selecting the appropriate relay to be energized by the cycling contacts is employed.

The input selector drive unit switches each analogue input in turn to the weighing network for measurement purposes, sets up alarm levels appropriate for each input, and provides information to digital tubes for point identification. This unit is identical in operation to the analogue-to-digital converter drive unit with the exception that its flip-flop is triggered from the last position on the analogue-to-digital converter drive unit ring, i.e. RL20B. The analogue-to-digital converter drive unit flip-flop derives its ground return from the input selector flip-flop and is prevented from running while the latter is de-energized.

Table 1

Analogue-to-Digital Converter Drive Unit Functions

Position	Function	Auxiliary Function
RL1B	Reset	
RL2B	High Alarm	Pull in high alarm setting
RL3B	Low Alarm	Pull in low alarm setting
RL4B	Switch 800	
RL5B	Switch 400	
RL6B	Switch 200	Print colour and store alarm
RL7B	Switch 100	
RL8B	Switch 50	
RL9B	Switch 20	Print thousands
RL10B	Switch 10	
RL11B	Switch 10	
RL12B	Switch 5	Print hundreds
RL13B	Switch 2	
RL14B	Switch 1	
RL15B	Switch 1	Print tens
RL16B	Spare	
RL17B	Spare	
RL18B	Spare	Print units
RL19B	Spare	
RL20B	Rest	Pulse input selector flip-flop.

Initiation of an automatic sample is by means of a START push-button located on the front of the instrument. Operation of this push-button switches on transistor switches TS1a and TS1b. At the same time, transistor switches TS1c and TS1d open circuit the common −24 V d.c. line in the drive units to prevent indiscriminate locking up of the relays in the rings.

TS1a operates relay RL1A, switching the first input to the weighing network for measurement, while TS1b operates relay RL1B which carries out the first function of the analogue-to-digital converter drive unit, namely resetting the weighing network relays in preparation for a new measurement cycle. Until the START push-button is released, the analogue-to-digital converter drive ring flip-flop cannot run since it is locked up to the 0 volt start line via TS1b. On releasing the START push-button, the analogue-to-digital converter drive flip-flop will run in synchronism with the decision amplifier at 25 cycles per second carrying out the functions outlined in Table 1. When the last position RL20B is reached, a trigger pulse is fed to the input selector flip-flop; relay RL2A will be energized, switching the second input to the weighing network for measurement.

The position of the SINGLE/REPEAT switch decides whether transistor switch TS2a is switched on or off. Should a single scan be required TS2a will be off and the input selector switch flip-flop will stop at the end of a single scan, automatically switching off the analogue-to-digital converter flip-flop; otherwise, the instrument will continue scanning. It can be seen that on an automatic mode of operation, transistor switch TS2b is inoperative, being permanently shunted by transistor switch TS3b.

On manual operation, transistor switch TS3a permanently shunts TS1c, preventing drop out of the selected relay when operating the START push-button, the ground return of the input selector flip-flop is permanently disconnected while the analogue-to-digital converter flip-flop is permanently connected. Operation of the START push-button will therefore initiate a single read or repeat read of the selected input, depending on the condition of TS2b.

3.2. *Weighing Network and Decision Amplifier*

Figure 3 illustrates the weighing networks. Relays RL3C–RL14C control the graded resistors R1–R12,

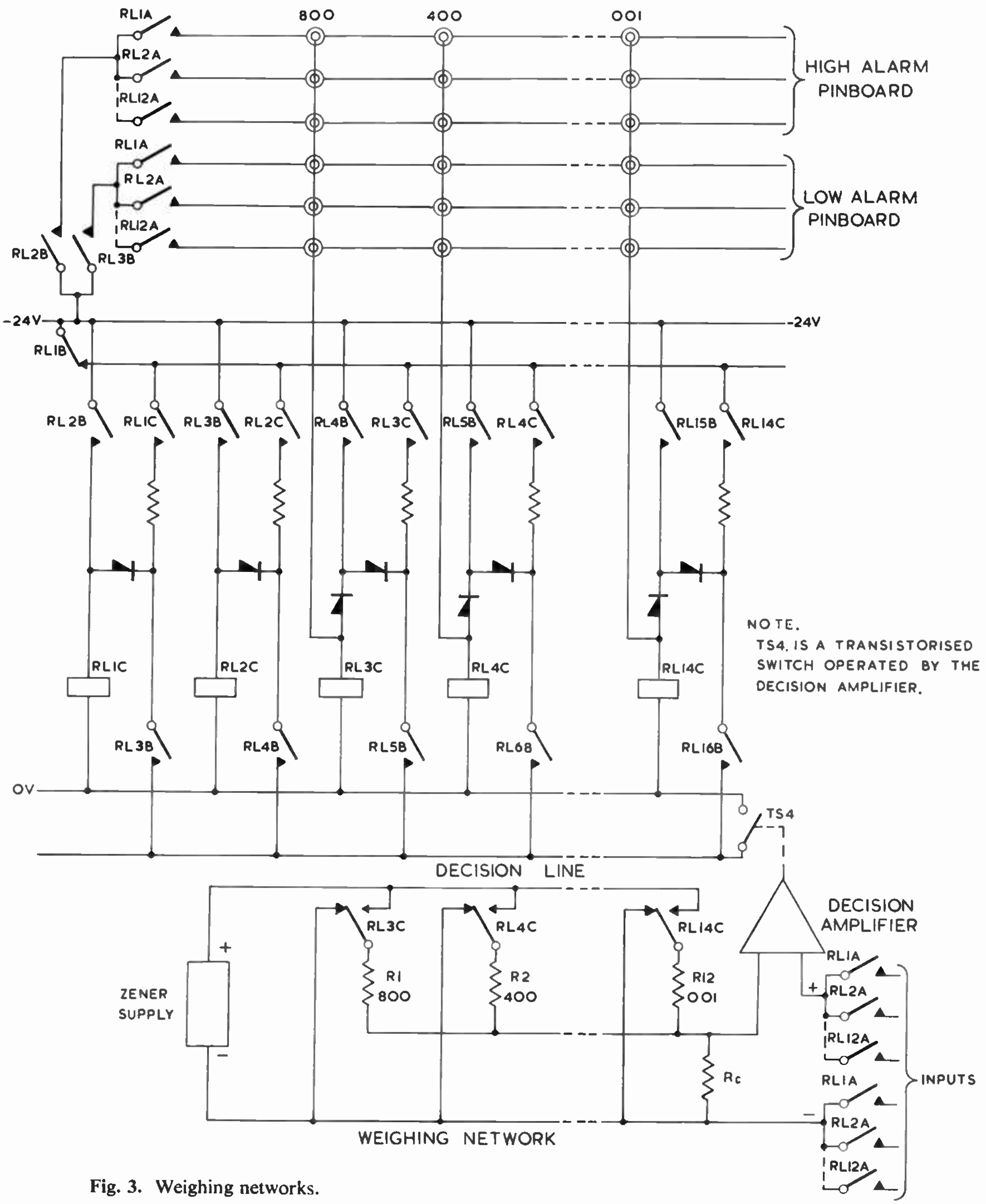

Fig. 3. Weighing networks.

proportioned in a binary coded decimal form as outlined in Table 1. These resistors are capable of switching currents through the reference resistor R_c to produce a minimum full scale voltage of 15·99 millivolts to an accuracy of 0·1% of value or 10 microvolts, whichever is the greater.

It can be shown that the reference voltage V_0 is given by the expression

$$V_0 = V_z \frac{\sum \frac{1}{R_s}}{\frac{1}{R_t} + \frac{1}{R_c}}$$

for any setting of the switching relays.

where $\frac{1}{R_t} = \sum \frac{1}{R_s} + \sum \frac{1}{R_p}$

$\sum \frac{1}{R_s}$ = total conductance in series with the supply V_z

$\sum \frac{1}{R_p}$ = total conductance shunting R_c

and V_z = supply voltage.

The reference voltage is therefore independent of the number to be decoded, being a function only of the accuracy of the resistors and the constancy of the supply.

The supply is a single Zener stabilized supply with no series regulator or feedback amplifier and has a stability better than 0·05%.

Ideally, the weighing network should present a constant load to the stabilized supply. However, the network shown in Fig. 3 will have a current drain from zero to a full load of 2 mA. By including a second set of proportional resistors (also controlled by the switching relays) to shunt the supply, it is possible to reduce the load variation considerably. Linearization of the weighing network by switched resistors further reduces the load swing to ± 1 μA, enabling scaling and linearization as discussed in Section 3.4 to be carried out by series resistor networks.

The programme of measurement is as follows. RL4B operates relay RL3C which switches a current through R_c proportional to half the analogue range or 800 digits. This voltage is "weighed" against the input signal and a difference signal passed to the decision amplifier. The latter is a high-gain pulse amplifier whose output is dependent on the sign of the difference signal.

Forty milliseconds later RL5B operates, shunting RL3C to the decision line. The decision amplifier also makes its decision at this time; however, the decision line is not switched for some 10–15 milliseconds to ensure that RL4B has opened. If the input voltage is greater than the reference voltage, transistor switch TS4 will be open and RL3C will remain operated, and the 800 proportioned voltage will be retained. Should the input voltage be less than the reference voltage, TS4 will be closed and RL5B will shunt relay RL3C and the 800 proportioned voltage will be rejected. The programme of switching questioning and answering is repeated for all the weighing resistors proceeding from the most significant to the least significant digit. Twelve questions will be asked, each one narrowing down the field, the final answer being within one digit.

The pattern of the RLC relay contacts existing at the end of a measurement cycle provides information on the input value in binary coded decimal form. This is transformed to a decimal form for visual presentation and print-out and can be further transformed by means of a diode matrix to a pure binary form for entry to a punch tape machine.

3.3. *High and Low Limit Alarm Setting*

It can be seen from Fig. 3 that relays RL3C–RL14C can be operated in parallel from the pinboard. By selecting the correct high and low limit alarm levels in binary coded decimal form and inserting pins in the appropriate positions of the pinboard, operation of RL2B and RL3B will set up the correct high and low limit reference voltage across R_c via the input selector relays, contacts RL1A to RL12A. The alarm decision will be taken in the same way as already discussed in Section 3.2, relays RL1C and RL2C retaining the alarm condition when it occurs.

It is interesting to note that should alarm scanning be required without measurement the analogue-to-digital converter ring drive unit will only require four relays, in which case the period of sampling each input will be 160 milliseconds compared to 800 milliseconds for alarm checking and measurement.

3.4. *Scaling, Linearization and Cold Junction Compensation*

Since the weighing network presents essentially a constant load, by inserting dividing networks between the Zener supply and the weighing network, it is possible to modify the analogue range of the measuring instrument. In the case of linear inputs, the analogue range is selected to match the analogue range of the transducer and provides an output in physical quantities. For non-linear inputs, the curve relating the analogue output of the transducer and the physical measurement is first approximated to a series of slopes, four or six approximations being taken according to the degree of linearization required. The analogue range of the measuring instrument is then modified automatically in four or six stages to correspond to these slopes. In addition to scaling, each slope will have a different intercept on the Y-axis and it is necessary also automatically to switch associated

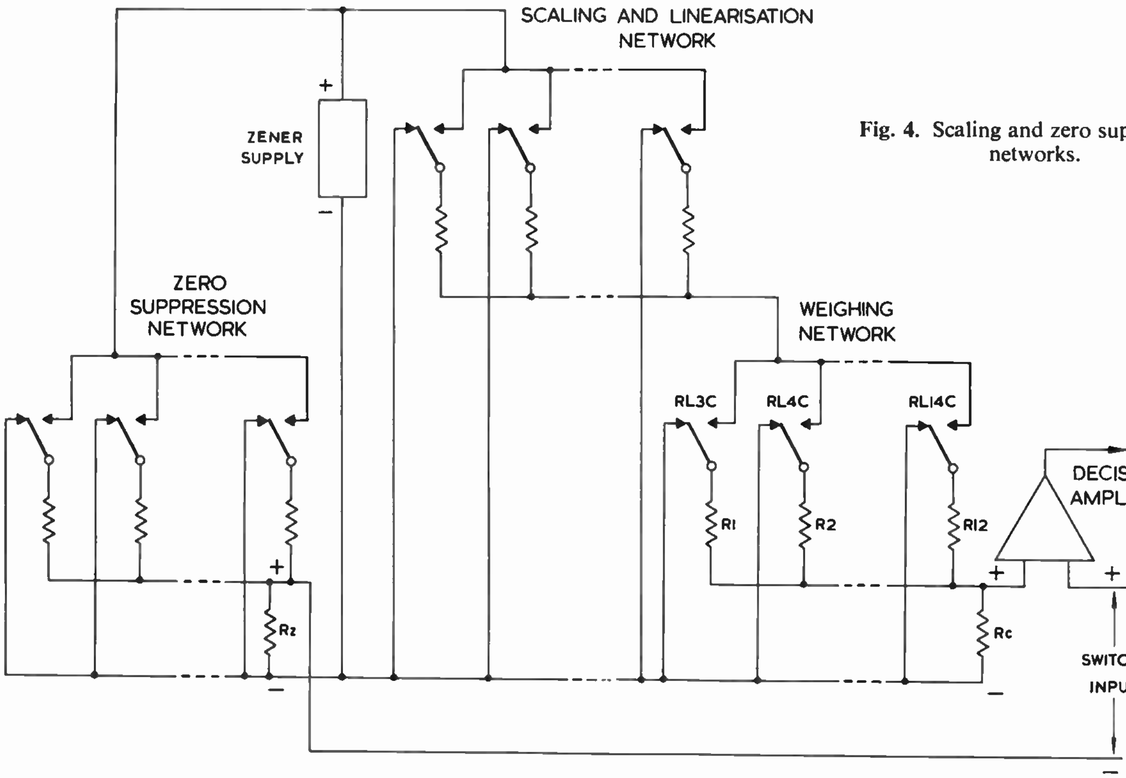

Fig. 4. Scaling and zero suppression networks.

zero suppressions. The appropriate scaling and suppression networks are switched in by the analogue-to-digital converter drive unit ring RLB relays. Figure 4 illustrates the scaling and zero suppression networks.

Cold junction compensation is essentially a zero suppression whose value is dependent on ambient temperature, and can be inserted in the instrument by replacing part of R_z by copper to the correct value.

4. Auxiliary Outputs

4.1. *Alarm Indication*

Remote alarm annunciation in the form of individual alarm lights indicating high, low, and normal conditions can be provided for each point with facilities for storage or automatic reset when the input condition returns to normal.

4.2. *Logging*

Automatic logging facilities can be provided extraneous to the instrument. The solenoids of the print-out device, which may be an electric typewriter, in-line printer or punch tape machine, are energized via a relay tree or diode matrix. No additional print-out sequencing unit is required, the print-out commands being auxiliary functions of the analogue-to-digital converter drive unit.

4.3. *Remote Indication*

Any number of remote indicating stations can be provided from the basic instrument, the indicator unit containing its own relay tree, digital indicator tubes and power supply.

4.4. *Chart Recording*

Chart recording of the instrument inputs can be provided extraneous to the instrument. The recording mechanism can be a simple chopper bar recorder synchronized from the analogue-to-digital converter drive unit and supplying trigger pulses to the input selector drive unit. The input to the recorder movement is a binary coded decimal contact arrangement with proportioned resistors, the movement acting as the current collector. Since high currents can be used, fast response can be achieved.

5. Conclusions

Some of the facilities available in the basic instrument have been discussed. However, there are many additional features which can be provided due to its inherent flexibility. An important facility of the instrument in its application to process monitoring is its ability to produce permanent records in chart or print-out form only when a process upset occurs. Thus, only relevant data of process behaviour will be accumulated.

Table 1 indicates that there is no severe limitation on the functions available in the instrument since the functions outlined in Table 1 can be permuted or modified to obtain such additional functions as polarity checking or alternative digital ranges.

No mention has been made in this paper of the application of the instrument to process control. However, it is evident that simple on/off two-position or three-position control can easily be incorporated. In the case of more sophisticated controls such as proportional, integral and derivative, consideration must be given to the relationship between the sampling rate and the process time-constants to ensure that instability in the closed-loop control system will not result. If the sampling rate is sufficiently fast, the instrument can supply an error signal to the conventional controller.

6. Acknowledgments

The author wishes to acknowledge the design effort of Mr. R. Swarbrick and Mr. A. Gibson and to thank the Directors of S. Smith & Sons (England) Ltd., for permission to publish this paper.

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DISCUSSION ON DATA LOGGERS

In the chair: Mr. P. Huggins (*Member*)

Mr. G. H. King: I am well aware that most data loggers for plant monitoring fitted with alarm facilities are required by the customer to make these independent of the digital output. I am sure that it was for that reason that both Messrs. Stuttard and Jardine have carried out their alarm monitoring on the analogue signal. However, I would make a plea for doing it at the digital stage, would it not be cheaper and just as effective?

Mr. E. B. Stuttard (*in reply*)*:* Digital alarms may be provided and are, in fact, cheaper since the requirement for an independent reference power supply unit is eliminated. Apart from the advantage of having the alarm detection independent of the digital part of the equipment, however, it is possible by using the analogue alarms coupled with an automatic system check performed in both analogue and digital form to locate any faults speedily to the appropriate part of the machine.

Mr. J. Jardine (*in reply*)*:* In data loggers of the successive approximation type no actual measurement is made on the analogue input signal to present it in digital form when alarm monitoring is required. Alarm monitoring is simply a voltage comparison, the alarm reference level having been preset in digital form. Should a digital comparison be required it would be necessary to carry out an analogue to digital conversion before making an alarm comparison. In many instances the input value is of interest unless an alarm condition exists and encoding time will have been wasted. In the digital instruments described, no economy can be achieved since the only additional component for alarm monitoring is the alarm setting pinboard.

Messrs. R. D. Brooker and R. B. Stephens (*in reply*)*:* Although this is not covered in our paper, we would like to express our agreement with Mr. King's comments.

In most of these data logging systems, periodic checking from input to output is specified, and under these circumstances it appears to us that the digital method is just as reliable and certainly as accurate as the analogue alarm system.

Mr. J. J. Hunter: If we examine different kinds of digital equipment, we find that there are only two types of digital output devices. These are:

(1) displacement encoders, either in rotary or linear form;

(2) counters, either directly counting a frequency variable output or preceded by either an analogue voltage to digital converter or a ratio to digital converter.

I would suggest that a general data collecting system should be able to deal with, or be compatible with, both these possibilities.

This is not difficult to arrange and it is worth the extra cost involved since there are more digital transducers available than seems to be generally realized, and it is certain that many more will become available. Thus the possibility of altering the type of transducers should be allowed for.

The compatibility requirements would be eased by use of a standard code at the input to the data recording devices, i.e. the tape punch, typewriter etc. For this, I would suggest binary coded decimal.

With regard to the linearization facilities, I wonder if these are often necessary since the impetus to the use of digital data collecting is provided by the availability of a computer and linearization is easily dealt with in the computer.

A small low cost digital system as described by Mr. Jardine should be very useful. I would like to know if the low input impedance prevents the use of capacitor storage which can be a very useful feature of the analogue voltage method.

This is illustrated by the case of discontinuous high-speed measurements where the voltages can be stored and read at a slower rate. This leads to a more economical arrangement than that described by Messrs. Brooker and Stephens, although of course this is necessary for continuous high-speed measurements.

Mr. Jardine (*in reply*)*:* Most data collecting systems are capable of accepting digital information directly and producing outputs in decimal or binary coded decimal form for entry to a typewriter or tape punch. While data handling systems are designed primarily to accept analogue input signals, a decimal input is equally acceptable, the only difference being that no analogue to digital conversion is necessary. The programmed logical circuits associated with the analogue to digital conversion unit can still be used to sequence the print-out of the decimal input. A "time" entry, common in most data handling systems is a typical example.

I would agree with Mr. Hunter that the necessity for linearization does not arise when a computer is available. However, in many instances the plant operator requires information in real time, in which case linearization is an attractive feature. Again, one must not lose sight of the fact that many smaller users do not have the facilities of a computer.

In cases where capacitor storage is employed the store can be isolated from the instrument to some degree by employing capacitor transfer techniques in the decision amplifier. The efficiency of conversion will then depend on the ratio of the store capacitor to the transfer capacitor.

Mr. Stuttard (*in reply*): The majority of industrial instrumentation is performed with analogue transducers and hence the design emphasis is in this direction. It is certainly far easier to design a data logging system to accept digital inputs since the switching problems become much simpler.

Linearization is commonly provided in the industrial system where an immediate read-out is required and in many of these applications there is no intention of performing further analysis by computer. In the case of scientific applications where analysis is to be performed one would agree that there is no point in linearizing.

Messrs. Brooker and Stephens (*in reply*): The choice of the system described in our paper was to some extent dictated by the availability of existing sub-assemblies, which gave us an advantage in cost and delivery time to the customer.

We did investigate the capacitor storage method, but this system is only advantageous when applied to discontinuous processes requiring a single set of readings taken at high speed, followed by a slow read-out on to an off-line device. In our particular case, high-speed data logging of an experiment lasting for several minutes was required, and obviously the capacitor method was impracticable.

Mr. D. Shaw (*Associate Member*): What have the authors done to make their equipment stand up to industrial environments?

Mr. J. Jardine (*in reply*): The instruments discussed have been designed to meet moderate environmental conditions, consideration being given to hermetic sealing, transistorization, and modular construction. It would be a simple matter to extend the capability of the instrument to withstand severe environmental conditions. However, the increase in cost and size would make the instrument less acceptable to a large portion of the market.

Mr. Stuttard (*in reply*): The specifications now being written for some industrial equipments are at least the equal of military standards. Unfortunately, the majority of data logging equipments are custom built and the customer is not prepared to pay the price necessary for the extensive development needed to meet these specifications. Development along these lines has, therefore, to be carried out as a private venture and is necessarily more slow. Consideration should be given to the possibility of locating large pieces of electronic equipment in a control room or other location removed from the plant in order to prevent the problem rather than cure it.

Messrs. Brooker and Stephens (*in reply*): The equipment described in our paper is intended for use in a research establishment, where the environment is more favourable than it would be, for example, in a steel works! However, the sub-assemblies from which the equipment has been constructed have themselves been designed to operate over a wide temperature range in adverse climatic conditions.

In our experience, however, data logging equipment is not usually required to work under very adverse conditions.

In such a system, data from a large number of transducers is collected and processed at a central station and information is then presented either to a human controller or electronic device. The central station can be sited some distance from the transducers and is usually in a control room which does not suffer from the extremes of temperature and humidity present in other parts of the works. It is to the advantage of the customer and the equipment manufacturer to site the equipment in the best conditions possible as inevitably the more extreme environmental conditions specified, the higher the cost.

U.D.C. 681.39

Low-Level High-Speed Data Scanning Systems

By

R. D. BROOKER†

AND

R. B. STEPHENS

(*Associate Member*)†

Presented at the Symposium on "Recent Developments in Industrial Electronics" in London on 2nd-4th April 1962.

Summary: A high speed data handling system for a hypersonic wind tunnel is described. The system is made up of transistorized modular elements allowing system flexibility and the possibility of further extension to cover increased scanning requirements. The low level inputs to the system are commutated by dry reed relays having characteristics which yield a fast reliable scanning switch. The advantages and disadvantages of these and other switching systems are discussed. The output from the system is in the form of a magnetic tape record which allows great flexibility in the subsequent processing operations.

1. Introduction

In the field of industrial process control the monitoring of relatively large numbers of parameters is often required in order that the process may be assessed or controlled either by a human operator or some automatic controller. Because of the large number of data sources involved it has often been found more practical and economical to commutate the inputs to a single precision measurement device. The input signal levels depend on the transducers used to measure the unknown parameters, but they are quite often obtained from thermocouples and strain gauges having full scale outputs of the order of 10 millivolts. The requirements have been met by the development of numerous low-level data scanning systems and the demand for such systems for both industrial and research applications is increasing.

A typical system consists of:

(a) A low-level scanning switch which connects the transducer outputs, one at a time, to a pair of common input terminals.

(b) An amplifier which is connected to the common input terminals and which provides the desired scale factor and linearization correction as required for a given input. This amplifier is usually of a differential type providing rejection of common mode voltages induced in the signal leads.[1]

(c) An analogue-to-digital converter which accepts the output of the amplifier and converts the commutated analogue signal to a digital form.

(d) An output drive unit which contains the code converters and amplifiers necessary to drive the output device, and finally,

(e) The output device itself. The latter may take several forms dependent upon the control requirements. If a human operator is involved, then a visual display or some print-out device is used. If the controller is electro-mechanical then digital-to-analogue conversion is required to give a signal which can be used to drive a shaft or valve. On the other hand, computer control requires the information to be put on to either paper tape, punched card or magnetic tape.

The choice of a particular system for a given application depends on several factors but, in general, systems may be divided into two groups, according to the rate at which the inputs are commutated. The first group is concerned with slow speed scanning switches operating at rates of a few points/second (typically 10 points/second). This type of system is usually employed where a continuous process is involved. In these cases the plant variables have very low rates of change, a few cycles per hour being typical, so that high scanning rates are not required. The output device is quite often a printer, and routine logs of the data are made at periodic intervals. The routine logging and alarm scanning of nuclear reactor fuel and moderator temperatures with reference to preset high and low alarm levels is an example of a system in this group.

The second group is concerned with high speed scanning switches operating at rates from 100 points/second to 10 000 points/second. This group is usually applicable to discontinuous processes (i.e. those only lasting for a few minutes), or to continuous processes having transient phenomena which must be examined. Due to the high speed involved, the information must be processed immediately in a computer or stored on a suitable medium for off-line processing. Collection of data on temperature and strain in a

† Mullard Equipment Ltd., Manor Royal, Crawley, Sussex.

model under test in a hypersonic wind tunnel is an example of a system in this group.

This paper is concerned with the second group of systems and has particular reference to the various types of scanning switch which may be employed. The advantages and disadvantages of three different types are discussed and reference is made to the problems involved in their design and construction. Finally, a brief description of an actual system for a hypersonic wind tunnel is given.

2. General System Requirements

The fields of application for high-speed data scanning systems are very varied but typically they are required for:

(a) High-speed wind tunnel instrumentation, where the duration of a test is short (10 minutes is a typical maximum figure) and the information must be collected from a relatively small number of points. A system for monitoring 52 transducers at a rate of 100 points/second is described at the end of this paper.

(b) Rocket motor testing, where again the duration of the test is small (3 minutes is a typical maximum figure) but information on a large number of rapidly varying quantities must be recorded. On a single chamber engine 10 to 15 measurements must be made in order to assess the performance of the engine. The sampling rate of each variable depends on the highest frequency to be reproduced, but a figure of 100 samples/second has been suggested.[2] Multi-chamber engines, with a corresponding increase in the number of measurements required, call for scanning rates of 10,000 samples/second for 100 measuring points.[3]

(c) Examination of the transient behaviour of plant processes, for example in a large steel works. On each of the strip, slabbing, rolling and tinning mills there may be as many as six stands, each having at least four parameters varying at rates up to five cycles per second. To reproduce these variables accurately it is required to sample each at least ten times per second, thus giving rise to a scanning speed of at least 240 points/second.

(d) Medical electronic applications such as contour mapping of the electrical activity over selected areas of the brain. To obtain the desired resolution there must be at least 200 electrodes/cm^2 of the cortex, scanned at a rate of 500–1000 times/second for each point.[4]

The design of a system to meet requirements such as those outlined above will depend on a number of important parameters. These are discussed in the following paragraphs:

2.1. *Linearity*

This is the measure of the constancy of the ratio between the input voltage and the output reading over the whole required input voltage range. The relationship is based on a straight line and the linearity is usually expressed as a percentage deviation from this straight line; 0·1% or better is often required.

2.2. *Accuracy*

This is the measure of how the output reading indicates the numerical value of the input with reference to an accepted standard and is usually expressed in the units of the input as a percentage. Again 0·1% (over the whole input voltage range) is a common requirement.

2.3. *Resolution*

The smallest signal input change which will result in a change of the least significant digit at the output. 0·1% or better of the full scale input is quite commonly required which in the case of a thermocouple may mean a resolution of 10 μV in 10 mV.

2.4. *Noise Level*

The noise level is specified as the noise introduced by the system while scanning the information channels. This is usually referred to the input and must obviously be considerably less than the smallest detectable signal.

2.5. *Common Mode Rejection*

In many data scanning systems the signal leads are of necessity very long. The transducers are also grounded at some point remote from the system ground, and consequently a potential, due to the ground loops, may exist between the two. This potential usually has both d.c. and a.c. components. The a.c. signals are at the power line frequency and as they are common to both signal leads, these unwanted signals are known as "common mode" voltages. Voltages of the order of tens of volts are not unknown and the system must be capable of rejecting such spurious signals.

2.6. *Input Impedance*

The desired input impedance depends on the source impedance involved and the accuracy required. Thermocouples and/or strain gauges have very low source impedances, but where long leads and loading networks are involved a value of 100 Ω may be typical and if an accuracy of 0·1% is required it is obvious that the input impedance must be at least 100 kΩ.

2.7. *D.C. Isolation between Channels*

In some cases adjacent channels may be at vastly different potentials. In the steel works application for example, where the measurement of several different control voltages obtained from Ward Leonard sets is required, the difference may be as high as 1 kV.

2.8. *Cross-talk between Channels*

Information on adjacent channels must not affect the operation of a channel selected by the scanner. The open and closed resistance of the scanning switch is the determining factor in most cases, and in general the system must be capable of accepting full-scale input on any one channel without affecting the input on any other selected channel by more than an amount corresponding to the resolution of the system.

2.9. *Zero Drift*

For most systems the output is required to be zero when the input is zero. Changes in ambient temperature or working conditions may give rise to long and/or short term drifts in the zero, and again this must be less than the resolution of the system.

2.10. *Operational Reliability*

For high-speed systems the utilization of the equipment may not be very high. Nevertheless the system should be as reliable as possible, although it may be periodically checked and set up prior to a run and needs only to retain its accuracy etc., during the period of the run.

2.11. *Output Requirements*

By this is meant the selection of a suitable storage device or on-line processor. The storage device is usually magnetic, either a tape recorder or ferrite store. In this case all the information is stored for further processing. The on-line device is normally a computer allowing processing while recording is taking place.

3. Scanning Switch Requirements

All the above items, with the exception of 2.11, must be taken into consideration when the design of a scanning switch is undertaken. In addition the maximum scanning rate required must be considered as all switches have a limiting speed of operation governed by factors such as life, transient behaviour etc. Particular attention must also be paid to the following three characteristics.

3.1. *Electrical Noise*

The electrical noise associated with a switch may be generated externally or internally. In the case of external noise, it is dependent on the source impedance involved and is usually at power line frequency. Adequate electrostatic and electromagnetic shielding usually eliminates this type of noise. Internally generated noise is caused by contact bounce at break and make, variation of the closed contact resistance, break-through of the driving waveform, thermal e.m.f.s, and capacitance of the switch. Contact bounce and break-through of the driving waveform do not usually cause very much trouble providing the duration of the noise generated is short compared with the sampling period. In such cases it can be gated out in the amplifier stage following the switch. Variations in the contact resistance, however, if significant in relation to the amplifier input impedance, do lead to noise generation which may effect the resolution of the system. Thermal e.m.f.s are caused by temperature gradients within the switch assembly and great care is required in the design stage to avoid such gradients at junctions of dissimilar metals, e.g. soldered joints which generate e.m.f.s of the order of 2 μV/deg C. Capacitance associated with the switch is also of importance especially when high source impedances are involved, as it leads to considerable make and break delay times in the switch.

3.2. *Environmental Conditions*

In many industrial applications the equipment may be operated under quite high ambient temperature conditions (35° C is not unusual) and in general, it is desirable to design a switch which will operate satisfactorily over a range $-10°$ C to $+60°$ C. Extremes of humidity and contamination of the air with dust or vapour lead to the need for hermetic sealing. Mechanical shock and vibration may also be a factor to be considered, but in general it is to the equipment manufacturer's advantage to obtain the customer's help in siting the equipment in a reasonable environment.

3.3. *Operating Life*

Ideally the scanning switches should be capable of operating for the whole of the life of the equipment (say between 50 000 and 100 000 hours) without adjustment or replacement. In practice this ideal may not be realized and some maintenance may be required. Thus electromechanical switches will have a definite life expectation in terms of total number of operations before contact adjustment is required or before complete failure will occur. The scanning speed will therefore determine the interval between scheduled maintenance. Again, with static switches long term changes in transistor parameters, for example, may necessitate re-adjustment of balance controls in order to restore the switch zero to its original value.

4. Scanning Switch Types

Consideration of all the above factors has led to the design of several types of scanning switch, which may be divided into three different groups.

4.1. *Electro-Mechanical Types*

4.1.1. Stepping switches

Gold-plated stepping switches have been designed to operate at speeds of 100 steps/second, but in general their use in high-speed systems is limited by life considerations.

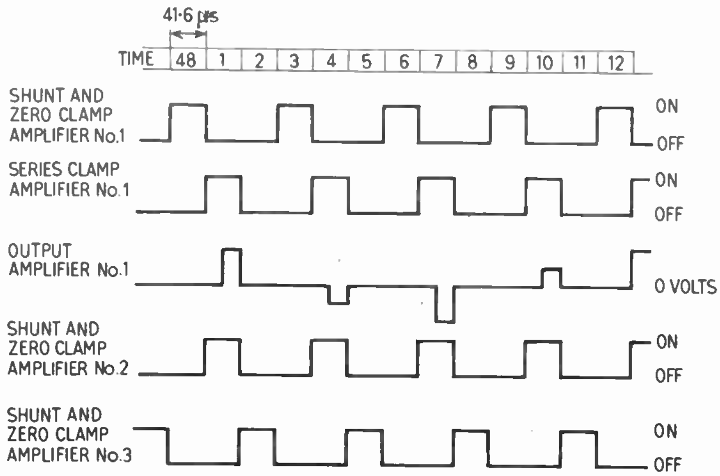

Fig. 3. Timing diagram.

high speed system, 48 inputs may be sampled at a rate of 500 samples/second. Referring to Fig. 2, the inputs are divided into six groups of eight. Each group, consisting of an input transformer having eight input windings, feeds one of three a.c. differential amplifiers which are gated as required to connect the selected input to the analogue-to-digital converter. The input transformer provides isolation between channels (1 kV) and high common mode rejection up to 180 volts (10^6 : 1), but it does introduce a problem in the design of the system. During a sample period energy is stored in the transformer core material. This energy must be dissipated before the next sample period and this necessitates multiplexing of the amplifier channels.

From Fig. 3 it is seen that as each point is sampled the appropriate amplifier is zeroed by shorting both the input and output terminals of the unit, thus restoring the d.c. reference. After a period of 41·6 µs the shunt and zero clamps are removed. At the same time the input gate is closed and the series clamp is also made. The input is thus applied to a newly zeroed amplifier. After a delay of 20 µs to avoid switching transients, the output gate is closed connecting the amplifier to the a.-d.c. At the end of the gate period, the input gate and series clamp are opened. The transformer is then allowed to recover during the time that the other five units are in operation. The series clamp is provided so that all loads are removed from the transformer, speeding up the recovery time. This system is installed in several plants in the U.S.A. and with its resolution of 10 µV for full-scale inputs of ± 5 mV, seems to be a solution to some of the semiconductor switching problems mentioned above.

G. W. Gore[14] describes another development of the Dorsett and Searcy circuit. In this circuit (Fig. 4) the transistors are connected as shunt elements rather than as series elements. Two resistors R1 and R2 are selected having values large with respect to the resistance of the transistors in their bottomed condition. Additional resistors R3 and R4 are connected between the emitter and collectors of the transistors having values which are large compared with R1 and R2. This enables R3 and R4 to swamp any changes in the internal resistances of the transistors, due to temperature say, when they are in the "off" state. Resistor R5 is variable and is connected between the two bases of the transistors allowing balancing of the switch when non-matched transistors are used. Using OC200 transistors with R1 = R2 = 50 kΩ and R3 = R4 $= \frac{1}{2}$ MΩ a successful switch is possible having "off" and "on" states which are only slightly affected by temperature. The main disadvantage is the need for a very high impedance amplifier to follow the switch since the effective source impedance of the signal is at least 100 kΩ. Crosstalk could also be troublesome when the number of points connected to a system exceeds about fifty.

4.3. *Magnetic Switches*

A commercially available system in the U.S.A. is based on a unit known as the Magne-Plexer.[14] In this system each input is connected to an individual

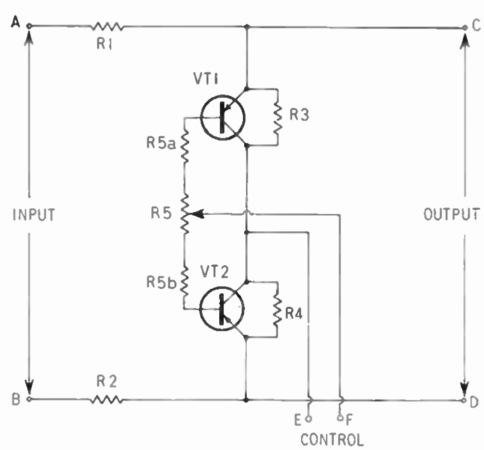

Fig. 4. Gore switch.

subminiature magnetic amplifier. Carrier current is supplied to only one amplifier at a time under control of a solid-state carrier gating matrix. Since carrier current is only applied to one magnetic amplifier at a time, zero e.m.f. is developed in the output windings of the other amplifiers. All the output windings can therefore be joined together and after amplification the resultant output signal is detected to provide a voltage output of 5 V for full-scale input. Overall feed-back stabilizes the linearity and accuracy (0·1%) and reduces the noise in the system (0·1% f.s.d.). Since a magnetic amplifier is basically a current-sensitive device, the full-scale input voltage is related to the resistance in the input circuit. The amplifiers have a nominal full-scale input of 50 μA and the windings have a d.c. resistance of 40 Ω. Thus assuming a thermocouple input with a source resistance of 100 Ω, full-scale output is obtained for 7 mV input. The amplifier requires at least 1 ms for stabilization but, by interlacing between input modules and output gates, rates corresponding to 50 μs per point can be obtained. A typical system has 96 inputs scanned in 16 ms. Common mode rejection is 120 dB with an input unbalance of 330 Ω between one side and ground. The isolation between inputs is 150 MΩ up to 180 V d.c. Crosstalk on to the selected channel from full-scale inputs on the remaining channels is less than 0·1% of full-scale deflection.

5. Conclusions on Switch Development

The foregoing paragraphs give a survey of the types of low-level high-speed switch developed or under development both in this country and the U.S.A. To the present date the semiconductor switch has not found very many applications due to troubles experienced with changes in ambient temperature and in transistor characteristics with life. Recent American developments and the introduction of new silicon chopper transistor types in this country suggest that in the near future the semiconductor switch will be used more frequently. Until such time, in this country at least, the authors feel that the best solution to the problem is the use of the dry reed or mercury-wetted contact reed relays.

6. A Data Handling System for a Hypersonic Wind Tunnel

To conclude the paper a brief description of a system developed by the authors is given below.

6.1. *System Description*

In a hypersonic wind tunnel speeds of up to a Mach number of 9 are obtained by charging air reservoirs to a high pressure and then allowing the air to escape at a controlled rate to a low-level pressure vessel through a small working section containing the model under test. The tests only last for a short time and during this time several temperature and strain measurements are required. For a particular wind tunnel a data handling system with the following specification was developed.

Number of inputs: Up to 52 points can be selected for scanning. The inputs are arranged in module form so that extension of the system can be made without major modification.

Linearity: Better than 0·2%.

Resolution: 10 μV.

Input ranges: 0–10 mV, 0–20 mV and 0–40 mV.

Speed: 50 points in ½ second.

Scanning modes: Continuous or single shot.

Input transducers: Thermocouples and strain gauges. A low level pressure transducer (Statham gauge) and manual selection of a high level pressure transducer is also included.

Output: Normal mode of operation is that of scanning 100 points/second and recording of the values in five bit code on magnetic tape.

Teleprinter output: Following recording, for local assessment, a teleprinter is used as a print-out device from the tape.

Visual read-out: Selection of a point for examination or calibration is possible, the output being displayed as three decimal digits plus point number on cold cathode indicators.

Analogue recording: A strip chart recorder can be connected to any one input by manual selection on the input modules. To avoid loading of the point during scanning the chart recorder is switched out while a measurement is being recorded on the tape.

Checking facilities: The inputs to the scan switches can be set to earth or to a calibration voltage set up against a standard cell. The calibration voltage is also wired to the first point in the scan cycle so that the system is checked once per complete scan.

6.2. *Equipment Design*

The system is made up of the following units.

(a) Scan modules
(b) Amplifier unit
(c) Analogue-to-digital converter
(d) Scan drive unit
(e) Writing unit
(f) Pressure scan unit
(g) Calibration unit (with control panel)
(h) Visual display unit
(i) Power supply unit
(j) Magnetic tape recorder
(k) Teleprinter.

With the exception of items (j) and (k), the units are mounted in a 6 ft 19 in. rack. Units are removable from the front of the rack and all connections are at the rear. All controls such as calibration voltage setting, scanning mode selection, and tape recorder control push buttons are mounted on the front panels of the units.

6.3. *Unit Descriptions*

Referring to the block diagram in Fig. 5, the following is a brief description of the main units of the system.

6.3.1. Scan modules

Each module is made up of four pairs of dry reed relay inserts mounted in four coil formers. The inputs to the module (from four transducers) are connected via screw terminals to the amplifier unit.

A switch is provided on the front of each module for selection of calibration, earth and chart recorder facilities. Driving signals to the relays, together with calibration voltages, are connected to the module by means of a gold-plated connector.

6.3.2. Amplifier unit

After selection by the appropriate relay the input signal is amplified by an American transistorized differential amplifier of advanced design[1] having the following specification:

Gain: 10 to 1000 in ten steps

Accuracy: 0·02%

Linearity: 0·01%

Stability: 0·01%

Bandwidth: d.c. to 100 kc/s

Common mode rejection ratio:
1000 × gain setting for d.c. to 150 c/s up to a maximum of 10 V.

Drift (referred to the input):

$$\pm\left(10+\frac{300}{\text{gain setting}}\right)\mu\text{V}$$

Input impedance:
1000 MΩ either line

Output: ±10 V at 100 mA.

The amplifier uses a new type of input circuit, which might be called a dynamic bridge. Unlike normal bridge amplifiers, the common mode rejection increases with gain and precision resistors are not required to meet this specification.

6.3.3. Analogue-to-digital converter

The analogue signal from the amplifier is converted to a 12-bit binary/decimal digital signal in a transistor a.-d.c. Brief specification details of this unit are:

Analogue input: 0–10 V positive

Input impedance: 8 kΩ

Output: 12-bit binary coded decimal parallel or serial.

Digitizing time: 1 ms

Accuracy: 0·1%

The fast digitizing time allows nine of the ten milliseconds available per point to be used for switching of the point to the amplifier, contact bounce, and settling of the amplifier.

6.3.4. Scan drive unit

Two sets of unique pulses are derived from circuits counting down from a standard 100 c/s square wave. These pulses are applied to the sides of an 8×7 matrix, energizing in sequence the reed relay coils. The reset point of the counter is controlled by switches on the front of the control panel, allowing for a variable number of points to be scanned.

6.3.5. Writing unit

This unit contains an 800 c/s clock pulse generator with a stability of ±2 c/s over a temperature range of 40 deg C which provides all the timing pulses for the recording of the information in the correct sequence on to the tape. It also provides the 100 c/s square wave used to drive the scanner. In addition all the

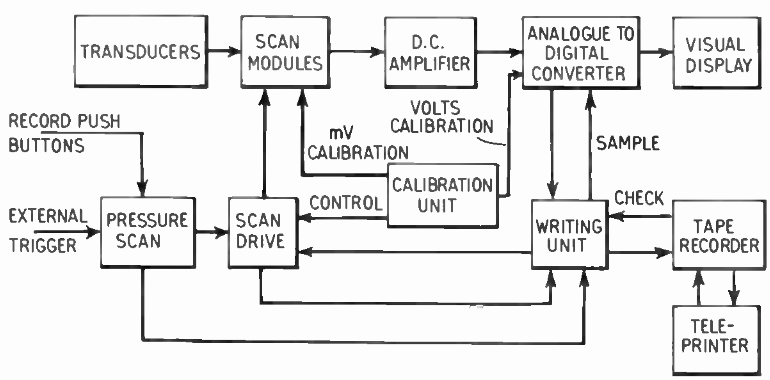

Fig. 5. Block schematic of data handling system for hypersonic wind tunnel.

logical operations necessary to convert the 12 bit parallel information into 5-bit serial form are carried out in this unit, together with the code conversion and warning character generation circuits.

Each word for a given point consists of the following seven characters:

(1) A character denoting start of scan or alternatively a "space" signal. As the information is processed in a computer point number information is not required, but a character at the start of each scan cycle is put on to the tape.

(2)
(3) Three decimal digits corresponding to the input on the point.
(4)

(5) Tunnel "on" or tunnel "off". A pressure pick-up in the tunnel is used to generate one of two characters depending whether the tunnel is running or not.

(6) Carriage return or space. In order to present a suitable print out, at the end of each block of eight words a carriage return signal is put on to the tape.

(7) Line feed or space. Line feed is put on the tape at the end of every eight blocks.

The logical operations necessary for the code conversion and word generation are carried out with NOR type logic using standard transistor NOR logic elements.

6.3.6. Pressure scan unit

This unit contains circuits which delay the recording of the information after starting of the tape deck in order that the tape may reach its proper speed. Automatic control of the scan is also carried out by circuits initiated by an external pressure trigger signal.

6.3.7. Calibration unit

All the controls for operation of the system including calibration voltage setting are contained on this unit. The calibration voltages are obtained from a transistor stabilized supply using a Zener reference diode source and are adjustable to within 2 mV of the desired levels by means of a centre zero meter which has, as reference, a standard cell. The latter is also contained in this unit.

6.3.8. Visual display unit

The visual display is only used on manual operation and contains a relay decoder (from binary coded decimal to decimal) to drive the cold cathode indicators.

6.3.9. Power supply unit

The supplies for the system are made up of a positive and negative 24 volts supply for the logical blocks, a variable stabilized calibration supply, and a 250 V supply for the cold cathode tubes. With the exception of the latter, standard units are used.

6.3.10. Tape recorder

A modified version of a tape recorder originally developed at Manchester University is employed. Brief specification details are:

Number of tracks:	8
Speeds:	$\frac{1}{16}$ in/s, $\frac{5}{16}$ in/s, 2 in/s and 10 in/s.
Heads:	Flux gate type with output independent of tape speed.
Packing density:	100 bits/in.
Recording rate:	1 kc/s at 10 in/s.
Recording modes:	Continuous or character by character at any one given speed.
Play-back facility:	To teleprinter at $\frac{1}{16}$ in/s if information is recorded at maximum rate.
Checking facility:	Recording heads read the information just recorded and circuits compare it with stored signals. If it is incorrect a signal is sent to the data handling rack causing the following character to be overwritten by an erase.

6.3.11. Teleprinter

The teleprinter is a standard 5-wire input, 5-wire output machine. The keyboard allows editing of the recorded tape and addition of information such as date, run, number etc.

7. Conclusions

A high-speed data handling system for a hypersonic wind tunnel has been described. The system is made up of transistorized modular elements allowing system flexibility and the possibility of further extension to cover increased scanning requirements. The low-level inputs to the system are commutated by dry reed relays having characteristics which yield a fast reliable scanning switch. The output from the system is in the form of a magnetic tape record which allows great flexibility in the subsequent processing operations.

Particular attention has been paid to the design of a suitable low-level high-speed switch. In the authors' opinion, a satisfactory solid-state switch is the ultimate aim of future development. The introduction of new types of transistor should enable this aim to be achieved, but until a semiconductor switch with characteristics equivalent to, or better than, the reed relay is developed, the latter will find extensive application in many systems during the next few years.

8. Acknowledgment

The authors would like to thank the directors of Mullard Equipment Ltd. for permission to publish this paper.

9. References

1. S. C. Brown, "Wide band differential amplifiers", *Electrical Design News*, January 1961, pp. 109–13, and February 1961, pp. 84–7.
2. F. K. Williams, "Requirements of a high speed, all electronic, fully automatic data handling system", *I.R.E. Conv. Rec.*, **4**, pp. 140–2, 1956.
3. M. C. Klein and R. B. Rush, "Techniques for a high speed, high quantity, data processing system—Idiot 11", *I.R.E. Conv. Rec.*, **4**, pp. 143–52, 1956.
4. C. Barus, "Low level high-speed switching system for brain mapping", *Trans. Inst. Radio Engrs* (*Medical Electronics*), No. PGME-7, pp. 1–13, December 1956.
5. J. G. Bannochie and R. A E. Fursey, "Sealed contact relays", *J. Brit.I.R.E.*, **21**, pp. 193–9, February 1961.
6. G. P. Bently and S. Ackerman, "Commutation switch developments", *Instruments and Automation*, **31**, pp. 1368–70, August 1958.
7. W. R. Davis, "Exploiting the mercury jet switch", *Control Engineering*, February 1957, pp. 69–72.
8. "Scanners", *Instruments and Automation*, **31**, pp. 1352–6, August 1958.
9. G. G. Haigh, "The use of mesa diodes in small-signal modulators", *J. Instn Elect. Engrs*, **7**, pp. 497–9, August 1961.
10. A. P. Kruper, "Switching transistors used as a substitute for mechanical low-level choppers", *Trans. Amer. Inst. Elect. Engrs*, **74**, Part 1, pp. 141–4, 1955. (*Commun. Electronics*, No. 17, March 1955).
11. R. L. Bright, "Junction transistors used as switches", *Trans. Amer. Inst. Elect. Engrs*, **74**, Part 1, pp. 111–21, 1955. (*Commun. Electronics*, No. 17, March 1955).
12. E. Dorsett and J. H. Searcy, "Low level electronic switch", *I.R.E. Conv. Rec.*, **5**, pp. 57–60, 1957.
13. C. E. Griffin, J. P. Knight and J. H. Searcy, "Low level multiplexing for digital instrumentation", *Electronics*, **33**, No. 41, pp. 64–6, 7th October, 1960.
14. "Improvements in or relating to selector circuit arrangements", British Patent No. 865,899 (G. W. Gore, Mullard Ltd.).
15. "High speed scanner", *Electrical Manufacturing*, May 1960, pp. 199–200.

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A report of the Discussion which followed the reading of this paper is given on page 275.

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A Direct Current Integrator with Digital Output

By

R. C. MARSHALL, M.A. †

Presented at the Symposium on "Recent Developments in Industrial Electronics" in London on 2nd–4th April 1962.

Summary: This paper outlines a number of ways of integrating with respect to time a 0 to 10 mA d.c. signal, which may be derived from an industrial measurement or control system. A new solid-state electronic integrator with superior stability and linearity is described in detail. A typical application is the measurement of total quantity of a liquid which has passed through a magnetic flowmeter.

1. Introduction

Current practice[1] for industrial process control is to employ a number of individual units, each performing a single function and operating on a common transmission signal.

The primary detecting element, such as a thermocouple or ionization chamber is followed by an amplifier to convert the signal into a standard transmission signal. Separate indicators, recorders and alarms can then monitor this standard signal. A controller can compare the signal with a desired value and operate a regulator to control the process.

A defined transmission current is often more useful than a defined voltage, as no loss of accuracy results from line resistance with a finite load resistance. Magnetic amplifiers and transducers in which a force is produced by a solenoid are both basically current sensitive, and fit naturally into a defined current system. As they are normally wound with copper wire the voltage sensitivity has a large temperature coefficient. A transmission current of 0–10 mA is recommended by a draft British Standard Specification.

A converter able to produce such an output is shown in Fig. 1. The input signal is applied to a d.c. amplifier A, whose output voltage is converted into a direct current of 0–10 mA by the output transistor VT1 and its emitter resistor. This current I is fed through a feedback resistor r to stabilize the gain. It also flows through a bridge rectifier B1 which, together with a 200 c/s square wave supply provides a floating h.t. supply for VT1. The alternating current flowing to the bridge is also passed through a low loss current transformer, and reconverted to d.c. by a second bridge rectifier B2. A floating output current is thus provided which is fed to the various load units in series. The maximum output is 20 volts, so that the units connected in series with its

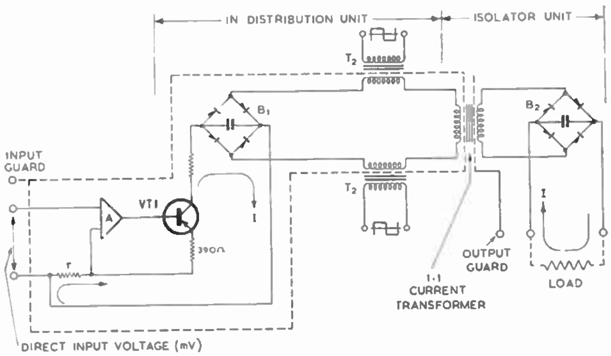

Fig. 1. A converter with current output.

output are generally arranged to have a voltage drop of 5 volts or less.

The integral with respect to time of the transmission signal is often required, as in the measurement of total quantity of a liquid that has passed through a magnetic flowmeter, or the total heat output of a boiler.

This need can be met by a device that indicates

$$K\int i\,dt$$

where the integration time varies from a few seconds to several years. The design of an integrator for use with the transmission system outlined above is discussed in this paper.

2. Available Types

Some commonly used electro-mechanical integrators will be considered to provide a background to the work described later in the paper. Two of these convert an input current or voltage into speed of rotation of a shaft, gear this down to a lower speed, and add the number of revolutions on a mechanical counter.

The d.c. ampere-hour meter incorporates a permanent magnet motor with its armature enclosed by an aluminium disk that rotates with it, and provides

† George Kent Ltd., Luton, Bedfordshire.

eddy current damping to give speed proportional to current. An undamped motor can also be used, giving a speed proportional to voltage. This runs at a considerably higher speed than the damped motor, with greater wear. To minimize friction, the brushes of both types only bear lightly on their commutators with consequent risk of high resistance troubles in industrial atmospheres. The operation of the commutator reflects a fluctuating load on the signal. The most accurate d.c. machine integrator uses a tacho-generator, driven by a servo-motor and amplifier so that its output voltage is equal to the signal voltage. By this means frictional errors can be eliminated, though the cost is high.

A sampling type of integrator is frequently used in conjunction with potentiometric recorders. Its principle is illustrated in Fig. 2. A continuously-rotating cam allows light to fall on a photocell for a fraction of the time that varies with the radial position of the lamp. The lamp is moved by the recorder pointer. Illumination of the photocell engages a clutch which connects a synchronous motor to a reduction gear and mechanical counter. Thus the counter is driven at a constant speed for a fraction of the time proportional to the deflection of the recorder pointer. The synchronous motor causes an error equal to the supply frequency error, and fluctuations of input that occur in less than one revolution of the cam are not correctly recorded. It is easy to provide any required law between input and output by shaping the cam. A square-root law is necessary if the recorder measures differential pressure across an orifice to indicate flow, and the total volume is to be shown on the integrator. None of the other integrators described can give other than a linear law, but in a modular electronic system non-linear laws are obtained in converters prior to the generation of the 0–10 mA transmission signal.

3. Electronic Integrators

The electronic integrators to be described are analogous to the electro-mechanical ones mentioned

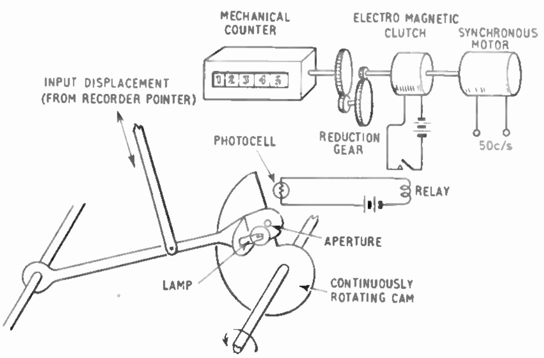

Fig. 2. General principle of the cam integrator.

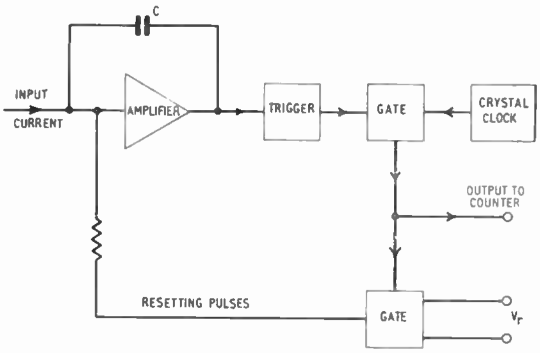

Fig. 3. A precision integrator.

above. They convert current (or voltage) into frequency, divide this down to a lower frequency, and add the number of cycles on an electro-mechanical counter. For summation over a long period the purely electronic counter is not practical as electronic display of the integral is too cumbersome. "Electronic" integrators operate an electro-mechanical counter at a speed giving the required life. This is usually between 10 counts per second and 0·1 counts per second. The electro-mechanical counter can be preceded by an electronic counter and display, if required to give very high resolution and long life. Some dividers are generally necessary to permit the use of practical values of components in the integrator itself.

The integrator may be based on the equation of the input current to a voltage repeatedly charging an inductor through a fixed flux range, or a number of pulses of fixed duration and current, or a current repeatedly charging a capacitor through a fixed voltage range. These three can be expressed as:

$$I = \frac{1}{R}\int V\,\mathrm{d}t = \frac{n\phi}{R} = \frac{nAB_{\mathrm{ref}}}{R}$$

$$I = \int i\,\mathrm{d}t = ni_{\mathrm{ref}}\,t_{\mathrm{ref}}$$

$$I = \int i\,\mathrm{d}t = nQ = nCV_{\mathrm{ref}}$$

where R is the gain (transfer resistance) of the current/voltage converter.

3.1. *Inductor Integrator*

The attraction of the inductor is that a saturable core will itself act as the reference which is required to set the limits of the integration. The Royer oscillator[2] uses this principle to give a frequency proportional to voltage. By using an alloy with a low temperature coefficient of saturation flux density,[3] it should be possible to produce a simple voltage integrator competitive with the d.c. motor types. It is difficult to make inductors as nearly ideal as capacitors, and when the basic input is a current, the accurate conversion to voltage in the face of the loading due to the inductor is troublesome, so this method has not been considered further.

3.2. *Precision Pulse Integrator*

The comparison of the input current with a number of current pulses of known amplitude and duration is complex but accurate. An integrator has been described[4] (Fig. 3) that uses a chopper-stabilized integrating amplifier, and generates precise resetting pulses of the required polarity whenever the amplifier output exceeds an arbitrary limit. The pulse duration is determined by a crystal clock, and the pulse current by a temperature controlled Zener diode reference. The stability and linearity are of the order of 0·01%.

3.3. *Capacitor Shorting Integrator*

One capacitor charging arrangement (Fig. 4) allows the capacitor to charge to a reference voltage V_r when it is discharged suddenly, and allowed to charge again. The output frequency is thus:

$$f = \frac{i}{CV_r}$$

if the discharge time is neglected. A practical design to achieve this is shown in Fig. 5. It includes a cascade grounded base amplifier VT1–3 (see Section 4.3)

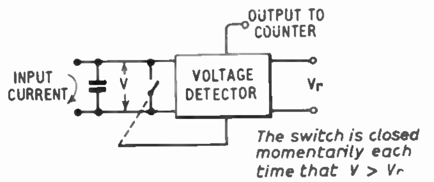

Fig. 4. Block diagram of a capacitor-shorting current/frequency converter.

that isolates the voltage swing on the capacitor from the input circuit. A complementary monostable circuit VT4 and 5 is triggered when the capacitor voltage exceeds V_r, and the power transistor VT6 then discharges the capacitor. With a 10 mA input, 25 volts reference and an 8 μF capacitor the charging time is 20 milliseconds, and the discharge time 70 microseconds. The non-linearity due to finite discharge time thus reaches 0·35% at full scale. The peak discharge current is about 4 amperes. This high current is the principal difficulty of the circuit as it imposes high stresses on the capacitor and VT6. The power supply has to provide 0·5 amperes base current for VT6, and thus needs large capacitors across its output.

The need for low leakage current when OFF, and low voltage drop at high peak currents when ON makes the choice of VT6 difficult. Transistor manufacturers do not seem to have given much consideration to pulse ratings.

A circuit using a *p-n-p-n* diode as combined reference and switch has been described[5] that overcomes most of these difficulties. However, the

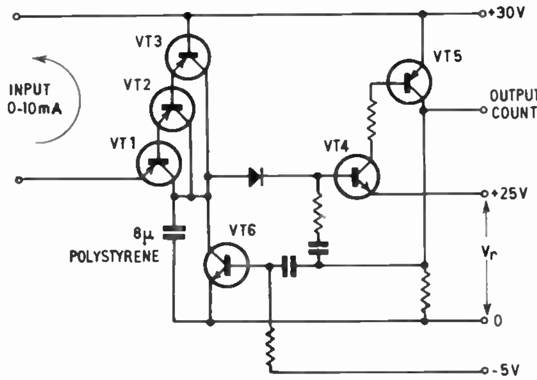

Fig. 5. Circuit of a capacitor-shorting current/frequency converter.

temperature coefficient is variable, and the starting current depends on temperature and cannot be adjusted for different applications.

4. The Chosen System

A more complex but lightly stressed circuit is shown in block diagram form in Fig. 6. This system has been chosen for development and will be described in detail. When the capacitor has been charged by the input to the reference voltage its connections are reversed, and it discharges at the rate defined by the input. Its connections are then reversed again, and the cycle repeats. The frequency is:

$$f = \frac{i}{2CV_r}$$

that is, half that of the previous design. Since in practice several binary dividers are needed between the integrator and counter, this saves one divider stage.

4.1. *The Reversing Switch*

The reversing switch circuit is shown in Fig. 7. If VT7 is held on, current flows out of the capacitor, through the input circuit, and into VT7 together with the 1 mA current i_2. The input current returns to the capacitor by way of the power supply. This is shown by the solid arrows. The voltage across D7 cuts off

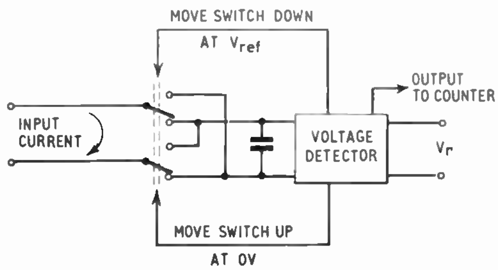

Fig. 6. Block diagram of a capacitor-reversing current/frequency converter.

VT8. If VT7 is off, i_2 supplies enough current to VT8 base to bottom this transistor, so that it can supply current into the capacitor. This current also flows in the input circuit, though, because of the diode bridge D1–4 its direction there is unchanged as shown by the dashed arrows. The actual value of the current is defined by the input circuit. In the complete circuit, Fig. 8, the cascade grounded base stage VT1–3 precedes the diode bridge (see section 4.3). VT4 and VT5 form a complementary bistable circuit: VT4 is switched on when the capacitor voltage exceeds $V_r + V_b$, and VT6 turns VT4 off when the capacitor voltage falls below V_b. The voltages V_a and V_b ensure that there is always some collector voltage for the cascade grounded base stage. The waveforms in this circuit are shown by Fig. 9.

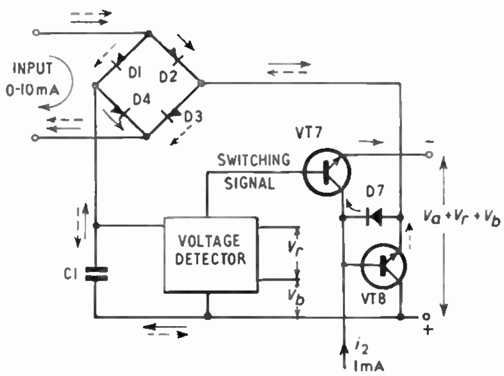

Fig. 7. The capacitor reversing switch circuit.

4.2. *The Cut-off Arrangements*

To switch VT4 off, the collector of VT6 must supply so much of the current through the feedback resistor R4 which is holding VT4 on, that VT4 is returned to its linear region. The emitter current i_{e6} is then $i_{e6} = \frac{1}{\alpha_6}(i_{R4} - i_{b4})$

where α_6 = common base current gain of VT6

i_{R4} = current through R4

i_{b4} = base current of VT4 at which loop gain of VT4 and VT5 exceeds unity.

Thus $i_{e6} \simeq \frac{V_a}{R_4}$

If the input is less than this, VT4 will never switch off, and no further integration will occur. This initial value of input current is called the cut-off current. It can be selected by choice of R4. This is important because, when the quantity measured is zero for long periods of time, the integrator must be prevented from summing whatever zero error appears

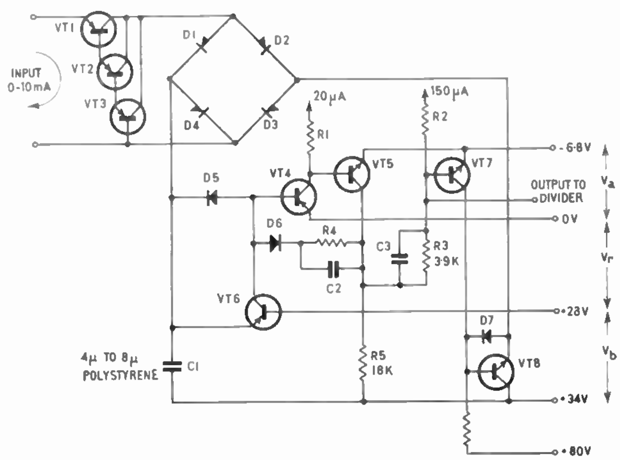

Fig. 8. Circuit of the capacitor-reversing current/frequency converter.

in the transducer or its associated amplifier. In the case of a differential pressure transducer followed by a square-rooting device, this may be as high as 10% of the full-scale value. To avoid error with signals just larger than the possible zero error, an overall characteristic like Fig. 10 is needed.

This is obtained from the circuit described. Since the cut-off current is sampled only for a small portion of each cycle just as VT6 starts to conduct, the loss of current integrated over the whole cycle is quite small. The measured performance is shown in Fig. 11.

4.3. *The Cascade Grounded Base Amplifier*

The cascade grounded base amplifier VT1–3 is a very useful component for current operated instruments. It provides a virtually infinite output impedance that can have a high voltage dropped across it, whilst only 1·5 volts are dropped across the input terminals. In this application the collector voltage varies from a volt or two to just over V_r, a swing of some 30 volts (see Fig. 9). By cascading

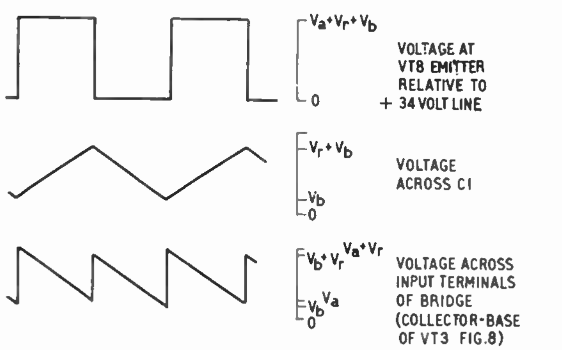

Fig. 9. Waveforms corresponding to Figs. 7 and 8.
NOTE.—All transistors and diodes are assumed to be perfect switches.

grounded base stages, the current gain approaches closely to unity, even with low-gain high-voltage silicon transistors. For three transistors, the overall current gain is $1-1/\beta_1\beta_2\beta_3$ where β_1, β_2, and β_3 are the common emitter current gains of the transistors. Figure 12 shows how the fall in gain at low currents of the transistors affects the overall performance. When low-voltage high-gain transistors can be used, two

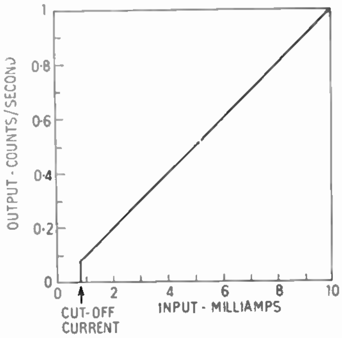

Fig. 10. Integration characteristic required for systems with a large zero error.

transistors are adequate. The third only contributes a current gain of about times five as its collector current is extremely low. An interesting property of the circuit is that the tolerance on $1/\beta_1\beta_2\beta_3$ is less than the product of the individual tolerances on β_1, β_2 and β_3. If β_1 and β_2 happen to be large, then the collector current of the third transistor will be very low, and its gain also very low. This partly compensates for the high values of β_1 and β_2.

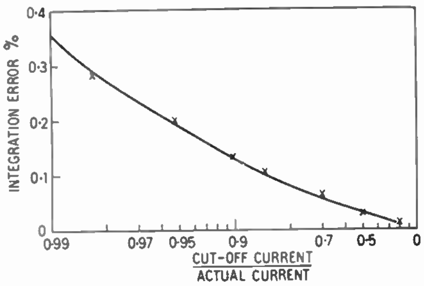

Fig. 11. Measured integration error due to cut-off current.

4.4. *Accuracy*

The integrator reading is reduced by currents lost as leakage in the diode bridge. At 70° C these would total less than 2 μA. Leakage of the cascade grounded base stage adds to the input current. This swamps the negative errors giving the rise in sensitivity at high temperatures shown by Fig. 13. Leakage through D5 and the base-emitter diode of VT6 when these are cut off adds to the input on one half-cycle, and subtracts on the other half-cycle, so that the final error can be neglected.

The delay between the capacitor reaching the required voltage and the diode bridge being switched over is about 2·5 μs. This means that the capacitor charging time is extended, and consequently the discharge time also. As this happens at each limit, an error of 10 μs occurs in a cycle of 40 ms, that is 0·025% error at full rate. The effect has been observed with an 0·2 μF capacitor substituted to give a cycle time of 1 ms. An error of 1% would then be expected: 0·9% has been measured.

Capacitance between the input circuit and the remainder of the integrator appears in parallel with the main capacitor on one half cycle, and is short-circuited on the other. It consequently has an effect on calibration, though with the normal 8 μF main capacitor 0·01 μF stray capacitance is needed to produce 0·06% error.

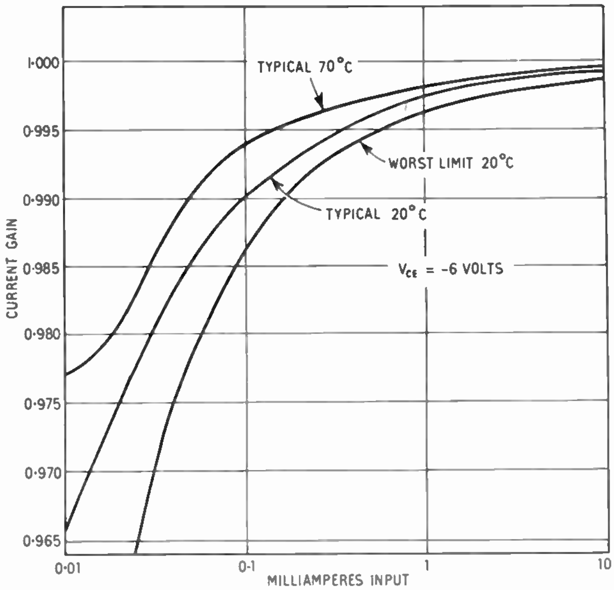

Fig. 12. Performance of triple grounded base amplifier.

4.5. *Temperature Compensation*

The polystyrene capacitor has a negative temperature coefficient, and the fall in forward voltage drop of D5 and the V_{be} of VT4 and VT6 cause an effective fall in reference voltage with rising temperature. The resultant of these is a negative coefficient of −130 to −260 parts per million per deg C. This is corrected by a Zener diode in the reference power supply, to give a resultant of ±85 parts in 10^6/deg C. Over a 30 deg C range this gives an 0·25% change in calibration. The use of silicon devices makes possible operation at 70° C ambient, though at this temperature, the uncertainty of low current integration is increased. The effect is a reduction of error, as leakage

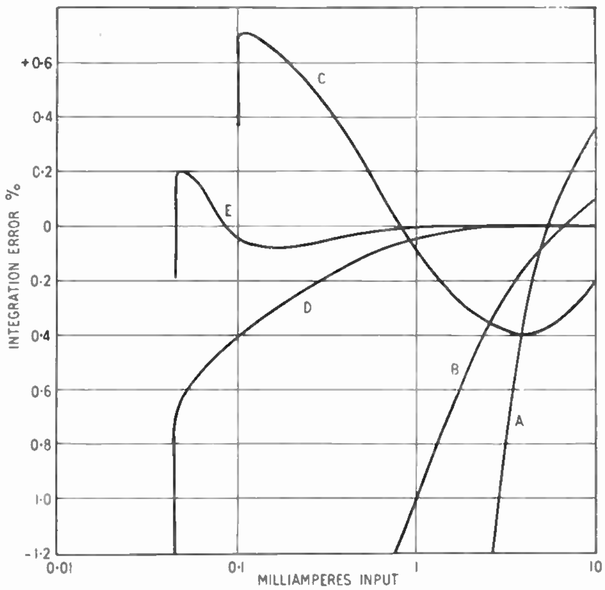

Fig. 13. Linearity of various integrators.

A Eddy-current damped mA-hr meter (measured).
B Low inertia integrating motor (half maker's limits, assuming exact current/voltage conversion).
C Cam type mechanical integrator (measured, excluding recorder errors).
D Capacitor reversing integrator 20° C ambient (measured).
E Capacitor reversing integrator 65° C ambient (measured).

and rising gain in the cascade grounded base stage make up for the normal lack of gain at low currents (see Figs. 12 and 13).

The linearity of this integrator compares favourably with that of its competitors. Figure 13 shows the true error (not the error referred to full scale), as a function of input current. The various instruments plotted are random samples of their type, and are not meant to represent the best or worst that can be encountered. The two motors behave similarly, with increasing negative errors at low currents due to friction. The mechanical cam integrator has errors due to dimensional tolerances that go either side of nominal. The electronic integrator has a negative error at low current due to the imperfections of the cascade grounded base stage. At high temperatures, leakage errors exceed this and the errors become positive. From 1/10th full scale upwards there is no detectable non-linearity.

4.6. *Ancillary Circuits*

The dividers and power supplies are conventional. The last divider is capable of driving two electro-mechanical counters.

Two counters are often convenient. One can be read and reset daily, and the other, with more digits but no reset facility, provides a permanent record. It is possible to fit a print-out counter, which should be particularly useful for automatic gas chromatography. For control purposes, counters with contacts operated after a number of pulses can be used.

Variation of the calibration constant K (eqn. (1)) over a 10:1 range is required to avoid scale multipliers other than powers of ten. It can be arranged in the electronic or mechanical parts of the integrator. The latter is simple, as it requires only the selection of gear wheels, but the counters must then be of a special type. To take advantage of the wide range of counters available, the electronic method has been adopted. Steps of 2:1 are obtained by adding dividers up to 8:1. The capacitor is variable in 0·5 μF steps from 4 μF to 8 μF, and the reference voltage varies from 25 to 30 volts to fill in between these steps. Soldered wire links effect the changes to capacitors and dividers, and a potentiometer sets the reference voltage.

5. Conclusion

The accuracy of integration that can be obtained for industrial purposes can be considerably improved by the use of an electronic integrator. For various ranges of current or voltage different methods offer advantages. A system that charges and discharges a capacitor, appropriate to a 10 mA maximum signal, has been described. Its temperature stability is similar to that of electro-mechanical integrators, and its long term stability and linearity over a wide range of currents are considerably better. By using an electro-mechanical counter selected from a commercial range to record the integral, variations in display and control facilities can be arranged easily.

6. Acknowledgments

The Directors of George Kent Ltd. are thanked for giving permission for the paper to be published.

7. References

1. G. B. Marson and I. C. Hutcheon, "Electronics in industrial, process control—Parts 1, 2 & 3", *Control*, **4**, Nos. 33–5, pp. 88–92, 102–5, 88–91, March, April and May 1961.
2. G. H. Royer, "A switching transistor d.c. to a.c. converter having output frequency proportional to d.c. input voltage" *Trans. Amer. Inst. Elect. Engrs*, **74**, p. 322, 1955.
3. H. L. B. Gould and D. H. Wenn, "Supermendur—a new rectangular loop magnetic material", *Electrical Engineering*, **76**, pp. 208–11, 1957.
4. R. V. Jarrett, "An analogue R-C integrator with a digital output", *Proc. Natl Electronics Conf.*, **16**, pp. 611–8, 1960.
5. G. M. Ettinger, "Some circuit applications of avalanche devices", *Proc. Instn Elect. Engrs*, **108B**, pp. 563–9, 1961.

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U.D.C. 621.317.757

Recommended Method of Expressing Electronic Measuring Instrument Characteristics

7. WAVE AND DISTORTION ANALYSERS†

Prepared by the Technical Committee of the Institution and based on a report compiled by M. B. Martin (Associate Member).

Introduction

This is the seventh set of recommendations in a series which has the objective of influencing the uniformity of presentation of information on the features, characteristics and performance of electronic measuring instruments and thus assisting in their comparative assessment and selection by potential users. The establishment of standards of performance is not an objective of these recommendations.

Wave analysers are marketed with a variety of titles, e.g. frequency analysers and frequency spectrometers, but in general, whatever the title, they normally behave as frequency selective voltmeters and usually operate on one of two principles: the heterodyne, or the selective filter.

There are two ways of using the selective filter; as a band-stop to suppress the fundamental or as a band-pass tuned to the components of the waveform being analysed. The band-stop type of instrument is usually termed a distortion factor meter as it can only be used to measure the total of all components other than the fundamental, therefore, it cannot be truly described as an analyser. For this reason, recommendations have not been specifically included for this type of instrument.

Table of Contents

FEATURE	METHOD OF EXPRESSION	REMARKS
	Part 1.—GENERAL DETAILS	
1.1 Power supply requirements	.. volts a.c./d.c. .. c/s and/or battery .. watts (voltage change .. %)	Maximum voltage and frequency variations for which the stated accuracies hold good must be given.
1.2 Temperature range	.. °C to .. °C	Maximum ambient temperature range for which the stated accuracies hold good with nominal power supplies.

† Approved by the Council for publication on 15th March 1962 (Report No. 25).

	FEATURE	METHOD OF EXPRESSION	REMARKS
1.3	**Overall dimensions**	Height .. in (.. cm) Width .. in (.. cm) Depth .. in (.. cm)	
1.4	**Weight**	.. lb (.. kg)	
1.5	**Construction and finish**		Where the construction conforms to a particular specification, the latter should be named.
1.6	**Valve and/or transistor complement**	Type numbers.	State if special selection is required.
1.7	**Accessories**		Details of any connectors and adaptors, etc.
1.8	**Ancillary equipment**		Where specific chart recorders, filters or other devices are recommended for use in conjunction with the instrument, their features should be expressed.
1.9	**Mode of operation**	State whether heterodyne, filter or other type of instrument.	

Part 2.—FREQUENCY

2.1	**Range**	.. c/s to .. c/s in .. bands. Give centre frequencies in the case of filter-type instruments.	The frequency range over which the stated accuracies hold good should be given. If the range can be extended by a special setting-up procedure or additional units, details of precautions and revised accuracies should be given.
2.2	**Calibration accuracy**	.. % ± .. c/s	This is the maximum error of the calibration of the main frequency control in relation to the input frequency, assuming the initial adjustment has been correctly carried out. The law and effective scale length of the control should be given.
2.3	**Re-setting accuracy**	.. %	This is the re-setting accuracy of the main control.
2.4	**Incremental adjustment**		The method used should be stated.
2.4.1	Range	± .. c/s or .. %	
2.4.2	Calibration accuracy	.. % or .. c/s	May be given as c/s per scale division.
2.5	**Selectivities and band-widths**	± .. c/s .. octave at 1 dB down.	The provision of response curves is recommended.
2.5.1	Variable band-width	From .. c/s at 1 dB down to .. c/s at 1 dB down.	State whether variation is continuous or stepped.

	FEATURE	METHOD OF EXPRESSION	REMARKS
2.6	**Sweep facilities**	YES/NO	State method.
2.6.1	Sweep speed	.. seconds per band; if log, .. octaves/second	
2.6.2	External drive torque	.. oz-in (..gm-cm), at input shaft.	If the drive shaft can be coupled to a pen or other recorder, this should be stated.
		Part 3.—INDICATOR	
3.1	**Type of indicator**	Meter or other form.	Give details affecting accuracy of reading, mirror scale, etc.
3.2	**Type of scale**	Linear, logarithmic, square law, etc. Length .. in (..cm)	Give for each range.
3.3	**Overload protection**		
3.3.1	Meter	YES/NO	Give type.
3.3.2	Circuit	YES/NO	Give type.
		Part 4.—VOLTAGE	
4.1	**Ranges**	.. μV to .. V	Give lowest scale reading that is within stated accuracy for each range.
4.2	**Parameter indicated**	Amplitude of signal components.	Peak, r.m.s. or mean. Give instrument rectifier law.
4.3	**Accuracy of calibration**	.. % of scale reading.	Maximum error to be given for each range.
4.3.1	Calibration reference	Internal or external.	Give type and method of setting.
4.4	**Drift**		
4.4.1	Short term	.. % change of indication without readjustment.	Maximum change of indication over any period of 10 minutes within a 7-hour period commencing 60 minutes after switching on. Input frequency, amplitude of signal, temperature and mains voltage assumed to be constant.
4.5	**Hum and noise**	Less than .. μV referred to input terminals at maximum gain.	State if the figure varies with band-width.
4.5.1	Residual modulation products	Less than .. μV referred to input.	State if the figure varies with band-width.

FEATURE		METHOD OF EXPRESSION	REMARKS
		Part 5.—INPUT	
5.1	**Input attenuators and/or controls**	.. dB in .. dB steps.	State if continuously variable control is provided and whether this can be switched out of circuit.
5.2	**Voltage limitations**	.. V	Give maximum voltage to earth and maximum voltage that can be applied between terminals.
5.3	**Impedances**	.. ohms	
5.3.1	Unbalanced	.. pF .. MΩ	
5.3.2	Balanced	.. pF .. MΩ	
5.3.3	Differential	.. pF .. MΩ	
		Part 6.—OUTPUT	
6.1	**Frequency**	Constant frequency of .. c/s and/or .. c/s—frequency dial reading. Dial frequency.	In general, an output at the frequency of the dial reading can only be obtained from the filter type of instrument.
6.2	**Level**	Open circuit voltage of .. V at f.s.d. at .. c/s.	Give output for each frequency available.
6.3	**Source impedance**	.. ohm	If source impedance varies for different output frequencies, give for each.

List of the Recommendations in this Series

1. "Amplitude-modulated or frequency-modulated signal generators", (January 1958).
2. "Cathode-ray oscilloscopes", (January 1959).
3. "Low frequency generators", (March 1960).
4. "Valve voltmeters", (April 1960).
5. "A.c. bridges", (August 1960).
6. "Stabilized power supplies", (February 1961).

The dates given are the issues of the *Journal* in which the particular recommendation appears; separate reprints may be obtained from the Institution, price 1s. 6d., post free.

U.D.C. 621.391.88: 621.394/5: 681.39

Telephone Circuit Evaluation for Data Transmission

By

K. L. SMITH, B.Sc.(Eng.),†

J. BOWEN‡

AND

L. A. JOYCE‡

Presented at the Symposium on "Data Transmission" in London on 3rd January 1962.

Summary: When using a telephone circuit for data transmission, the effect of certain characteristics on the transmitted data signals have to be evaluated. Because of the random nature of many of these disturbances, recourse is made to statistical methods based on extended tests using special measuring equipment and computer-based data reduction methods. From these, mathematical models of the various circuits can be formulated whereby the optimum arrangement of the transmitted data and the effectiveness of various error detection and correction systems, can be determined.

1. Introduction

With the extension of data processing systems to incorporate data transmission links, the systems designer is faced with a new type of problem, since he has to rely on a data path over which he has little control in either performance, specification or administration; a path which was not even designed for the use he intends. Yet the designer has a primary responsibility to provide the necessary data processing facilities with the specified reliability and availability and to guarantee a minimum performance for the whole system. For this he needs an exact knowledge of the probable performance of every systems component he uses, preferably expressed as a simple mathematical model.

An exact formulation for the telephone network is made difficult by the random nature of the disturbing phenomena and recourse has to be made to statistical methods, based on thorough and extensive testing of a wide range of circuits to obtain a sufficiently broad assessment of the network characteristics and hence determine the basic model. Subsequent testing can then provide particular parameter values for the particular model to be used for a particular system design.

The purpose of this paper is to describe some of the network characteristics which have to be evaluated, the approach adopted and field equipment used for data gathering and primary reduction, and some of the preliminary results obtained.

2. General Nature of Telephone Circuit Imperfections

The imperfections of telephone circuits for data transmission have little effect on speech communication for which the network was primarily designed and generally these imperfections represent a deliberate relaxation of specification to achieve a worthwhile economy and yet still maintain a satisfactory performance of speech transmission.

Furthermore the effects of these imperfections on data transmission systems vary with the design of the modulation equipment, and any test results should always be qualified by reference to the modem employed. For example, a modem employing phase modulation has a superior performance to one using frequency modulation in the presence of noise, yet the performance of the phase modem is affected far more by the presence of envelope delay distortion in the telephone circuit than is the frequency modem.

For data transmission the main concern is the data speeds which can be achieved over the circuit, and the error rate to be expected at those speeds. The error rates influence the choice of the data speed for the system.

The major imperfections of the telephone circuit which affect the error rate are:

(1) Bandwidth limitations
(2) Frequency shift
(3) Amplitude distortion
(4) Envelope delay distortion
(5) White noise
(6) Impulse noise
(7) Circuit interruptions.

Bandwidth limitations are imposed at the low-frequency end of the spectrum by the presence of in-channel 2VF signalling systems and at the high-frequency end by the cut-off of either the line plant or the terminals, the main source of difficulty being the existence of heavily-loaded tie cables. For this reason the G.P.O. in the United Kingdom restrict the bandwidth available for data transmission to the range 900–2100 c/s, together with a smaller band from 300–400 c/s used for error control and supervisory purposes in some systems.

† IBM United Kingdom Ltd., 101 Wigmore Street, London, W.1.
‡ IBM World Trade Laboratories, Hursley, Winchester, Hants.

The main effect of bandwidth restrictions is, of course, limitation of data speed because insufficient spectrum exists to carry the sidebands. For this reason the major emphasis in modem design has been directed towards systems requiring a minimum bandwidth and the most promising is a vestigial sideband phase modulation scheme. Unfortunately its use is restricted to circuits without frequency shift.

Frequency shifts are caused by lack of synchronism between the master oscillators of multi-channel carrier telephone terminals. Consequently in phase modulation systems it is necessary to transmit both sidebands from which a modified carrier can be derived at the receiver. Consequently the maximum data transmission speed which can be achieved is less than half of the intrinsic circuit capability. Frequency shift modulation systems are not affected seriously by frequency shifts, but they do not exploit the available spectrum as efficiently as phase modulation schemes. Although the first order sidebands are the same distance from the carriers as are those of the phase modulation system, the distance between the two carriers in frequency shift modulation represents an additional bandwidth requirement.

Amplitude distortion causes unequal attenuation of the various sidebands associated with the modulation of a carrier. At a given data transmission speed this affects the recovery of the data signal and, in some suppressed carrier systems, the recovery of the carrier signal itself. Furthermore, as the data transmission speed varies the range of sidebands produced suffer different attenuations, which appears as a variable receive level effect. By careful choice of carrier frequency the relative attenuation of the two sidebands at a given speed can be made approximately equal by positioning the carrier at the mid-frequency of a symmetrical loss/frequency characteristic. Sometimes a compromise amplitude equalizer needs to be added to the line circuit to shift the mid-point to a carrier frequency set by speed considerations. In modern phase modems using sideband multiplication to extract the carrier, sideband amplitude distortion is not a serious problem.

Envelope delay distortion on the other hand, which represents an unequal delay in arrival of the sidebands at the receiver, does hamper carrier reconstruction, although again a compromise in phase equalization can sometimes centre the characteristic on the carrier frequency.

White noise is seldom met on telephone circuits, and then it is usually caused by a poor contact in the circuit. It is uncommon to operate telephone circuits at such low levels that amplifier noise becomes observable. Performance in the presence of white noise is often used as a comparative measure of the performance of a modulation system, since the results can usually be compared with theoretically derived predictions as a check that the modem is indeed functioning correctly. Comparative white noise test results can also be correlated with performance under actual line noise conditions and are the most valuable test performance index to take for modems, since the disturbing noise can be rigorously specified.

Impulse noise is caused by disturbances associated with the automatic switching plant, whether by crosstalk from dial current trains in exchange lines, or by magnetic induction from the switch solenoids. The levels of these disturbances are often such as to cause a catastrophic loss of the line signal and so to introduce a burst of errors into the data. Their worst effect occurs at the receiving end of a circuit, i.e. at the receiving subscriber's exchange, where the signal level is quite low and the disturbance has its maximum effect.

Interruptions in a circuit can be caused by a variety of effects, but the most common is poor contacts in switches, relays and plugs. Such interruptions are usually intermittent and often last for only a few milliseconds, but like impulse noise, their effect on data transmission is disastrous, causing bursts of errors.

A further source of errors on public network trunk circuits is the timing signal, but its effects can usually be allowed for and it can probably be suppressed for a data call.

The first five of the imperfections listed above are relatively easy to measure and for a given circuit they are usually sufficiently stable that, once measured, their effect on the data transmission system can be evaluated and hence allowed for in determining the overall system performance.

On the other hand, the occurrences of impulse noise and circuit interruptions are completely random and hence considerable measurement of error rates on actual circuits is essential before the general statistical characteristics can be determined and expressed as a mathematical model. Furthermore, once the general model has been determined, it is still necessary to measure the actual circuit it is intended to use in order to determine the principal coefficients.

3. The Experimental Approach

In adopting a statistical approach to the problem of evaluating line characteristics, it is clearly desirable that the basic data should be gathered in a form which is suitable for reduction in a data processing system.

The International Consultative Committee for Telephones and Telegraphs have recommended the preservation of as much detail of the original characteristic information as possible. Fortunately this requirement is helped by the availability of data processing facilities, and, in fact, every bit of information transmitted, is recorded along with an indication whether it was received correctly or in error, or whether a circuit interruption occurred.

As certain line characteristics influence the performance of some modulation schemes more than others it is important to test the combination of telephone circuit and modem, so that a typical operating environment is created, and the results have direct application to system performance evaluation. These influences are further worsened by certain patterns, so it is necessary to provide facilities for repeated transmission of selected message patterns. To simulate actual traffic a pseudo-random number generator is most suitable, a block length of 511 bits being particularly convenient. By using pseudo-randomness it is easy to reconstruct a similar pattern at the receiver for checking incoming data bit by bit.

Such checking principles are essential for end-to-end testing, which is generally considered to be the only valid method. Loop testing eases the test equipment design problem but is difficult to apply without setting up special testing arrangements on specified circuits, which generally defeat the object of random testing. Direct looping back on a given circuit is difficult to achieve unless the four wire circuits are carried right back to the test points. Even if such facilities are available, the test results tend to give a pessimistic picture because the subscriber's exchange connection and "noise injection" area are each passed twice. In some exchange areas it is difficult to dial back into the same exchange, and, again, special testing arrangements have to be made.

Having decided on end-to-end testing, it is necessary to provide the following field testing equipment at each end of the circuit:

(a) *Transmission measuring set* to measure the overall insertion loss of the selected circuits in each direction, since most circuits will be four-wire. If suitable equipment is available the phase-shift at various frequencies should also be determined to correlate error rates with envelope delay distortion.

(b) *A modulator/demodulator of the type* to be used to provide a data transmission service. This should be lined-up and optimized for the particular circuit under test.

(c) *A pattern generator* capable of generating:

(i) a repeated pattern of 8 bits selected by the operator to determine pattern sensitivity of the particular line-modem combination;

(ii) a pseudo-random pattern of 511 bits to simulate actual traffic.

The pattern speed should be variable from 100 to 3000 bauds to cover most of the data speeds expected on the public network. Means should be provided for synchronizing the receiving pattern generator to the received line signal and to maintain this synchronism during a noise disturbance so that no apparent errors are added to the actual errors due to inadequacies of the testing equipment, and an accurate error picture is obtained.

A comparison circuit should also be included to compare continuously the received data with the locally generated data.

(d) *An error recorder* which can record each error as it occurs, suitably clocked, so that a detailed analysis is possible. Errors due to line interruptions should be distinguished from those

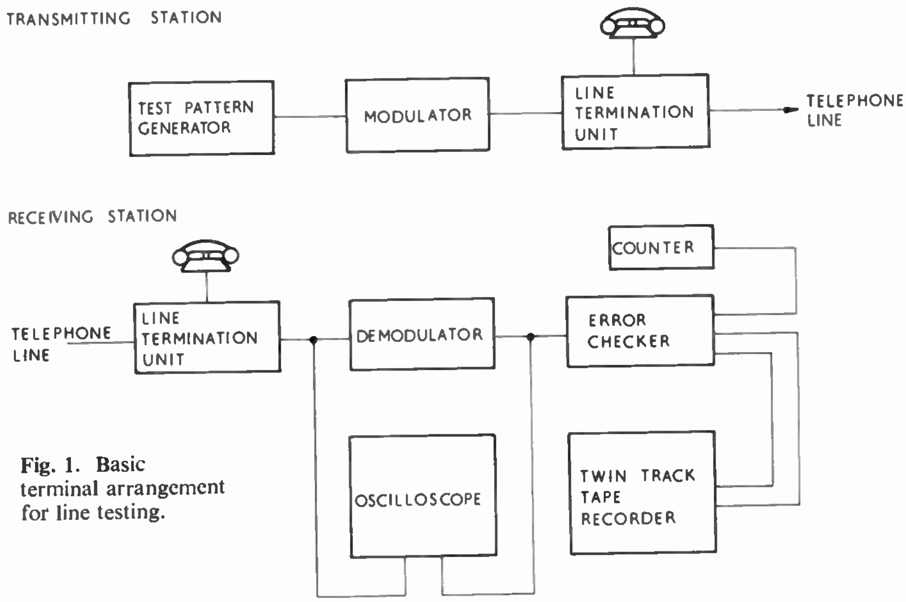

Fig. 1. Basic terminal arrangement for line testing.

caused by noise interference. The beginning of each block of information should also be recorded so that the error distribution within blocks can be determined.

(e) *An error counter* to determine the total errors occurring during the test as a check on the recorder. It gives also a quick indication of the average error rate. A second counter measuring the total number of blocks containing errors provides a ready assessment of the performance of systems incorporating error detection schemes.

Typical terminal arrangements are shown in Fig. 1.

The recorded magnetic tape provides a convenient input to a central data reduction calculator which translates the recorded data to discrete digital form, based on the number of error-free data bits between successive errors, the output being printed on a tabulator as an error-listing and also punched on cards.

These cards provide the input to a data analysis computer to obtain a statistical evaluation of the line error pattern characteristics and the efficiency of various error detection and correction schemes.

4. Line Testing Equipment

Much of the line testing equipment mentioned in the previous Section is conventional, but the pattern generator and error checker incorporate some unusual circuit features which are worth considering in some detail.

4.1. *The Pattern Generator*

The basic pattern generator consists of a nine-stage shift register which has two alternate feedback paths, as shown in Fig. 2.

For a repeated eight-bit pattern, the output from the eighth stage is connected to the input to the first. Once a pattern has been set on the register, therefore, it will continue to circulate this sequence of digits. The pattern is introduced by feeding pulses to the appropriate side of each stage via the character switches on the front panel.

For pseudo-random pattern generation, the outputs of the fifth and ninth stages are gated in an "exclusive OR" circuit which feeds the input to the first stage. This produces a repetitive 511-bit pattern containing all combinations of eight bits with the exception of eight zeros.

In order to verify that the incoming information is correct, the character generator in the receiver is run synchronously with the incoming data and an error is recorded when parity is absent.

When the synchronizing switch on the receiver is depressed, the receiver clock generator is synchronized in both frequency and phase. When the switch is released, the shift registers are reset to zero. At the same time, the feedback loop is disconnected and the input to the first stage of the shift-register is derived from the incoming information. The local feedback loop for either single character or pseudo-random pattern is restored by the appearance of a digit from

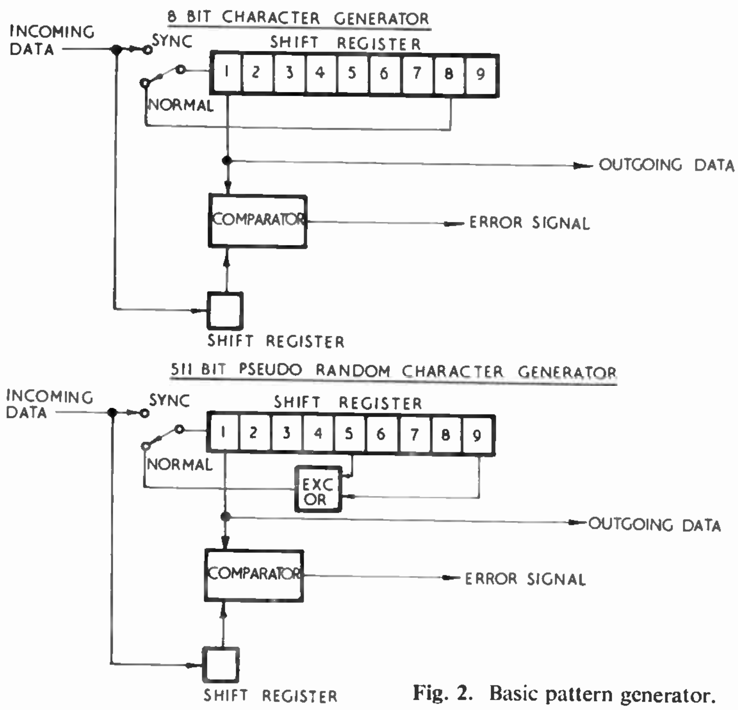

Fig. 2. Basic pattern generator.

the ninth stage of the shift register. This means that the register contains the nine proceeding bits of the incoming message so that when the feedback loop is connected the generator will continue to run synchronously with the incoming data.

4.2. *The Oscillator Circuits* (Fig. 3)

In order to compare the incoming data signal against locally generated information it is necessary to produce clock pulses having the same repetition rate as the transmitter clock and the correct phase relationship relative to the incoming data. These pulses are supplied to the local pattern generator which produces the same digital sequence as the transmitter. The incoming signal can then be compared sequentially with the output of this local generator and an error recorded when parity is absent.

The equipment has been designed to have a continuously variable bit rate from 30 pulses/second to 3000 pulses/second in two over-lapping ranges and can be set to an accuracy better than ± 2%.

The oscillator incorporates four divider circuits which give output signals corresponding to 0–5%, 0–30%, 0–50% and 70–100% of the duration of each oscillator cycle. Two of these signals, 0–30% and 70–100%, are supplied to two gate circuits so that one of these gates is open for the first 30% of each cycle and the other for the last 30%. The leading and trailing edges of the incoming data are differentiated to obtain a short duration pulse for each change of state of the incoming data. These pulses are supplied to the two gates mentioned above.

The local clock generator is assumed to be synchronized to the incoming data. If a subsequent output occurs from the gate, which is open from 0 to 30% of the cycle, then the local generator must be running fast and the output pulse is used to decrease the oscillator frequency. Similarly if an output occurs from the gate which is open from 70% to 100% then the local generator must be slow and the output pulse is used to increase the oscillator frequency.

In practice these two sets of output pulses are applied to the two inputs of a bi-stable circuit so that this circuit rests in one state when the generator is slow and in the other when the generator is fast.

Interference pulses occurring on the incoming data will cause incorrect information to be presented to this bi-stable circuit. To avoid this effect, the output is integrated over a period which is made sufficiently long that the oscillator responds only to the general trend of the incoming information.

Initially the transmitting and receiving generators can differ in frequency by as much as ± 4%. During normal operation the time-constant of the frequency control circuit has to be extremely long in order that the oscillator may remain in synchronism over line interruptions periods of up to 300 ms. This means that it would take an extremely long time for the frequency control circuit to impart a 4% change. To overcome this delay the time constant of the frequency control circuit is made variable and is changed to a short time constant when the synchronizing switch is depressed.

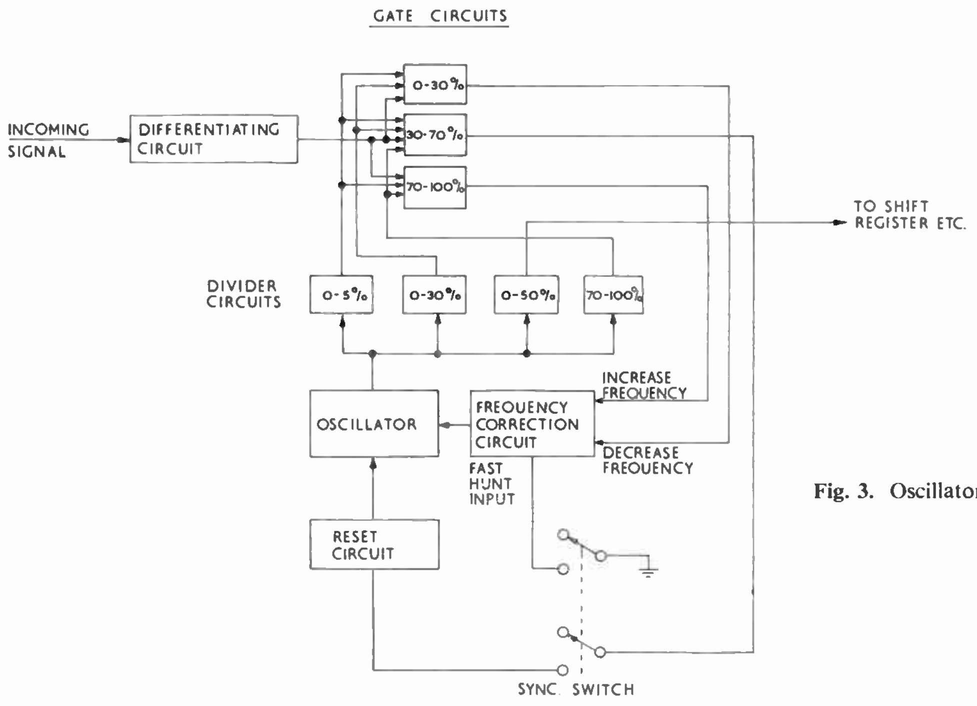

Fig. 3. Oscillator comparison circuit

In practice it is also necessary for an oscillator re-setting facility to be brought into operation when this switch is depressed since the incoming data will "slip" continuously relative to the local generator if the oscillator is a long way off frequency. This will result in an equal number of pulses appearing from both the 0 to 30% and 70 to 100% gates and the frequency control circuit will be unable to decide whether the oscillator is fast or slow.

In order to overcome this problem the differentiated data pulses are fed to a third gate which is arranged to be open from 30% to 70% of each oscillator cycle. An output occurs from this gate when the oscillator has drifted by more than 30% of a cycle and the pulse is used to re-synchronize the oscillator to the incoming data. This is performed by feeding the pulse to one side of a bi-stable circuit and using the output of this to "hold" the oscillator in the condition which normally occurs at the start of each cycle. The oscillator is released by feeding the next incoming data signal to the other input of the bi-stable circuit.

By this means a continuous cycle will recur whereby the incoming pulses drift relative to the local oscillator from 0% to 30% or 100% to 70%, according to whether the oscillator frequency is fast or slow, and are then reset to the 0 or 100% point, until such time as the oscillator is synchronized. The switch can then be released and the system restored to normal operation.

However, due to the short time-constant of the feedback circuit during this initial alignment, the oscillator frequency will tend to hunt rapidly to each side of the nominal frequency. If therefore the feedback were reduced instantly when the oscillator frequency happened to be at the extreme of its excursion then the correction would be insufficient to restore the frequency to that of the incoming signal without slipping occurring. It is necessary therefore to reduce the feedback time-constant gradually. This is achieved by making the time-constant continuously variable and deriving its control signal via an RC network to give a slowly decaying waveform.

The 0–5% divider circuit is used to provide a blanking pulse at the start of each cycle to the frequency control gates, and during the resetting cycle described above the oscillator is released in such a manner that the incoming waveform will coincide with the centre of this 5%. This reduces the chances of false information being presented to the frequency control circuit during the resetting operation.

The 50% divider provides both the timing pulses for sampling the incoming data and a suitable waveform for the tape recorder output circuits.

The basic oscillator is a conventional cross-coupled multivibrator, the frequency of which can be varied in two ways (Fig. 4). The manual adjustment takes the form of a helical potentiometer RV1 which controls the time-constant of the RC timing waveform and so determines the basic data test speed. In addition to this the frequency can be altered by varying the voltage to which the timing resistor is returned. This method is used to synchronize the oscillator when the equipment is in the receive data mode. In order to provide a suitable waveform for the divider circuits the mark-to-space ratio is made large so that the variable RC decay occupies most of the timing cycle.

The comparison of the local clock generator and the incoming data is determined by sampling the data edges in the periods 0–30% and 70–100% (Fig. 3). The gating waveforms which control the sampling gates are derived from the divider circuit shown on the right-hand side of Fig. 4. This circuit is designed to produce an approximately constant percentage pulse irrespective of the oscillator frequency.

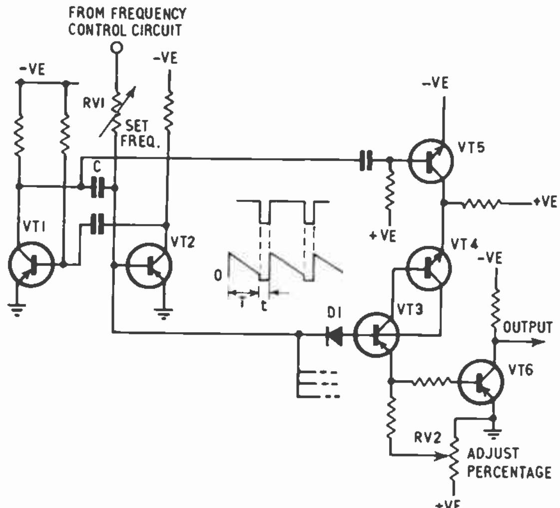

Fig. 4. Basic oscillator and divider circuits.

VT3 and VT4 form a complementary bi-stable pair which are initially reset to the state in which both transistors are cut off. The emitter of VT3 is returned to a variable voltage determined by the setting of RV2. The input to the circuit is derived from the base of VT2 and takes the form of an exponentially decaying waveform which is initially positive relative to the emitter of VT3 so that D1 is reverse biased.

As the input signal decays the voltage across D1 reduces until eventually the diode becomes forward biased, and a small current flows into the base of VT3 turning the transistor on. This in turn causes VT4 to conduct and the subsequent regeneration results in both transistors conducting fully. VT3 acts as an emitter follower so that its emitter potential is now approaching the negative supply voltage and D1 is again reverse biased. By this action only a very small current pulse is taken from the trigger source and the oscillator timing waveform is not unduly disturbed.

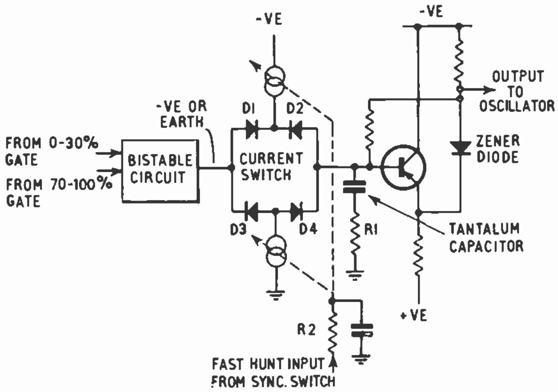

Fig. 5. Frequency control circuit.

Resetting is achieved at the end of each cycle by cutting off VT5 which effectively removes the negative supply to VT3 and VT4.

The outputs of the two sampling gates are fed to each side of a bi-stable circuit so that this rests in one state if the oscillator frequency is high and the other state if the oscillator frequency is low (Fig. 5).

The output from this bi-stable circuit is used to control the current switches consisting of diodes D1, D2, D3 and D4. When the input is at earth D1 conducts so that the current from the upper generator passes to earth and D2 is reverse biased. At the same time D3 is reverse biased so that current from the lower generator passes through D4 to discharge the capacitor; the action is reversed when the input is negative so that the capacitor charges.

For satisfactory operation the time-constant of the circuit has to be very long. Silicon semiconductor components are used in the current switch and generators and the output is fed into a compound emitter follower with d.c. feed back to increase the input resistance. The capacitor itself is a low leakage tantalum device.

The small resistor R1 which is placed in series with the capacitor produces a small voltage step in the output when the current switch changes state and has the effect of damping out any tendency for the oscillator to hunt about the true frequency.

4.3. *Magnetic Tape Recorder*

Four separate classes of information are recorded on a conventional two-track magnetic tape recorder using $\frac{1}{4}$-in. tape at a speed of $3\frac{3}{4}$ inches per second. The four signals are:—

(a) clock pulses to index each bit of data
(b) start of each 511-bit message
(c) received errors
(d) line breakdown.

Advantage is taken of certain exclusive relations between these four classes of signal to record them all on two tracks by a simple form of amplitude modulation, the actual waveforms using being shown in Fig. 6.

The tape recorder waveforms are designed to present a constant d.c. reference level to the tape recorder and so avoid certain adverse effects due to a.c. coupling within the recorder. Modulation of the waveform is effected by bridge modulators fed by balanced driver stages and controlled by the outputs from the pattern generator, the comparator circuit, and the modem carrier detector.

4.4. *Line Testing Procedure*

The telephone call is established in the normal way and the operators agree on the details of the test to be carried out.

An assessment of the speech quality and the background noise may be made and the steady state response of the line measured. The line is switched from the telephone instrument to the data transmission test equipment. A number of different eight-bit words are generated by the test pattern generator and the receiving operator measures the approximate bias distortion with the oscilloscope, the particular eight-bit words are chosen in an attempt to find the sequence showing the greatest amount of distortion.

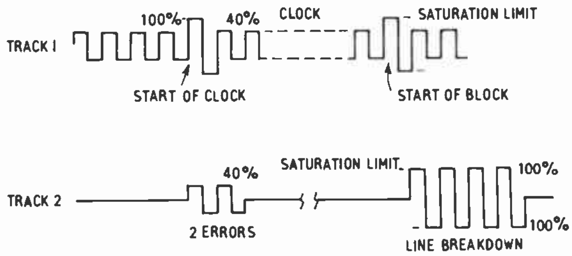

Fig. 6. Tape recorder waveforms.

The test pattern generator is then used to produce the pseudo-random sequence and is allowed to operate for the agreed test period. The receiving operator synchronizes his error checker on the incoming data, resets the error counter and starts the drive on the tape recorder. During the test which usually consists of transmitting at least one million bits the telephone line is monitored and a note made of any errors that occur as shown by the counter. Each error that occurs is indicated by the counter and recorded as a distinctive signal on the tape recorder.

5. Data Reduction and Analysis

5.1. *Data Reduction*

Data on the magnetic tapes prepared from all line tests is translated on a central calculator to discrete digital form and punched out on cards and listed on an accounting machine. The calculator used is a wired-program machine with 2080 characters of storage grouped into 416 words of 5 characters. For the translation procedure, the tape is read at $7\frac{1}{2}$ in. per second, i.e. at twice the recording speed, to reduce the time required.

Clock pulses corresponding to error-free bits are accumulated into a particular register until an error occurs, when they are routed into the next adjacent register. To simplify analysis by block, each register is split into two distinct words, recording the number of complete blocks (2 characters) and bits of a block (3 characters) respectively, the carry over from one to the other being determined by a start-of-block pulse.

Thus each register records the distance in error-free bits between adjacent errors. Should a register be filled before the next error occurs it is left at zero and the next register used. Errors occurring during a line breakdown are recorded as a special word so they are not included in subsequent analysis but rather indicate the duration of the failure.

When all registers are filled, the input tape is stopped and the data punched out on standard IBM cards, each recording 15 intervals together with a 3-digit test identification number and a 2-digit card number. A manually punched master card records details of the line, test speed, time of testing, location, etc.

A listing can also be printed for each magnetic tape showing:

(a) the number of blocks and bits between each error as punched on the card;

(b) the total number of blocks and bits transmitted from the start of the test until the error occurs;

(c) the number of bits between consecutive errors;

(d) the time of error occurrence;

(e) the number of errors in each minute;

(f) the total number of errors;

(g) the error rate expressed as errors per 1000 bits.

5.2. *Data Analysis*

With the data recorded on the cards pin-pointing the location of every error, an IBM 704 scientific computer can be programmed to extract a considerable amount of statistical information from this data.

So far there have been three main areas of interest:

(a) Line and modem performance evaluation, obtained from a direct analysis of the test results. Average error rates can be determined and tested for time sensitivity. Direct comparisons of particular modems can be made on different lines and at various data speeds.

(b) Distribution of errors within blocks. It is known that errors tend to occur in bursts, and we need to know the distribution of the lengths of these bursts and the distribution of errors within a burst, so that the performance of various data transmission systems over such lines can be determined and any weaknesses exposed.

(c) Coding simulation. We can simulate the behaviour of various error detection and error correction schemes, and, in particular, determine the undetected error rate. Such simulation gives a quick appreciation of probable performance without building test models.

The most valuable feature of this approach is that the lengthy, expensive and monotonous work of actual line testing is performed once only. As new studies are required, it is necessary only to write new programs for the computer; the original data can be used for all of the new investigations. Furthermore the same data can be used for any computer capable of accepting punched card input.†

6. Outline of Interim Results

Already a considerable number of lines have been tested in various European countries, including the United Kingdom by various IBM groups using similar testing equipment to that described in this paper. As manufacturers of data processing equipment the interest is not so much in the cause of errors, which is the responsibility of the telephone administration, as in the effect of such errors on the design and performance of data transmission systems, in particular the distribution of errors within message groups. It is known that errors occur in bursts, so there is certain error distribution behaviour. With the aid of the collated data the average form of the distribution can be determined and, what is more important, values can be assigned to the particular parameters and hence system performance can be determined. The general form of these results has already been presented[1] to C.C.I.T.T. Study Group A, but are repeated here to illustrate the type of result obtained in general. In the next Section the use of such characteristics to determine the performance of a data transmission system employing error detection by coding will be discussed.

6.1. *Distance between Errors*

The "error free" regions of the transmission medium are those of most interest, so perhaps the most important characteristic is the distribution of lengths of error free periods or "distance between errors". Figure 7 shows typical distribution of intervals between errors which exceed a value *d*. A short range correlation up to 15–30 bits is normally obtained, and this suggests a useful "guard space" of 20 bits, i.e. two errors separated by less than 20 correct bits are

† The work on data reduction and data analysis described in this Section is being undertaken by the IBM French Laboratory in Nice as part of the IBM World Trade Corporation Data Transmission Test Program.

considered to belong to one burst. Beyond this guard space the curve follows a Poisson distribution.

6.2. *Length of Bursts*

Using a guard space of 20 bits a curve such as Fig. 8 is obtained which shows the distribution of error bursts which exceed a given length *b*.

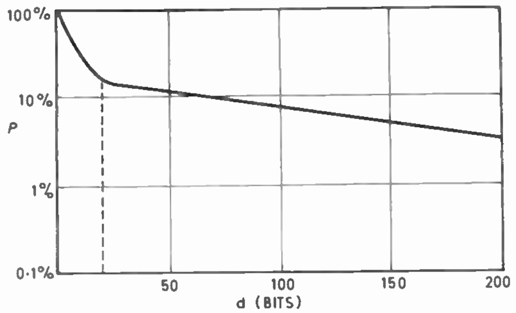

Fig. 7. Probability of intervals between errors exceeding *d* bits.

One interesting result is that most error bursts have a duration of 2 bits, suggesting disturbances of less than 1 millisecond. (These data refer to NRZ coding† in which errors always occur in pairs.) The longer bursts have an exponential distribution in length. The slope of the exponential portion of the characteristic depends on the guard space used. Within a burst the frequency of bit errors is just less than 0·5, indicating that the longer bursts correspond to strong line disturbances, which, as discussed above, are rather difficult to combat effectively by modem design, although the incidence of single isolated errors can be reduced by good design and choice of modulation method.

6.3. *Distance between Bursts*

Figure 7 showed that errors tend to group themselves into bursts and, using a guard space of 20 bits, Fig. 8 shows the distribution of error burst lengths. The distribution of the distance between bursts is shown in Fig. 9 and yet another interesting phenomenon emerges—that bursts of errors themselves tend to form bursts—and a suitable error free guard space between bursts of bursts is approximately 5 seconds.

This final "model" of the error pattern on telephone circuits produces analytical results which correspond very closely to the results obtained from actual line measurements (it can be related to the principal source of impulse noise on telephone circuits which are dialled pulse trains, each pulse itself causing a burst of errors, and the train of pulses (up to 10 in number) producing a burst of error bursts).

† NRZ coding: A coding method in which a binary digit is represented by a change (or no change) in the condition of a logical element rather than a particular condition.

This result provides the theoretical basis on which the capabilities of a telephone circuit for data transmission are judged. The method is to send a large number of known messages of fixed length and to obtain from these tests the percentage of messages in error. The justification of the method and its use is discussed later in the section on assessing system performance.

6.4. *Pattern Sensitivity of Messages*

Tests have also shown that the actual content of messages has an effect on the error distribution; and for this reason test messages must be designed to reveal such characteristics.

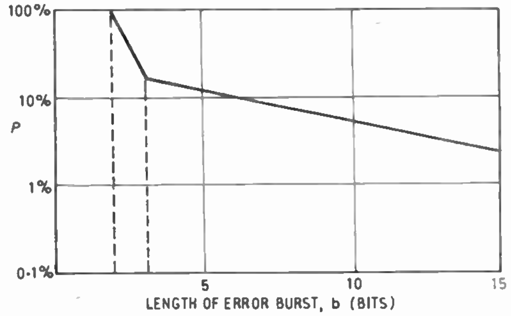

Fig. 8. Probability of error burst exceeding *b* bits.

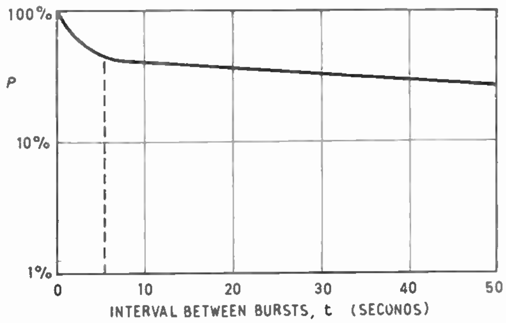

Fig. 9. Probability of intervals between bursts exceeding *t* seconds.

A particular example to illustrate this point is the excessive error rate which occurred in the last bit of a string of 0's when transmitted by a particular p.m. system. This was apparently due to envelope delay distortion phenomena.

6.5. *Time Sensitivity of Errors*

It has also been found that the error rate on both local and trunk calls depends upon the time of day, and can be directly related to the volume of telephone traffic at any time. The lunch hour between 1 and 2 provides a particularly quiet time and the user of data transmission equipment is well advised to send his data at this hour. Perhaps a tradition could be established to keep this time as the Data Hour.

6.6. *Probabilistic Model*

From these preliminary results a probabilistic model of a line with errors can be constructed.

It is apparent that the model must incorporate three basic states for the line, corresponding to:

State 1 low error;

State 2 in error burst;

State 3 between error bursts.

These are represented in Fig. 10.

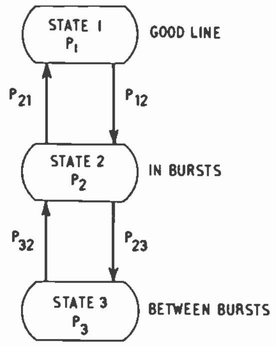

Fig. 10.
Probabilistic model.

Four parameters are then introduced such as P_{xy} which is the probability of the line changing from one State X to another State Y.

The model is completed by three parameters such as P_x which is the probability of an error occurring when the line is in State Y. Thus P_1 is the probability of a single error occurring in the low error condition, a typical value for P_1 being 10^{-5}, for $P_2 = 0{\cdot}5$, $P_3 = 10^{-5}$.

For each line values for the individual parameters can be determined and, by fitting them into such a probabilistic model, a good measure of the performance of error detecting and error correcting codes, etc., is obtained.

7. Prediction of Data Transmission System Performance

To illustrate the use of the simple mathematical model of the telephone network, let us consider the prediction of the performance of a data transmission system using error correction by retransmission.

The overall performance is the product of the two individual efficiencies:

(a) the intrinsic machine efficiency;

(b) the transmission medium efficiency.

7.1. *Machine Efficiency*

The data is arranged in blocks of length B, as shown in Fig. 11, with P check characters associated with each. A gap of length τ is provided between successive blocks for:

(a) start-stop time of the data source;

(b) envelope delay of the transmission path, go and return;

(c) turn-around time of any echo-suppressors included in the circuit.

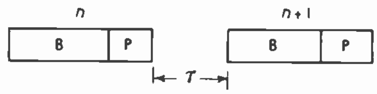

Fig. 11. Machine efficiency.

Let B = number of information characters
P = check characters associated with each message
b = character rate of line transmission
d = average distance between adjacent bursts of errors measured in characters
τ = turn-around time between messages

Then intrinsic machine efficiency $\eta_i = \dfrac{B}{B+P+\tau b}$

Probability of message encountering an error $P_i = e^{-B/d}$

Therefore actual machine efficiency $\eta_i = \dfrac{B}{B+P+\tau b} \cdot e^{-B/d}$

Then the intrinsic machine efficiency is readily derived as:

$$\eta_i = \frac{B}{B+P+\tau b} \qquad \ldots.(1)$$

and this is plotted in Fig. 12. It is the efficiency of the system through an error-free medium, and obviously improves as the block length is increased.

7.2. *Overall Efficiency*

Errors occur in bursts of bursts and follow a Poisson distribution of distance between adjacent bursts of bursts.

If the average distance is d characters, then the probability of a block of length B characters containing no error is

$$P_i = e^{-B/d} \qquad \ldots.(2)$$

which is plotted in Fig. 12.

Thus the overall machine efficiency over a transmission path producing errors is given by:

$$\eta_a = \frac{B}{B+P+\tau b} \cdot e^{-B/d} \qquad \ldots.(3)$$

This combined characteristic is superposed over the individual characteristics, and as might be expected, shows the existence of an optimum block length for maximum transmission efficiency, since the longer the block length, the greater is the chance of a line error occurring within it.

The optimum overall performance occurs when the intrinsic machine efficiency equals the line reliability, and corresponds to a block length

$$\hat{B} = \sqrt{d(P+\tau b)} \qquad \ldots.(4)$$

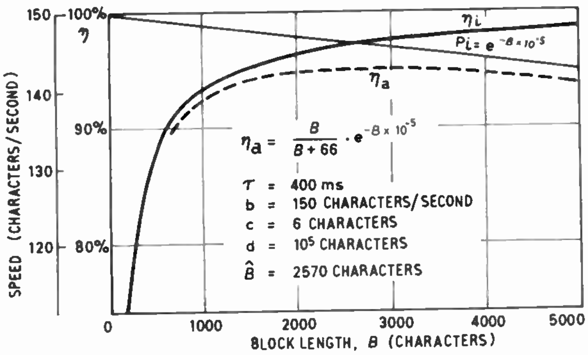

Fig. 12. Transmission efficiency.

and the corresponding optimum performance is

$$\hat{\eta}_a \simeq \left(1-\frac{B}{d}\right)^2 \qquad \ldots(5)$$

In practice a block length shorter than the theoretical optimum is normally used, since actual bursts tend to occur more frequently during bad periods than the average value suggests.

7.3. *Circuit Testing*

From Fig. 12 it is seen that the line error characteristic is approximately linear over the range of interest. This follows from the approximation of the exponential mathematical model

$$e^{-B/d} \simeq \left(1-\frac{B}{d}\right) \qquad \ldots(6)$$

and provides the basis of a simple method of assessing the probable performance of a given circuit.

By sending repeatedly a pseudo-random message of known length and determining the percentage of messages in error, one point on the straight-line characteristic is obtained, and hence the whole characteristic defined with sufficient accuracy.

8. Conclusions

From this brief description of the studies undertaken so far in Europe it can be seen that already there exists a sufficiently realistic model of the general network to enable a reliable estimate to be made of the probable performance of data transmission systems.

At the same time these and similar studies undertaken by other groups have shown the major weaknesses of the existing network for data transmission services and as the volume of such traffic increases in future so can improvements be expected to reduce these effects. In particular a reduction in noise level at exchanges and an extension of the available bandwidth by replacement of old, heavily-loaded cables need urgent attention by communication authorities.

Further studies are still necessary to broaden understanding of data transmission problems, but the results of such studies will be to increase the sophistication of the model presented in this paper rather than effect any drastic change in its basic structure.

The detailed recordings of the actual line characteristics have provided most valuable test data for new error detection and correction schemes, although it is not possible to predict the general future for such systems with certainty. To illustrate the problem, a particular circuit produced errors which involved the retransmission of 10% of 512 bit blocks. By introducing single-bit error correction, the number of blocks retransmitted was reduced to 5%, but only at the expense of adding 9 check bits to each 512 bit block, and a corresponding increase in transmission time of 2%. Thus the overall gain was 3% of transmission time, at the expense of a considerable increase in equipment. Economically it is probably better to restrict error correction schemes to essential control data, or to overcome difficult pattern sensitivities. Alternatively, the better use of error correction would be to use a simple scheme to improve a bad circuit to an average circuit and then rely on error detection and correction by retransmission which is most effective with low error rates.

Such studies are made more conveniently with the aid of the statistics collected and analysed by the methods described, and lead generally to a better understanding of the problems of data transmission.

9. References

1. IBM (World Trade Europe Corporation) Contribution No. 1 to Study Group A of CCITT (Data Transmission).
2. E. P. G. Wright, "Data collection and transmission", *Computer J.*, **4**, No. 2, pp. 103–7, July 1961.
3. A. C. Croisdale, "Error rates and error detection on telegraph circuits", *Trans. Inst. Radio Engs*, (*Communications Systems*), **CS-9**, No. 1, pp. 28–38, March 1961.
4. R. G. Enticknap, "Errors in data transmission systems" *Trans. I.R.E.*, **CS-9**, No. 1, pp. 15–20, March 1961.
5. T. A. Maguire and E. P. G. Wright, "The examination of error distributions for the evaluation of error detection and error-correction procedures", *Trans. I.R.E.*, **CS-9**, No. 2, pp. 101–6, July 1961.
6. E. P. G. Wright, "Error rates and error distributions on data transmitted over switched telephone connections", *Trans. I.R.E.*, **CS-9**, No. 1, pp. 12–15, March 1961.
7. M. B. Williams, "Present and future facilities for data transmission", *Computer J.*, **4**, No. 2, pp. 88–93, July 1961.
8. A. A. Alexander, R. M. Gryb and D. W. Nast, "Capabilities of the telephone network for data transmission", *Bell Syst. Tech. J.*, **39**, No. 3, pp. 431–76, May 1960.
9. E. N. Gilbert, "Capacity of a burst noise channel", *Bell Syst. Tech. J.*, **39**, pp. 1253–65, September 1960.
10. P. Mertz, "Model of impulsive noise for data transmission", *I.R.E. Con. Rec.*, **8**, Part 5, pp. 247–59, March 1960.
11. P. Mertz, "Model of error burst structure in data transmission", *Proc. Nat. Electronics Conf.*, **16**, pp. 232–7, October 1960.

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APPLICANTS FOR ELECTION AND TRANSFER

The Membership Committee at its meeting on 27th September 1962 recommended to the Council the election and transfer of 21 candidates to Corporate Membership of the Institution and the election and transfer of 39 candidates to Graduateship and Associateship. In accordance with Bye-Law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

CORPORATE MEMBERS

Direct Election to Associate Member

*DEHN, Rudolf, M.Sc. *Thurso, Caithness.*
HOLMES, Major Ronald. *Manchester.*
SOAMES, Spencer Thomas. *Middlesex.*

Transfer from Associate to Associate Member

GORHAM, Horace Henry James. *Watford, Hertfordshire.*
MANCHESTER, John Kay. *London, W.4.*
MARTIN, Alexander Duncan. *Lisburn, N. Ireland.*

Transfer from Graduate to Associate Member

AZAR, Yoram. *London, W.4.*
DANIEL, Michael. *Manchester.*
GABOR, Peter Reuben. *Tel-Aviv, Israel.*
GODDARD, Frederick Edward. *Chalfont St. Giles, Buckinghamshire.*
HARDING, Robert Harold. *Loughborough, Leicestershire.*
HUBBARD, Raymond Thomas. *Oranjemund, S.W. Africa.*
MARSHALL, Terence Alfred Anthony, B.Sc., (Eng.). *Accra, Ghana.*
WELLEN, Peter Antony. *Chatham, Kent.*

Transfer from Student to Associate Member

DIXEY, Graham Edward. *Ickenham, Middlesex.*
FORESTAL, Peter. *Hong Kong.*
GRAY, Reginald Edward. *London, N.1.*
GREAVES, Alan Bradley. *Guildford, Surrey.*
HARRIS, Herbert Arthur. *Johannesburg, S. Africa.*
PARK, Kenneth Charles. *Shepton Mallet, Somerset.*
TOWNSEND, Brian Joseph. *Chelmsford, Essex.*

NON-CORPORATE MEMBERS

Direct Election to Associate

HILL, Ernest Mammen. *Cheadle Hulme, Cheshire.*
STEENKAMP, Thomas, B.Sc. *Auckland, New Zealand.*
TAM KAI CHEONG. *Hong Kong.*

Direct Election to Graduate

ANDERSON, Peter Charles. *London, N.18.*
ARNOLD, Edwin Charles. *Southall, Middlesex.*
BALL, George Frederick. *Bristol.*
BIRD, Peter Frank. *Colchester, Essex.*
BROMLEY, Fred Arthur. *Bolton, Lancashire.*
BUTLER, Walter James. *Bristol.*
DENCH, Edward James. *Weymouth, Dorset.*
EVANS, Peter John. *Ilford, Essex.*
GANDERTON, Richard Alan. *Luton, Bedfordshire.*
GREGORY, Norman John. *Wells, Somerset.*
HARVEY, Alan John. *Stanmore, Middlesex.*
LOCK, Dennis Joseph. *Hemel Hempstead, Hertfordshire.*
MATTHEWS, Ronald Alan. *Hillingdon, Middlesex.*
PEGRAM, Terence William. *Chelmsford, Essex.*
PIGOTT, Peter Malcolm. *Orpington, Kent.*
PLEVIN, John. *Reading, Berkshire.*
PURNELL, Brian Thomas. *Greenford, Middlesex.*
ROWLINSON, Stanley Alfred. *Belfast, Northern Ireland.*
SHENINGTON-GUNN, John Dennis. *Basingstoke, Hampshire.*
SHERMAN, John Charles. *Portland, Dorset.*
SIMS, Brian John, B.Sc. *Fareham, Hampshire.*
SKINGLE, Gerald David. *Greenford, Middlesex.*
THORNBURY, Arthur Patrick. *Coatbridge, Lanarkshire.*
WILLS, Trevor George. *Farnborough, Hampshire.*
WILSON, Christopher Geoffrey Charles. *Axbridge, Somerset.*

Transfer from Student to Graduate

CHUNG, Yook Kei. *Hong Kong.*
GIBSON, George Arthur. *Aberdeen, Scotland.*
GOODAY, John Hamilton. *Chelmsford, Essex.*
HILL, Michael Edward. *Birmingham.*
HORWOOD, Peter John. *London, E.12.*
MARR, Robert Whyte. *Chadwell Heath, Essex.*
PEIL, Fred. *Ramsey, Huntingdonshire.*
SALTER, Martin Thomas Ardley. *Croydon, Surrey.*
SEABURG, Bernard Douglas. *Hornchurch, Essex.*
SKERRY, Christopher Scott. *Petersfield, Hampshire.*
WILSON, James Albert. *Slough, Buckinghamshire.*

STUDENTSHIP REGISTRATIONS

The following students were registered on the 27th September.

ABDUL-AHAD, Shamoon. *Banias, Syria.*
AFZAL, Muhammad, M.Sc. *London, S.W.6.*
AROOZOO, Edmund Percival. *Malacca, Malaya.*
AZAVEDO, Peter Thomas St. Anna. *London, S.E.4.*
BAKER, Arthur T. *Annan, Dumfriesshire.*
BEACHELL, Iain Alexander. *Leicester.*
BHADURI, Saurindra N. *Chelmsford, Essex.*
BHASKARAN, K. P. *Kerala State, India.*
BROWN, John. *Lerisick, Shetland Isles.*
BURKE, Brian. *Dublin, Ireland.*
CARTER, Reginald L. *London, N.1.*
HOON, Chuah Chin, B.Sc. *Kuala Lumpur.*
DALJIT SINGH, B.A. *Uxbridge, Middlesex.*
*DOOLEY, Russell J. *Hemel Hempstead, Hertfordshire.*
ENG TOON YEW. *Singapore.*
FARRELL, Anthony Henry. *London, N.5.*
FORRESTER, Roy. *Southampton, Hampshire.*
GILL, Brian. *Reigate, Surrey.*
GOAD, Peter. *Wembley Park, Middlesex.*
HIDER, Phillip Frank. *London, S.E.3.*
HOGLEY, Bernard, B.Sc. *Galashiels, Selkirkshire.*
IBRAHIM, Hag. M. *Khartoum, Sudan.*
JEFFERIES, Ian Donald. *Romford, Essex.*
JHA, Anil C., B.Sc., *London, W.10.*
JONES, Ronald Nelson. *Plymouth.*
KANITKAR, Sharadchandra V. *Poona, India.*
KEEN, Kenneth J. *Freetown, Sierra Leone.*
WAH, Lin Koon. *Hong Kong.*
LANGDOWN, Cecil W. G. *B.F.P.O. 151.*
LIM, Joseph Clement K. S. *London, N.W.11.*
McGREGOR, Sidney. *Basildon, Essex.*
MAHAJAN, Virendra, B.A. *Punjab, India.*
MARTIN, Helmet. *Dartmouth, Canada.*
MILLER, John R. *Arborfield, Berkshire.*
*NARGAS, Balbir Singh, B.Sc. *New Delhi.*
NARAYANA MURTHY, S. *Bangalore.*
NAZIR, Capt. Ahmad Malik, M.Sc. *Rawalpindi Pakistan.*
NGODDY, Charles I. *Lagos, Nigeria.*
NOROHNA, Oscar L. L. *London, S.E.4.*
RAVEN, Trevor George. *London, E.C.3.*
SMITH, John Gardiner. *Woodford Green, Essex.*
SMITH, Lawrence Hugh. *Rhondda, Glamorgan.*
STOKES, John Edward. *Swindon, Wiltshire.*
WOLFENDEN, John M. *Bolton, Lancashire.*
WRIGHT, John B., B.A. *London, N.2.*

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* Reinstatements

U.D.C. 534.75: 362.41

Auditory Perception and its Relation to Ultrasonic Blind Guidance Aids

By

L. KAY, B.Sc., Ph.D.†

Presented at the Symposium on "Practical Electronic Aids for the Handicapped" in London on 28th March 1962.

Summary: Experiments have been conducted to determine the optimum form of auditory presentation of signals received with ultrasonic echo-location guidance aids. Pulses of tone and continuous-wave frequency-modulated transmission were used, and there was no doubt that the latter form of transmission produced the most suitable auditory signals.

A portable guidance aid was made, using frequency-modulation echo-location principles, and the results justify tests with a small group of blind people.

1. Introduction

There have been several attempts to produce a guidance aid for the blind, based on the use of sound or electromagnetic waves, and covering a very wide range of frequencies. The sound waves ranged from audible frequencies to about 100 kc/s, and the electromagnetic waves have covered the radio, visible light and ultra-violet bands of the spectrum; but all systems were unsuccessful in the sense that they have not come into general use. The results did however lead to the conclusion that devices employing light would have most chance of success, and the sensory stimulus should be tactile. A brief survey of these devices by Bushor[1] reflects the current thought on the subject which apparently excludes almost completely the possibility of ultrasonic devices producing the required results in the future.

Very little work on the use of ultrasonics for blind guidance appears to have been reported since around 1950.[2] The work described here was started in the University of Birmingham in 1960, because recent progress in ultrasonic techniques and the studies of the bat's behaviour by Griffin and others[3] during the past decade led to the belief that a successful ultrasonic aid may now be possible.

The behaviour of bats under adverse conditions has shown that these particular animals can make good use of ultrasonic waves, and recordings show that much of the information needed and obviously obtained by the bat must be obtained as a consequence of the broad frequency band used in the ultrasonic emission. Whilst it is impossible at this stage to say just how the bat actually uses the information, there can be no doubt that adequate information is available in the signal it receives. Two theories have been suggested for the reason why bats use a wide frequency band, during the emission of the ultrasonic waves, which is in the form of a linear (or nearly linear) sweep of frequency of up to one octave in the region of 50 kc/s.

Strother[4] suggests the system is similar to that used in "chirp" radar. This is basically a pulse system using a frequency-swept transmission and a time delay in the receiving channel which is a function of the signal frequency being received. High frequencies are delayed longer than the low frequencies, so that a change from a high to a low frequency during the pulse transmission will result in an echo signal in the receiver output being compressed in time, and the amplitude relative to the noise background thereby increased. Two factors make this method unlikely; the resolution enjoyed by the bat shows that echoes from objects are received before the transmission has ceased, and there is no satisfactory evidence of a time delay which is a function of frequency in the auditory neural system.

The alternative system proposed by the author,[5] and at the same time and quite independently by Pye,[6] has more attractive possibilities. A beat note produced during the transmission between the transmitted frequency and that of an echo will have a pitch which is proportional to the distance of the object, and it is physiologically possible for such a beat note to be produced. Echoes received during the transmission can therefore be resolved in a frequency scale instead of a time scale. It will certainly be a very long time before either theory can be proved or disproved—as is evident from our inability to explain even the performance of our own auditory system.

It is nevertheless attractive to put forward the hypothesis that since a bat can "see" with ultrasonic waves, a blind person should also be able to do so, and therefore a copy of the bat's system of navigation may have a very good chance of success. This

† Electrical Engineering Department, University of Birmingham.

argument is not necessarily sound. Humans are of a very much more intelligent order and require more from life than a bat if they are to enjoy a satisfying existence. The bat probably concentrates on a single task whilst a blind person undoubtedly has many thoughts passing through his mind whilst mobile, and any really successful guidance aid should not prevent these. Many more such reasons can readily be found. The information obtained by a bat through the medium of sound waves may not in fact be adequate for the satisfactory guidance of blind humans, and it is very clear that psychological factors are of paramount importance.

On the other hand, the dependence on their natural senses for mobility causes considerable stress to blind people, and the relief of this may be of sufficient justification for the general use of a device. If the information gathered can be analysed subconsciously by the neural system, more than this relief of stress may ultimately be achieved. Audible sound waves already provide a medium whereby blind people can gather information about their surroundings without any conscious knowledge of the mechanism, and if this can be greatly improved by using ultrasound, this guidance medium should be fully exploited.

2. Laboratory Ultrasonic Guidance Aids

Both pulse and frequency-modulation systems tried in the past proved ineffective, and some evidence that a new system could be much more effective is required. Equipment was developed in the early stages of the research programme to reproduce some of the effects obtained with previous guidance aids,[7] using both pulses of tone and frequency-modulated transmission. Schematic diagrams of the equipment are shown in Figs. 1 and 2. Wide-band ultrasonic transducers of the type developed by Kuhl *et al.*[8] were used to cover the frequency range required. They were not available in the past and it is partly due to these that encouraging results have now been obtained.

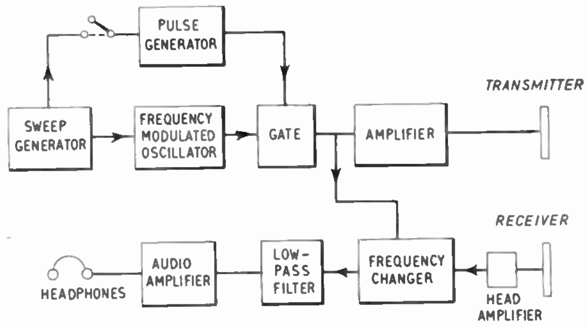

Fig. 2. Schematic diagram of frequency-modulation system.

2.1. *The Pulse System*

The frequency of the transmitter oscillator in the pulse equipment could be varied over a range of frequencies from 20 kc/s to 100 kc/s, and the drive to the output stage was gated to produce pulses of tone of varying duration from 1 to 20 ms. The repetition rate was adjusted to suit the maximum distance from which it was required to receive an echo. When, for example, a distance of 20 feet is required, the time taken to receive an echo from this distance is approximately 40 ms. The period between the pulses must then be adjusted to 80 ms to avoid ambiguity in range. When ambiguity arises it is because the ear is unable to tell whether the echo or the transmission is received first, and the time interval between transmission and echo cannot therefore exceed half the transmission period. The received echo was heterodyned with a second oscillator having a frequency which was between 1 and 4 kc/s lower than the frequency of transmission. An audible note could then be heard in the headphones when an echo was received. An arrangement was provided for coupling the transmitted pulse to the receiver at an attenuated level so that both the transmission and the echo could be heard. The audible output to the headphones was in the form of either a low frequency

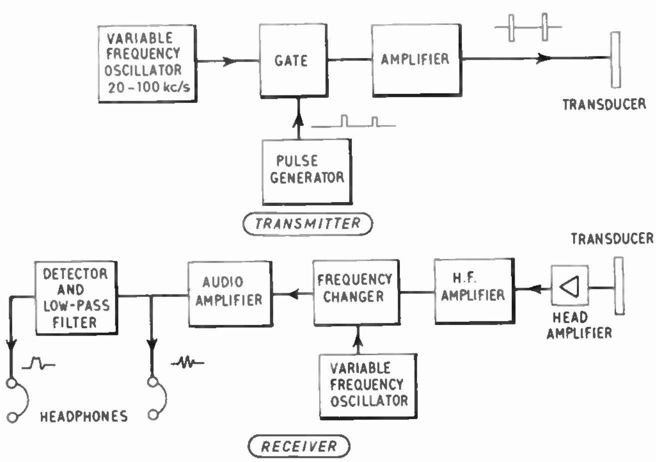

Fig. 1. Schematic diagram of pulse system.

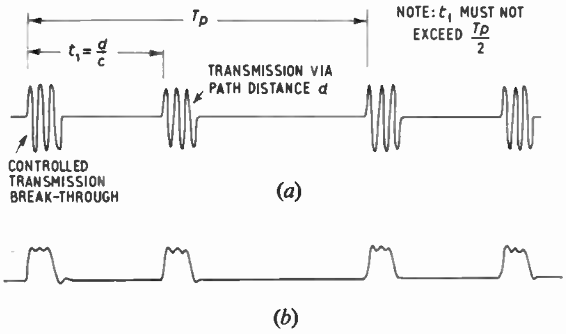

Fig. 3. Pulses applied to headphones.
(*a*) Audible frequency pulses applied to headphones.
(*b*) D.c. pulses applied to headphones.

pulse of between 1 and 4 kc/s, or a d.c. pulse obtained from the rectified audio frequency pulse as shown in Fig. 3.

2.2. *The Frequency-Modulation System*

The parameters of the system can best be understood from the graphs of Fig. 4; it will be observed that the transmission is continuous and that the sweep lasts for five times the period between the transmission pulses in the pulse system. The frequency sweep was from 60 kc/s to 30 kc/s and the maximum audible frequency was 3 kc/s. This maximum frequency was chosen because the response of the ear falls off beyond 3 kc/s; between 100 c/s and 3 kc/s it rises by approximately 40 dB, and is very effective in correcting for the attenuation in the air as the range increases. A wide frequency band of 30 kc/s in the medium has two advantages. Firstly, the time required to sweep this band is 400 ms for a maximum range of 20 feet, and an echo therefore becomes an almost constant tone; quality then becomes apparent. For example, an echo from 20 feet will last $(400-40) = 360$ ms. A short break of 40 ms maximum has little effect on the quality of the echo pitch. Secondly, the frequency band in the medium should be as large as possible if information about the reflecting surface is to be obtained. For instance, changing the wavelength by a factor of 2 : 1 produces a modulation on the echo waveform at the audio output when the reflecting surface has two or more reflecting points separated in range by about one wavelength. The pattern changes throughout the sweep period.

By adjusting the frequency sweep rate, shorter maximum ranges than 20 feet were also obtainable. The highest sweep rate gave a maximum range of 5 feet.

To reproduce the f.m. arrangements used in earlier guidance aids, a gate was included in the transmitter circuit to produce pulses of frequency-modulated signal, and the repetition rate was adjusted accordingly. One arrangement was to allow the transmitter to sweep a band of 3 kc/s over a period of 40 ms followed by an "off" period of 40 ms to eliminate ambiguity in range.

2.3. *Transducers*

Dielectric microphones of the type described by Kuhl were used, giving a uniform overall response up to 70 kc/s. The beam of the transmit–receive arrangement was about 10 deg at 50 kc/s but, of course, varied with the frequency. This beam of 10 deg was chosen initially to obtain high sensitivity and satisfactory angular resolution. A very narrow beam would be difficult to use and too wide a beam may have given too little sensitivity. Later developments show that wider beams can be used with some advantages.

2.4. *Laboratory Tests*

Tests were carried out with a few sighted colleagues to determine the approximate relative performance of the systems. The subjects were not trained in any way and could not therefore judge distance absolutely from the sounds they heard. They were only asked to state the order of merit in which they would place the systems on the basis of the information they received from the signals.

Initially, the transmitter was separated from the receiver and pointed at it from various distances. Only one "echo" was then received. The results are discussed briefly below.

2.4.1. Pulse system with no direct coupling to the transmitter

Pulses could be clearly heard but there was no indication of the distance from which the signal was being received. This applied equally to audible

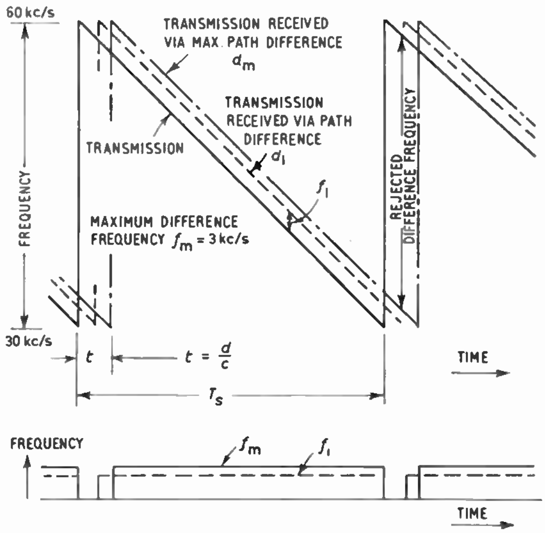

Fig. 4. Parameters of f.m. system.

frequency pulses and d.c. pulses in the headphones. Two headphones were used in all the tests to balance the auditory system, although this is not absolutely necessary.

2.4.2. Pulse system with direct coupling to the transmitter

(*a*) *Audible frequency pulses.* Both the transmitter breakthrough pulse and the echo pulse were adjusted to be of the same amplitude. The subject could tell the difference between the sounds produced by the transmission pulse alone and the transmission and "echo" pulses together, but could not say just how close they were. Thus he was unable to judge distance. There was a faint sensation of hearing a note with the click, the pitch of which varied with the distance to the transmitter. Once the attention is drawn to this, the pitch can be judged, but only with some difficulty. The effect is reduced as the relative amplitudes of the two pulses is changed. Altering the repetition rate of the pulses also alters the pitch.

(*b*) *D.c. pulses applied to the headphones.* The results were better than those obtained with audio-frequency pulses because the pitch of the note, heard together with the clicks, was much more marked. The pitch could be judged sufficiently well to give a rough indication of distance. Even so, the click was still predominant.

2.4.3. The frequency sweep system using a continuous transmission

This system is silent when no "echo" is present, but immediately an echo is received the echo tone can be heard. (In the pulse system the transmission click has to be tolerated at all times.) The "echo" was in the form of a clear but interrupted note; when the break in the note, due to the fly-back of the transmission sweep, was long—20 milliseconds say—there was a faint thump accompanying the tone which increased as the range was reduced. The frequency of the echo tone also reduced. With this system the tone pre-dominated. The pitch was easy to judge—as easy as a pure tone.

The pulsed f.m. arrangement gave results which lay between the pulse system and the continuous f.m. system.

3. The Spectrum of Signals Presented to the Ear

3.1. *Pulse System*

3.1.1. A.c. pulses

The signals presented to the ear during tests with the pulse system are shown in Fig. 3. These pulses were derived from two continuous sinusoids, one of which was amplitude modulated by short rectangular d.c. pulses of duration τ, having a repetition period T_p. The "echo" pulses were separated from the transmission break through—or marker-pulse by the time interval

$$t_1 = \frac{\text{distance between the transmitter and the receiver}}{\text{velocity of sound}}$$

The spectrum of either the marker pulses or the "echo" pulses taken separately can be expressed by the Fourier components of the amplitude modulated wave. Since the amplitude of the marker pulse and the "echo" pulse were arranged to be of equal amplitude, the periodic function of these two series of pulses can be expressed by

$$\mathrm{f}(t_M) = \mathrm{M}(t)\, V \cos \omega_a t \text{ for the marker pulses}$$

$$\mathrm{f}(t_E) = \mathrm{M}(t+t_1)\, V \cos \omega_a t \text{ for the echo pulses.}$$

$\mathrm{M}(t)$ for a series of d.c. pulses of duration τ and repetition period T_p is well known, and can easily be shown to be,

$$\mathrm{M}(t) = k + 2k \sum_{n=1}^{\infty} \frac{\sin nk\pi}{nk\pi} \cos n\omega_m t \qquad \ldots\ldots(1)$$

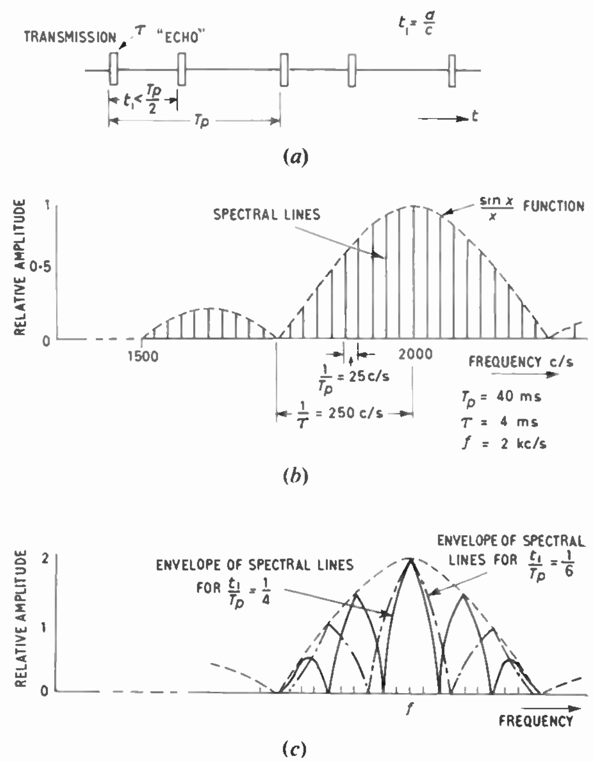

Fig. 5. Spectrum of audio frequency pulses from which distance information is to be obtained by the ear.

(*a*) Time scale of direct transmission break through and transmission via path distance *d*.

(*b*) Spectrum of audio pulses (transmission or "echo" alone).

(*c*) Spectrum of transmission and echo pulses together. *Note:* amplitudes equal.

where $k = \frac{\tau}{T_p}, \quad \omega_m = \frac{2\pi}{T_p},$

and $\frac{\omega_a}{2\pi}$ = audio frequency of pulse.

Hence

$$f(t_M) = kV \cos\omega_a t + $$

$$+kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} [\cos(\omega_a + n\omega_m)t + \cos(\omega_a - n\omega_m)t] \quad ......(2)$$

The spectrum of this periodic function is shown in Fig. 5(*b*). There will obviously be no difference between the spectrum for the echoes alone and the marker pulses alone, as there is no phase information contained in the spectrum. There is however, a phase difference between the spectral components of the two series of pulses due to the time interval t_1, as can be seen from

$$f(t+t_1) = kV \cos\omega_a t +$$

$$+2kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} \cos(n\omega_m t + n\phi) \cos\omega_a t \quad ......(3)$$

where $\phi = \frac{2\pi t_1}{T_p}$

Then

$$f(t+t_1) = kV \cos\omega_a t +$$

$$+kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} \cos[(\omega_a + n\omega_m)t + n\phi] +$$

$$+kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} \cos[(\omega_a - n\omega_m)t + n\phi] \quad ......(4)$$

When both the marker and the "echo" pulses are present in the audible output, the spectrum is the sum of the two spectra, from which is obtained

$$f(t)+f(t+t_1) = 2kV \cos\omega_a t + kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} \{\cos(\omega_a + n\omega_m)t + \cos[(\omega_a + n\omega_m)t + n\phi]\} +$$

$$+kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} \{\cos(\omega_a - \omega_m)t + \cos[(\omega_a - n\omega_m)t + n\phi]\} \quad ......(5)$$

The phase angle ϕ varies between 0 and π as t_1 is increased from 0 to $T_p/2$, the maximum time delay which can be used without introducing ambiguity in range. Thus when $t_1 = T_p/4$, $n\phi$ assumes values of $n\pi/2$, and zeros occur at intervals of $4\omega_m$ as can be seen from Fig. 5(*c*). When $t_1 = T_p/6$, zeros occur at intervals of $6\omega_m$. It will be seen that this spectrum now has a series of peaks (following the sin x/x envelope), and it is possible that these produce a similar effect to that of the formants of a signal with the fundamental removed. When the pulses were presented to the subjects, as described in the previous section, they had the sensation of hearing a low frequency signal, as well as the higher frequency "click". Because the pitch of this low frequency is rather indistinct, it was not possible to match it to a pure tone, but it was certainly in the region of $1/t_1$ the "fundamental" of the peaks of the spectrum. Non-linearity of the middle ear has not been considered in the analysis. This would produce two weak components of frequency, $1/t_1$, and $1/(T_p - t_1)$, as well as numerous other components due to intermodulation. It may well be that it is these the subject hears.

It should be noted that the peaks do not generally coincide with one of the spectral lines, i.e. when $mt_1 \neq T_p$ (m is an integer).

3.1.2. D.c. pulses

The spectrum of the d.c. pulses is easily seen, from the analysis of the a.c. pulses, to be

$$f(t)+f(t+t_1)$$

$$= 2k + 2kV \sum_1^\infty \frac{\sin nk\pi}{nk\pi} [\cos n\omega_m t + \cos(n\omega_m t + \phi)] \quad ......(6)$$

This is described graphically in Fig. 6, for values of $t_1 = T_p/4$ and $T_p/6$, where it will be seen that the peaks of the spectrum open out from ω_m/π, for $t_1 = T_p/2$, as t_1 is reduced. The subject was therefore presented with a strong component at the frequency he was expected to hear.

Whilst the analysis has assumed a d.c. pulse of duration τ, this pulse of pressure may not have been applied to the cochlea in this form, and strict agreement with the analysis can neither be expected nor measured. Nevertheless the difference between the a.c. and d.c. pulses is quite evident, and may explain the reason why the d.c. pulses produced a more pronounced effect from which distance could be gauged.

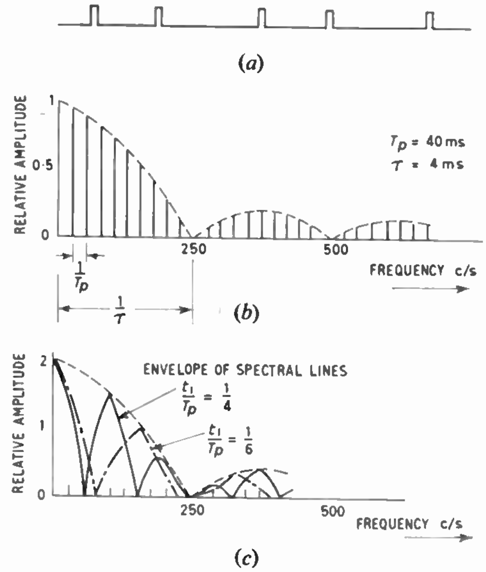

Fig. 6. Spectrum of d.c. pulses from which distance information is obtained by the ear.

(*a*) D.c. pulses applied to headphones.

(*b*) Spectrum of d.c. pulses (transmission or "echo" alone).

(*c*) Spectrum of d.c. pulses due to both transmission and "echo" of equal amplitude.

These are complex spectra and it would not be expected that the low frequency pitch be dominant; most of the energy is spread over a wide frequency band and is heard simply as a "click".

Results of a related nature have been obtained by other workers.[9–12]

3.2. *Frequency-modulation System*

The waveform of the "echo" using the f.m. system, is shown in Fig. 7(*a*). Each pulse of audio frequency is ideally (and very nearly in practice) locked to the start of the transmission, and is therefore repetitive. The spectrum can be found from the Fourier series of the waveform given by (for an even function)

$$f(t) = V \sum_{n=1}^{\infty} a_n \cos n\omega_s t \qquad \ldots\ldots(7)$$

where $\omega_s = \dfrac{2\pi}{T_s}$

(note $T_s = 5T_p$ for this system arrangement)

and $$a_n = \frac{1}{\pi}\int_{-\pi}^{\pi} f(t) \cos n\omega_s t \, dt$$

$$= \frac{2}{T_s}\int_{-\left(\frac{T_s-t_1}{2}\right)}^{+\left(\frac{T_s-t_1}{2}\right)} \cos\omega_a t \cos\omega_s t \, dt \qquad \ldots\ldots(8)$$

since $f(t) = \cos\omega_a t$ (the audio difference frequency) for

$$-\left(\frac{T_s-t_1}{2}\right) < t < \left(\frac{T_s-t_1}{2}\right)$$

and is equal to 0 for the remainder of the period T.

Let $$\frac{T_s-t_1}{2} = k'$$

then

$$a_n = k'\frac{\sin(\omega_a+n\omega_s)k'}{k'(\omega_a+n\omega_s)} + \frac{\sin(\omega_a-n\omega_s)k'}{k'(\omega_a-n\omega_s)} \qquad \ldots\ldots(9)$$

The spectrum is shown in Fig. 7(*c*). It should be observed that the spectral lines occur at intervals of $1/T_s$ and the peak of the sin x/x envelope only coincides with one of these when ω_a/ω_s is an integer. It will be observed from Fig. 7(*b*) that the spectrum is very narrow compared with that for the pulse system (of the order of 1 : 100), and varies slightly as the range is changed. The subject is presented with a signal having distinct tonal quality accompanied by a "thump" resulting from the break in the signal of duration t_1. Since $t_1 < T_s/10$ in this system arrangement, the thump is quite soft in character as compared with the harsh click of the pulse system.

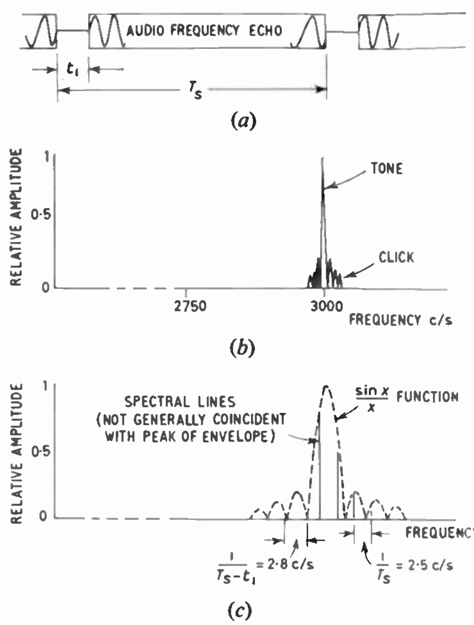

Fig. 7. Spectrum of echo from f.m. system (position of spectrum on frequency scale gives distance information).

(*a*) Amplitude modulation on echo with f.m. system.

(*b*) Spectrum of echo with f.m. system (same scale as for pulses).

(*c*) Spectrum on expanded scale.

4. Multiple Echoes

The systems were then used for detecting objects and, provided the background noise was negligible, single objects could be detected equally well with both pulse and f.m. arrangements. It is only when the signal/noise ratio is small that any difference is noticed and this affects both the maximum range of the system and the smallest object which can be detected. As was expected, the f.m. system was superior since semi-continuous signals were being used which always give better performance than pulsed signals, because of the difference in bandwidth and mean power.

When more than one object was in the field of view, the results of the various systems were very different. Under realistic conditions two echoes are rarely of the same amplitude since they are usually from different surfaces, and when this occurs the subject has great difficulty in hearing the weaker echo with the pulse system. The strong echo apparently suppresses it. This was not the case with the f.m. system except when two echoes were of almost the same frequency. When the strong echo is only slightly modulated by the weak one, discrimination becomes difficult. Generally however two notes are distinctly heard and the relative distance between the two objects and the distance to them can be easily judged.

It will be appreciated that the subject was being asked to interpret a very complex spectrum when more than one echo was being received with the pulse system, as can be imagined from the spectrum of Fig. 6. With the f.m. system the spectrum is very much simpler. The spectrum of two echoes does not seriously overlap until the frequency of the echoes are separated by less than $4/T_s$. In the system used, this was only 10 c/s.

Several objects close together, however, could not be resolved with the f.m. system—three is about the maximum—and the echoes merge into a musical sound, or a sound pattern. The ability to learn a sound pattern is well known from our every day experience, but it does not follow that these particular patterns can be learned.

The conclusions drawn from the laboratory tests were very definite. There was no doubt that the wideband continuous f.m. system was very much superior to all the pulsed arrangements, and was therefore the system worth trying as a portable device. As a go/no-go guidance aid the pulse arrangements could suffice, but judging from experiments in the past, more was required. Since the continuous wideband f.m. system gave results which were so much better than what must have been obtained from previous guidance aids, there was good reason for making a portable model.

5. Portable F.M. Ultrasonic Guidance Aid

A portable guidance aid has been made using a frequency band of 30 kc/s in the air, sweeping from 60 kc/s to 30 kc/s, and with a maximum audible frequency of 3 kc/s. Two sweep rates are provided to give 10 feet or 30 feet maximum range, and the duration of the echo tone is approximately 180 ms with a maximum break of 20 ms, or 540 ms with a maximum break of 60 ms respectively. Hearing aid receivers (earpieces) are used for producing the audible sounds, and instead of placing these over the ears, as is normally done with the hearing aids, a small diameter plastic tube is used to feed the sound into the outer ear (meatus). It is envisaged that a plastic mould will be used which supports the tube in the meatus but does not seriously affect normal hearing. The two ears are used because it is believed that auditory balance will be maintained and normal binaural auditory perception unimpaired. One ear can, of course, be used if found preferable. The "torch" has an effective beam of about 10 deg.

The purpose of making a portable aid with a "torch" was to determine, mainly, the nature of the auditory signals which would be received from the common objects which are encountered in everyday life, and to test the unit for its response to especially dangerous obstacles such as descending steps and step-down curbs. Manholes etc. come into the same category. It is also necessary to see if the type of signal presented by the aid could be interpreted by blind persons. Quite clearly, if this was not possible there would be little point in proceeding further. *It is emphasized here that the use of a torch is not envisaged as the final form of the aid; a binaural presentation is the ultimate aim*, but even a torch has already been found to have useful features.

6. Results Obtained with the Guidance Aids

Once the aid was made portable the potentialities became obvious. The results are best described in terms of the object and obstacles encountered, but it is impossible to describe the sounds as these are characteristic of the system alone. A sketch of the frequency spectrum does, however, give some idea of the character of the audio signals when these are complex.

(1) *Simple objects*

Smooth wall. This can be detected up to 30 feet provided the torch is pointed directly at the wall. Rotation of the torch over an angle of $\pm$ 5 deg produces a rapid change in the amplitude of the almost pure tone echo. Since the signal intensity is greater than can be obtained from any other object there is no difficulty in deciding it is a smooth wall and the direction in which it runs.

Post. The distance at which a post can be detected depends upon its diameter, but a 2-inch diameter road signpost can be located at 10 feet provided the torch is pointing horizontally. Again, rotation of the torch produces a rapid change in the signal intensity, but because the distance (as determined by the frequency of the signal) is less than for a wall echo of the same intensity of signal, the object is quickly classified. A sharp corner produces a similar effect but with much lower signal level. Since a 1 millimetre diameter wire can be detected at 4 feet, most sharp corners can be detected before a person collides with one.

Corner of room, etc. A strong signal is received which is of the same order as that from a wall, but on rotating the torch the sound changes in a different manner because of the different geometrical shape. Some practice is required in order to notice the difference.

Simple objects such as those described produce signals which are nearly pure tones, and many such examples can quickly be brought to mind.

(2) *Complex objects having a well defined character in the signal pattern*

Ascending steps. Many tones in an ascending scale are heard as the torch is directed up the steps. The sound is musical and each step can be counted as one note after another is heard to start. This can be understood better from Fig. 8(*a*).

Railings. These give a similar sound pattern, but of course from a different plane, and can therefore only be a succession of upright posts.

Bush. Each leaf or small branch produces its own weak signal, which when added to all the others, each of a different frequency, produces noise in a limited band of frequencies. This band depends upon the size of the bush as seen in Fig. 8(*b*).

Gravel path. Each stone produces its own signal as for the leaves of a bush, but the frequency band is much greater—depending upon the angle of the beam. Grass has a similar effect (Fig. 8(*c*)). When compared with a smooth path or road, the signal is very strong, but even concrete paving gives a detectable return, as seen from Fig. 8(*d*).

Descending steps. These are detected not by the presence of a signal, but its absence. The background from the pathway (paving, etc.) suddenly ceases at about 6 ft distance if the torch is pointed slightly towards the ground. The subject then realizes the break in the surface and proceeds cautiously to check if it is a hole or steps. If a hole, there is an echo from the far wall, indicating the width of the hole. Descending steps can be counted for a distance of three or four steps by noting the change in signal as the torch is directed down the steps at an almost vertical angle. Difficulty is experienced when the floor is smooth, since the background return is weak, but this the subject knows about and can take extra care if steps are likely to be encountered.

Pedestrians are observed by the rapid change in the frequency of their echo signals as they approach, or pass ahead.

Objects fluttering in the breeze are observed by the wavering of the echo frequency; tall blades of grass are easily heard.

The performance of the device is unimpaired in a 60 miles/hr gale and no serious fading of signals has been observed. This is partly accounted for by the fact that few echoes are of absolutely constant amplitude because of the wide frequency sweep in the transmission medium. Variation in amplitude due to atmospheric conditions is therefore unnoticed. Even heavy rain seems to have no effect.

There are many more complex objects than those given, and it may now be appreciated that an infinite variety of signals is possible. We all have learned many sound patterns—speech is an excellent example —and it is not unlikely a blind person could learn the sound patterns from this system, but just how well has yet to be determined.

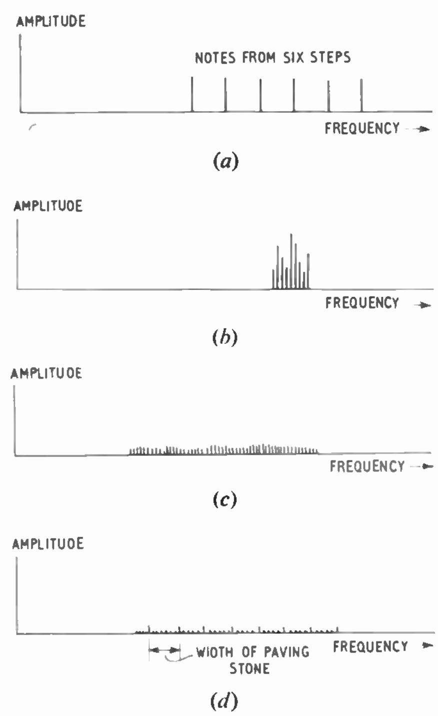

Fig. 8. Examples of sound spectra from various objects.

(*a*) Sound spectrum from ascending steps.
(*b*) Sound spectrum from a small bush.
(*c*) Sound spectrum from gravel or grass.
(*d*) Sound spectrum from paving.

Only a few blind persons have tried the guidance aid at the time of writing and these for only short periods of time and without training. Their reactions were sufficiently encouraging, however, for an order to be placed with a manufacturer for ten pocket units so that a controlled evaluation could be made over a period of time. One unit, tried by untrained persons, is not likely to provide the required information. The fact that ten units have been ordered does not, however, mean that the aid is already considered to be a success.

7. Conclusion

It has been shown that the most efficient guidance system, using ultrasonic transmission, is in the form of a frequency-modulated wave similar to that used by bats, who rely entirely upon ultrasonic waves for their orientation. Information can be gathered about one's surroundings in the form of sound patterns—some simple. and others complex—and it is possible a blind person can learn to interpret these. It is essential that the normal auditory cues obtained by a blind person are not destroyed, but they could be suppressed, if desired, when superior information which was easily learned was available. Tests are to be carried out on a small group, each with his own aid, over a period of time, to determine how effective and acceptable the device can be.

8. Acknowledgments

The author is pleased to acknowledge the financial assistance given by St. Dunstan's and the encouragement received from St. Dunstan's Scientific Committee and the National Research Development Corporation.

9. References

1. W. E. Bushor, "Medical electronics, part IV: prosthetics, hearing aids and blind guidance devices", *Electronics*, **34**, No. 25, p. 43, 23rd June 1961.
2. R. L. Beurle, "Electronic guiding aids for blind people", *Electronic Engineering*, **23**, p. 2, 1951.
3. D. R. Griffin, "Listening in the Dark" (Yale University Press, 1958).
4. G. K. Strother, "Note on the possible use of ultrasonic pulse compression by bats", *J. Acoust. Soc. Amer.*, **33**, p. 696, 1961.
5. L. Kay, "A plausible explanation of the bat's echo-location acuity", *Animal Behaviour*, **10**, Nos. 1–2, p. 34, January–April 1962. (Received by Editor, May 1960.)
6. J. D. Pye, "A theory of echo location by bats", *J. Laryngol. Otol.*, **74**, p. 718, 1960.
7. "Research on Guidance Devices for the Blind" (Haskin Laboratories, New York, 1946).
8. W. Kuhl, G. R. Schodder and F. K. Schroder, "Condenser transmitters and microphones with solid dielectric for airborne ultrasonics", *Acustica*, **4**, p. 519, 1954.
9. W. R. Thurlow and W. A. Small, "Pitch perception for certain periodic auditory stimuli", *J. Acoust. Soc. Amer.* **27**, p. 132, 1955.
10. J. L. Flanagan and N. Guttman, "On the pitch of periodic pulses", *J. Acoust. Soc. Amer.*, **32**, p. 1308, 1960.
11. J. L. Flanagan and N. Guttman, "Pitch of periodic pulses without fundamental component", *J. Acoust. Soc. Amer.*, **32**, p. 1319, 1960.
12. R. A. Jenkins, "Perception of pitch, tembre, and loudness", *J. Acoust. Soc. Amer.*, **33**, p. 1550, 1961.

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North Western Section Report

At the conclusion of the 1961–62 session a party of members of the Section visited the Mullard valve factory at Blackburn. This is a completely self-contained plant which includes sections concerned with fine wire and glass manufacture.

In the production of tungsten wire—some of it eventually to be drawn to a diameter of 0·005 mm (a tenth of the thickness of a human hair)—the basic raw material is scheelite ore which is first ground to a fine powder. Subsequent stages are reduction, compression and sintering, swaging into rods, and finally several stages of drawing through diamond dies. The glass factory processes are essentially continuous and lengths of about four feet are fed to semi-automatic machines which cut it to the appropriate length for the valves.

Members then saw the valve manufacturing plant where the electrode assemblies are built up by operators on pins set in the bases. (An experimental automatic assembly machine for this purpose was also seen.) The sealing and pumping of the valve is effected on rotary machines holding 30 valves at a time. H.f. heating fires the "getter" and the cathode heater is "formed" by passing a current. Extensive individual testing and quality control random testing follows.

Members of the North Western Section were the guests of A.B.C. Television Limited at the Didsbury studios in Manchester on the evening of 3rd September last. Approximately 160 people were in the studio circle when Mr. B. R. Greenhead, general manager of A.B.C. Television, opened the proceedings. In his opening address of welcome, Mr. Greenhead made special reference to the continuous activity for over 30 years of the Institution's North Western Section. He then introduced Mr. J. S. Sansom who read a paper on the SECAM System of Colour Television. (A resumé of Mr. Sansom's paper is given below.)

After the presentation of the main part of Mr. Sansom's paper, a colour film was transmitted from the A.B.C. Teddington Studios by Post Office land lines to Didsbury and was projected on to a large screen. The picture was simultaneously shown on two colour receivers and on two monochrome receivers which acted as monitors. Excellent results were obtained over this long link of some 200 miles.

The Chairman of the North Western Section, Mr. F. J. G. Porter, was invited to close the proceedings and thanked Mr. Greenhead and his Company for making the evening possible, Mr. Sansom for his most interesting paper and all those who had been concerned with the admirable arrangements.

Colour Television—A Review of the SECAM System

Mr. Sansom opened his remarks by stating that the fundamental requirements of a public service colour television system were outlined by the American National Television System Committee (N.T.S.C.) over twelve years ago and were exemplified by the system they devised. While the French SECAM system uses identical coding practice in the formation of luminance and colour difference signals, significant divergence occurs in the method of modulating the sub-carrier with the latter signals.

A simplification is achieved by avoiding the need to convey both signals simultaneously; the arrangement adopted is one of alternate lines of red and blue colour difference signals frequency modulating the sub-carrier. Reconstitution in the receiver involves the use of a delay line of one line duration and implies that any two consecutive displayed lines are modulated with the same colour information; the inherent reduction of vertical resolution, namely approximately 2 : 1, is not observed by the eye due to the same physiological phenomena which permit the reduction of chrominance bandwidth by a factor of at least 3 : 1.

The undeviated sub-carrier frequency is chosen to be nominally the same as N.T.S.C. for a given scanning system but the use of frequency modulation obviates the effectiveness of dot interlace in reducing its visibility on a monochrome picture. Satisfactory compatibility is provided by phase switching techniques in the encoder and employing a low level of sub-carrier. The use of frequency modulation implies a superior noise protection performance than on a.m. systems and this is further enhanced by l.f. and h.f. pre-emphasis techniques.

The function of the reference signal is solely to control the line frequency switch in the receiver, correction being available at field rate and provided by signals gated into the post field sync suppression period. These signals may be suitably derived to provide running checks of performance in a similar manner to vertical interval test signals.

Cross-colour is greatly reduced by a technique whereby the sub-carrier amplitude is momentarily increased in the presence of luminance transitions, reducing their visibility as a spurious colour signal. No significant worsening of compatibility is observed.

It is claimed that use of the above mentioned techniques facilitates the design of simple receivers which are not critical in adjustment. In a similar way the performance of transmitters and links and studio equipment is not critical and not liable to produce hue and saturation errors.

F. J. G. P.

U.D.C. 621.397.621

Tuner Design Considerations for a Combined 405/625 Line Television Receiver

By

F. C. COX†

Presented at a Television Group Symposium on "Dual-Standard Television Receivers" in London on 7th February 1962.

Summary: U.h.f. tuner design is comparatively standardized except for means of meeting the probable requirement of 60 dB image signal rejection. The system of feeding the u.h.f. tuner into the v.h.f. stage by directly coupling to the mixer grid appears to be the logical approach, bearing in mind both technical and commercial considerations. The general design requirements to meet the use of u.h.f. in combined receivers are comparatively straightforward. The only outstanding problem is that of the eventual v.h.f. system and a scheme for meeting the likely changeover to C.C.I.R. channel allocations is put forward.

1. Introduction

Design considerations for tuners for a combined 625/405 line receiver can be grouped under three headings:

(1) The u.h.f. tuner.

(2) The system of coupling the u.h.f. tuner into the v.h.f. tuner or the i.f. amplifier of the receiver.

(3) The additional requirements that the system places on the v.h.f. tuner.

These problems must be considered in the light of the system that will most likely be adopted in this country. In the case of (1) and (2), these are relatively independent of the system. The requirements for the v.h.f. tuner are, however, extremely dependent upon the system to be used and must be considered for each of the proposed systems; these will be dealt with later when considering this section of the design.

2. The U.H.F. Tuner

The design of tuners has now been stabilized into two groups: one utilizes an r.f. amplifying stage and the other does not. (In the latter group should be included v.h.f. tuners which use the harmonic of the v.h.f. oscillator and a crystal mixer, mounted on an alternative coil biscuit, to receive a u.h.f. signal over a limited band.)

These two groups can be related, in the first case to the Continent and, in the second case, to America; if the present legislation going through Congress is introduced then it can be anticipated that America will tend to change over to the somewhat better tuner designs of the Continent. The use in America of tuners without an r.f. amplifier has been due to the local nature of the u.h.f. transmissions, whereas on the Continent the u.h.f. bands have been used to provide national coverage on one or more networks.

In the case of tuners with an r.f. amplifier, designs have become so stabilized that, irrespective of manufacture and ignoring slight physical differences, the circuit and performance are almost identical; all tuners are based on a three-gang air-spaced capacitor which tunes lecher lines. The only notable exception to this practice employs continuously tuned lecher lines short-circuited by contacts.

It is not proposed in this paper to enter into the design data for the tuner but to look at the specific requirements of this country. A typical u.h.f. tuner is shown in Fig. 1 and Fig. 2 gives its circuit. Apart from the

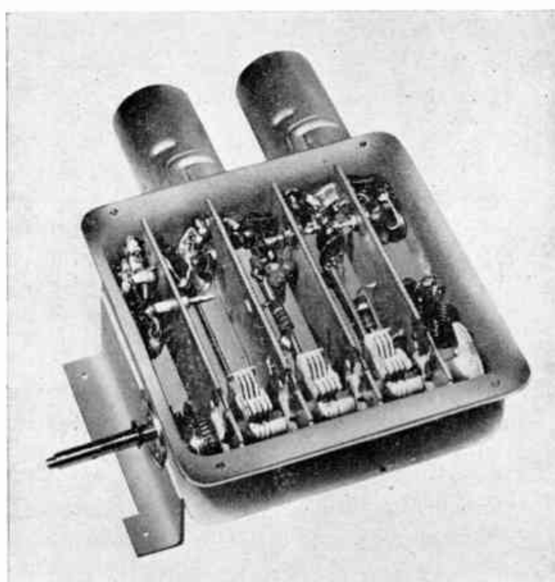


Fig. 1. A typical u.h.f. tuner (shown with cover removed).

† Sydney S. Bird & Sons Ltd., Poole, Dorset.

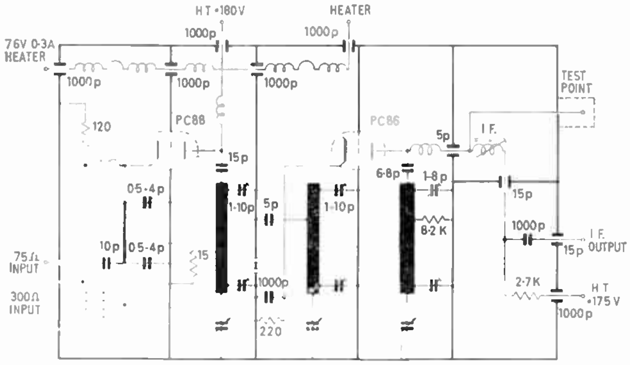


Fig. 2. Circuit diagram, u.h.f. tuner (series heater version).

obvious comments that an improvement in gain would be an advantage and an improvement in noise figure is highly desirable, the only specific requirement imposed by the proposed frequency allocation to this country would be an image frequency rejection ratio of at least 60 dB, the proposed channel allocation being such that in all areas an alternative local programme is present on the image frequency. The performance of the current u.h.f. tuner can just achieve 60 dB at the best point of the band, but is not better than 45 dB over the majority of the band; for any designer to claim more than 60 dB rejection would be optimistic. To improve the rejection ratio, a further tuned circuit must be introduced and the logical place for this would be in the aerial stage, in fact a tuner constructed in this manner by the author's Company has achieved better than 60 dB.

3. Coupling the U.H.F. Tuner to the Receiver

The coupling of the u.h.f. tuner to the receiver has been approached in a number of ways.

(i) The u.h.f. tuner, working independently of the v.h.f. tuner, couples into the i.f. amplifier of the receiver. With current v.h.f. signal gain of the order of 50 dB and u.h.f. signal gain of only 20–23 dB it is doubtful whether this method will prove adequate for use in this country, even bearing in mind the difference in the usable sensitivity due to the higher noise figure.

(ii) A second method, usually adopted with turret tuners, is to place, in one position of the turret, coil "biscuits" tuned to the intermediate frequency, thereby adding to the i.f. gain of the receiver the complete gain of a v.h.f. tuner. The comparative gains are now of the order of 80 dB u.h.f. and 50 dB v.h.f. In this case problems of stability are experienced, and, in existing tuners, there is the difficulty of stopping the oscillator as its anode dissipation must be controlled; in the case of tuners in which the oscillator provided bias for the mixer stage, the dissipation of the mixer valve must also be controlled.

(iii) With a new design, contacts would be provided to allow the output of the u.h.f. tuner to be applied to the grid of the mixer stage, thus reducing considerably the stability problem. The gains would now be approximately 50 dB for both v.h.f. and u.h.f.

(iv) Some Continental tuners utilize a small low-capacitance relay in the grid of the v.h.f. mixer stage to enable switching to be accomplished by the single control which changes from v.h.f. to u.h.f.

A current v.h.f. tuner has been designed to accommodate u.h.f. facilities. Means have been provided for switching at the grid of the mixer by the normal contact switching arrangement. Figure 3 shows this tuner and Fig. 4 its circuit diagram. The tuner is an incremental inductance type whereby channel switching is accomplished by moving a sliding contact down the length of the inductance system. At the low frequency end this contact runs on to a u.h.f. position, where oscillation ceases; sufficient resistance is used in the anode circuit to ensure that the valve will not exceed its rated anode dissipation, and will provide sufficient h.t. voltage to avoid poisoning the cathodes. The u.h.f. input socket is connected to the mixer grid and v.h.f. circuits are switched off frequency. Two methods are used to avoid over-dissipation of the mixer valve in the absence of oscillator volts, according to the type of valve. Either lower screen volts than would normally be the case are applied at all times or cathode bias is included at all times.

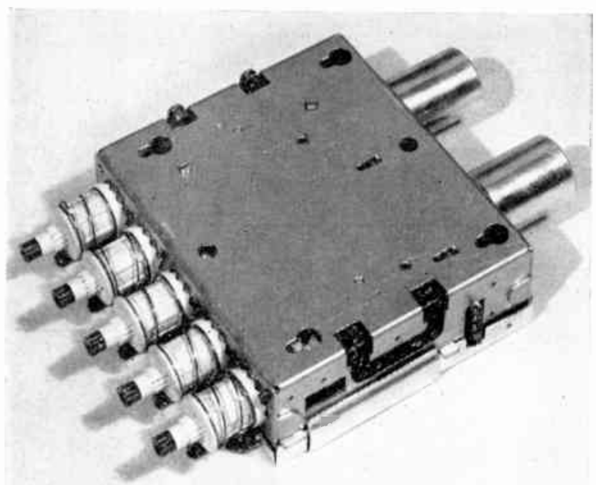


Fig. 3. The v.h.f. tuner described in the text.

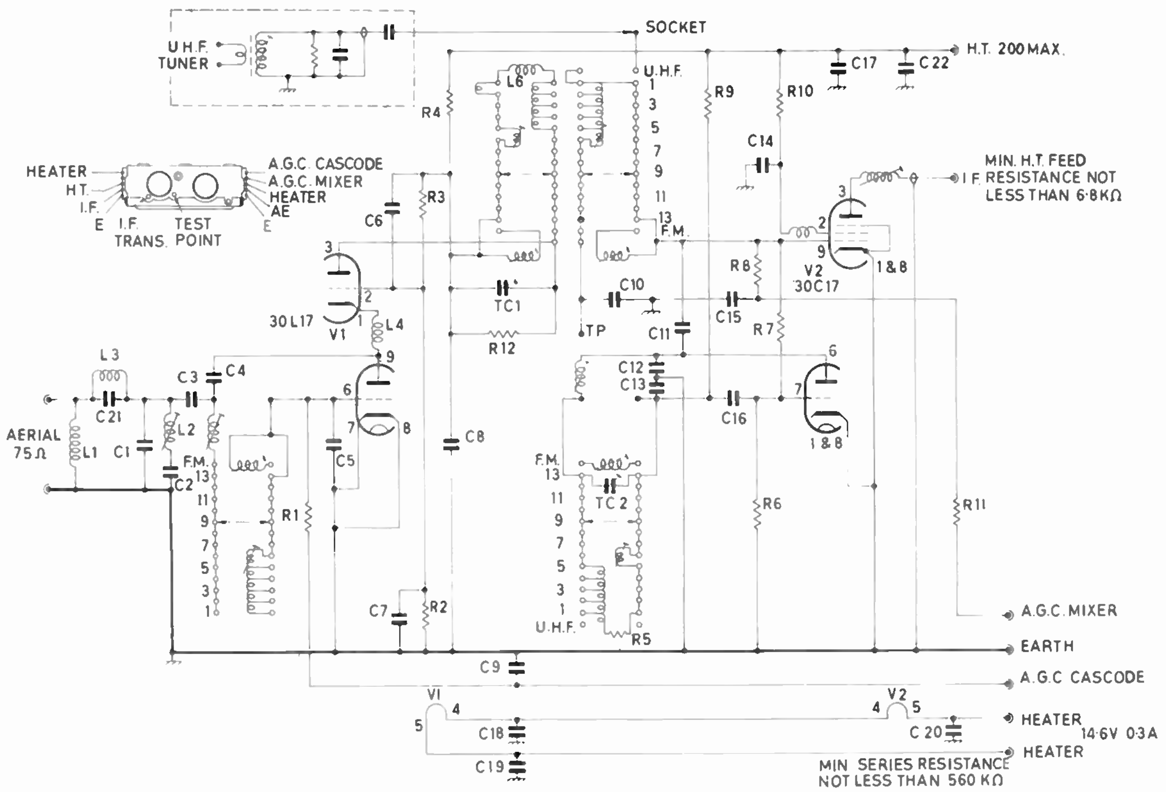


Fig. 4. Circuit diagram of v.h.f. tuner discussed in the text. Note the u.h.f. input position and a.g.c. control to mixer stage.

In this tuner the capacitance between the u.h.f. input socket and the rest of the tuning system is maintained sufficiently low to enable the u.h.f. tuner to be left functioning when using v.h.f. if this is desired, but to enable this to be done care must be taken to ensure that "pick-up" from other directions is also minimized. It has been stated by the valve manufacturers that the valves used in a u.h.f. tuner can be safely operated with only heater volts applied, without endangering the performance of the valve by cathode poisoning. This system has been employed in a u.h.f. convertor intended for Continental use without adverse effect and it is also used in many convertors of Continental manufacture. Therefore, if it is felt desirable to switch off the u.h.f. tuner it could be arranged that only the h.t. is switched; should the receiver designer wish to take all precautions a small h.t. voltage applied to avoid cathode poisoning might be used. Another alternative is to switch valve heater voltage as well as h.t. voltage in which case there will be a considerable warm-up time on switching to u.h.f.

To facilitate this switching and that of time-bases, i.f. circuitry, etc., the knob on the tuner which selects u.h.f. operation will have provision for a mechanical linkage to an external slider switch, so that by this selection alone the entire receiver switching is accomplished.

A relevant point which might be mentioned at this stage is that of a.g.c. When the receiver is operating on u.h.f., a.g.c. voltages cannot be applied to the u.h.f. tuner, consequently the a.g.c. is insufficient for optimum receiver performance. It is, therefore, desirable that the v.h.f. mixer stage, which is in use as the first i.f. amplifier in this mode of operation, be capable of control by a.g.c. voltages. Such a method is already in use for additional control on v.h.f. reception.

4. V.H.F. Tuner Design Requirements

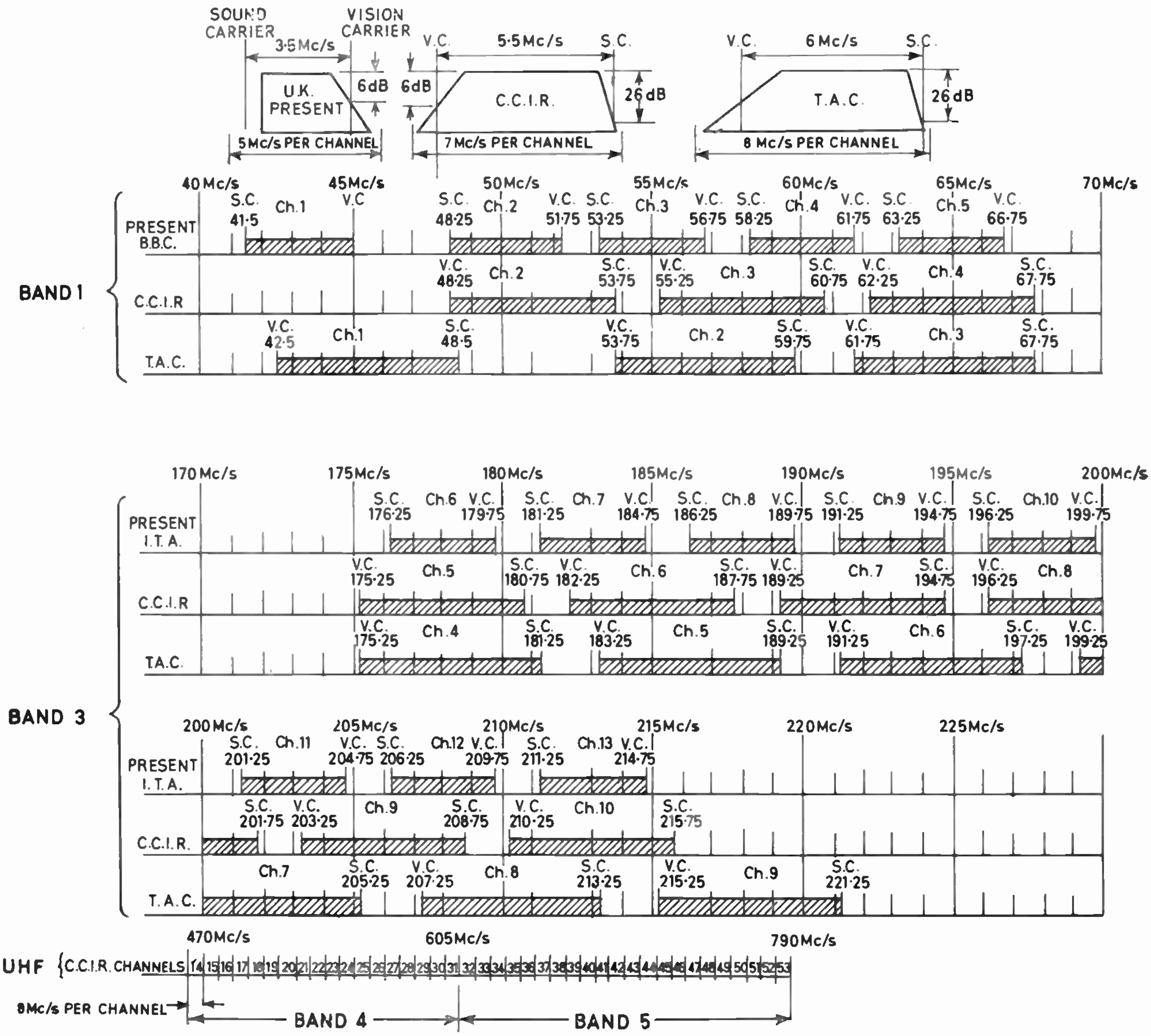
The more general design requirements of the v.h.f. tuner can be placed into two categories: firstly, those which are required in a v.h.f. tuner to be used in areas in which u.h.f. tuning is also to operate; secondly, those that may be required should the change be made from v.h.f. 405-line transmissions to v.h.f. 625-line transmissions after a given stated period of say, seven years.

In the former group probably the most serious problem is u.h.f. radiation from the v.h.f. tuner, due to harmonics of the v.h.f. oscillator. To overcome this, the design of the v.h.f. tuner must follow the methods employed on u.h.f. tuners, that is to say chassis and covers with a minimum number of holes, totally enclosed valve shields, and a true mating surface between all associated metalwork. Care should be taken on the precise point of earthing the v.h.f. tuner to the receiver. This can only be determined in a practical manner. The other major point for consideration is the method of providing connection from power supplies, a.g.c., i.f., etc., to the tuner. These connections must be earthed by suitable decoupling capacitors at the point of exit from the tuner. In the case of a standard turret tuner construction, the habit of running wires across the tuner from decoupling on the interscreen is not suitable, and further decoupling must be introduced at the point where the wires leave the tuner. In the case of a printed circuit design, the internal decoupling cannot prove satisfactory at these frequencies, and it is essential to provide decoupling immediately at the termination of connections to the outside of the tuner.

The author has already dealt with points that have to be considered when the tuner is used as an i.f. amplifier, the remaining consideration must be therefore what will happen if at some predetermined time the u.h.f. transmission is changed from 405 to 625 lines. This will of course necessitate the re-allocation of channel frequencies with an increase in channel bandwidth. Here again we must consider the implications with regard to the v.h.f. tuner system. Undoubtedly, in this respect the most flexible system will

Table 1

Present U.K., C.C.I.R., and proposed T.A.C. Frequency Allocations.



be a permeability tuner, providing the difficulties which are experienced in the design of permeability tuners could be overcome. Here the tuner could be operated now with a greater bandwidth sufficient to accommodate any proposed system. A re-adjustment to the tuner would only be necessary through the normal tuning system.

The turret tuner could also be viewed as being comparatively flexible. But in this, consideration must be given to the enormous dealer problems that will arise if, on one stated day, transmissions change from 405 to 625.

Incremental inductance tuners could be considered as the least flexible; but bearing in mind the problems associated with the reloading of turrets, this type of tuner is probably of equal merit, since it may be designed to include means of interchanging the coil assemblies.

5. Tuner Considerations at a Changeover of Standards

To assess the magnitude of the problems associated with a rapid changeover, the systems which might be expected to be used will be considered. One of the main arguments which has been put forward for the introduction of 625-line transmissions to this country has been the need to make the industry more flexible with regard to exports, bearing in mind the implications of the Common Market. To achieve this flexibility it has been stated that a similar system to the Continent should be adopted in the United Kingdom thereby enabling receivers to be exported direct from home market production, without incurring the overhead costs which are necessary to run small specialized production units solely for export. If this argument is followed to its logical conclusions it may be assumed that C.C.I.R. channel allocations will be adopted. If this in fact be true it is easy to see a comparatively simple solution to the problems associated with turret and incremental inductance tuners.

Table 1 shows the frequency allocations of the current British channels together with C.C.I.R. and T.A.C. propositions. It is customary in tuner design to choose a bandwidth about a half megacycle greater than is required above and below the channel. This is to avoid a too stringent requirement in tuner tracking, etc. By taking this principle a little further, Table 2 may be drawn up which shows that a 14-channel tuner may be designed for 405/625 line use which, upon changeover, will necessitate only channel selection and fine tuner adjustment by the user. Tuners made in this way would probably be aligned in positions 4 and 12 with attention being paid to the excess bandwidth required in position 7.

Table 2

Scheme based on the C.C.I.R. System to show how a 14-channel tuner may be constructed to receive both 405 and 625-line transmissions.

Arbitrary Channel No.	405-line Channel No.	C.C.I.R. Channel No.	Additional Bandwidth Requirement
1	1		
2	2	2	
3	3		
4	4	3	+ 1 Mc/s high-frequency side
5	5	4	
6	6	5	
7	7	6	+ 1 Mc/s low-frequency side
8	8		
9	9	7	
10	10	8	
11	11		
12	12	9	+ 1 Mc/s high-frequency side
13	13	10	
14		11	

Should, however, the T.A.C. channel allocation be used, with the attendant disadvantage of insufficient channels to provide national coverage on two programmes, then a combined system cannot be embodied in the incremental tuner. In this case provision would have to be made for completely removing the coil assembly, which is probably equivalent to interchanging coil "biscuits" in a turret tuner.

6. Acknowledgments

The writer is indebted to the Directors of Sydney S. Bird & Sons Ltd., for permission to publish this paper, and to Mr. P. C. Ganderton, Chief Engineer, for comments and advice.

Manuscript received by the Institution on 20th February, 1962 (Contribution No. 55/T14.)

# of current interest . . .

Paper Tape Handling Conventions

Occasionally confusion is caused in the handling of paper tape due to there being no clear indication on tape as to which way round the information is punched. In addition, it is not always clear which way punched tape should be fed into a reader. To eliminate these sources of confusion the following two conventions are suggested by the Data Processing Section of the Electronic Engineering Association.

Paper Tape Convention. It is proposed that paper tape be marked with arrows printed in low density red ink showing both the direction of motion and also defining the "printed face". (The copies of the conventions show the precise position of the arrow in relation to the various coding arrangements specified in B.S. 3480 : 1962.) The spacing of the arrows along the tape should be not more than 6 in. or less than 4 in. throughout the whole of its length.

Paper Tape Equipment Convention. In order that paper tape be fed to readers and punches in the correct manner, it is necessary to define whether the "printed face" should be upwards or downwards in the equipment. It is therefore proposed that an arrow be placed on the equipment in close proximity to the gate, indicating which way up the tape should be placed. A second arrow may be placed on the equipment to indicate the intended forward direction of movement of the tape, if so desired.

At this stage the proposals are a minimum and the Association hopes that once these procedures become normal practice it will be possible to broaden the scope of the conventions.

Copies of the conventions are being issued widely to manufacturers of peripheral equipment and others. Further copies are available from the Technical Secretary, Electronic Engineering Association, 11 Green Street, Mayfair, London, W.1.

Transistor Repeaters in U.K.–Belgium Cable

Transistors will be used in the U.K.–Belgium submarine cable when new repeaters are inserted in the cable in mid-1964. This will be the first British cable using transistors and will be a major advance in the development of submerged repeater techniques. It will provide a big increase in circuit capacity.

The cable, laid in 1948 between St. Margaret's Bay, Kent, and La Panne, at present carries 216 telephone circuits and by inserting two transistor repeaters the capacity will be raised to 420 circuits of standard C.C.I.T.T. quality for 4 kc/s spacing. The present system uses frequency bands of 60–956 kc/s and 1152–2048 kc/s for the 216 circuits now in service; the new system will use bands of 312–2044 and 2544–4276 kc/s. The cable, which is 47·5 nautical miles long, is a single coaxial cable of 1·7 in. diameter in which the polythene core incorporates a partial air space. It is the largest type of submarine coaxial telephone cable in service but, since it was laid, the rapid development of submerged repeaters has made the use of such large and expensive cables less attractive.

Transatlantic Data Transmission Service

The first international high-speed data transmission public service was introduced on 5th September by the British Post Office between London and New York. Customers in London rent special lines to a Post Office switching centre where "datatelex" calls are connected to customers in New York over a circuit in the transatlantic telephone cable. The New York end of the service is operated by the Radio Corporation of America (Communications) Inc. Datatelex calls to New York cost £1 6s. 8d. a minute with a minimum of 10 minutes. The speed of transmission is generally 1200 bits a second.

Air Traffic Engineering as an Academic Subject

The Electrical Engineering Department of the University of Birmingham is offering among its post-graduate courses one on Air Traffic Engineering. This is believed to be a new topic in the British Universities. Research work in radar systems, arrays, displays, communications, computers, etc., has, of course, been carried on in the Department for many years; it is the integrated course of instruction which is new.

Students who contemplate entering industry or government establishments concerned with air traffic work—or who are already in such employment and are returning to the University—may take a 12-months' full-time course which lays a sound academic foundation for research and development work in Air Traffic Engineering and leads to the M.Sc. degree (or the Diploma in Graduate Studies) by examination. A combination of the following subjects has been thought most suitable: information theory and statistics, mathematics, communication systems, radar and navigational aids, computers, and air traffic systems. In the latter subject, in particular, many outside specialist lecturers contribute to the course, coming mainly from the Ministry of Aviation (who have given much support to the course) but also from industry.

The course is open to all students with a good first degree (or similar qualification) in an appropriate subject. Enquiries should be addressed to the Deputy Registrar, The University, Birmingham 15.

U.D.C. 621.397.332

Switchable 405/625 Line Time-Bases

By

A. CIUCIURA, B.Sc.(Eng.)†

Presented at a Television Group Symposium on "Dual Standard Television Receivers" in London on 7th February 1962.

Summary: The design of switchable line time-bases can be channelled along two lines: constant flyback time and constant flyback to scan ratio. Some theoretical aspects connected with the two modes of operation are reviewed and possible solutions are illustrated with practical designs.

1. Introduction

In preparation for conversion from the 405-line system to a higher definition system in the United Kingdom, much attention has been given in television laboratories to the receivers for the transition period. The views of one laboratory on the problems of switchable line time-bases are offered in this paper in which the basic ideas evolve along the two possible approaches. The subsequent treatment is limited to evaluation of differences between operation on the 405-line system and the 625-line system within the framework of these approaches.

2. Design Target

In a modern television receiver, the line time-base performs the following functions:

(a) line scan

(b) e.h.t. generation

(c) heater supply to e.h.t. rectifier

(d) source of a_1 (and often focus) potential for c.r.t.

The scan itself requires linearity correction and waveform shaping (s-correction). Furthermore, it is usual to provide third harmonic tuning. Most of the functions are critical in application; that is, any change in performance, especially when switched from one system to another, is liable to be noticed by the viewer.

Thus, the aim of a switchable line time-base is to provide these functions at the same level on 405-line and 625-line television systems in spite of over 50% change in line frequency and about the same in power. All the functions of the line time-base must be taken into consideration. The changeover must be done with a minimum number of switch operations; the necessary switching points must be limited to low voltage portions of the circuit.

Finally, the operating conditions of the valves used in the time-bases must be kept within the published data limits on each line standard.

† The Mullard Radio Valve Co. Ltd., Mitcham Junction, Surrey.

3. Third Harmonic Tuning

There are still some reservations in the industry as to the practicability of third harmonic tuning in a mass produced receiver. In the case of switchable line time-bases it is tempting to abandon it and thus gain a degree of freedom in design. This is particularly attractive in the constant flyback to scan execution.

Such an approach is quite practicable for a time-base handling a moderate amount of power. As the power handling capacity of the transformer is increased, there is a point where it is difficult to avoid some degree of tuning. In a transformer designed for operation on one line frequency a small amount of tuning could pass unnoticed; however, in a dual system the tuning produces differences in e.h.t. when switching from one system to another. Perhaps it can be shown that there exists a relationship between two high order harmonics which gives the same e.h.t. value on both systems. In practice, such an approach could be at least as sensitive to production tolerances as the third harmonic tuning. Thus, unless drastic steps are taken to reduce the effect of leakage inductance, third harmonic tuning seems to be indispensable.

4. Possible Approaches

Considering operation of a switchable line time-base from every aspect, the design can be based on one of the following factors which remain the same on switching from one system to another: constant flyback time or constant flyback to scan ratio. Individual requirements may justify some departure from these approaches; nevertheless it is always useful to develop main reasoning on these two lines.

4.1. *Constant Flyback Time*

In this mode of operation the flyback time is maintained the same on the 405-line and the 625-line system. The resulting flyback-to-scan ratio is shown in Fig. 1 as a function of flyback time. This approach is often taken in multi-standard receivers on the Continent. However, the line systems used there are 625/819 lines and the difference in frequencies is about a half that of the 405/625 dual system considered here.

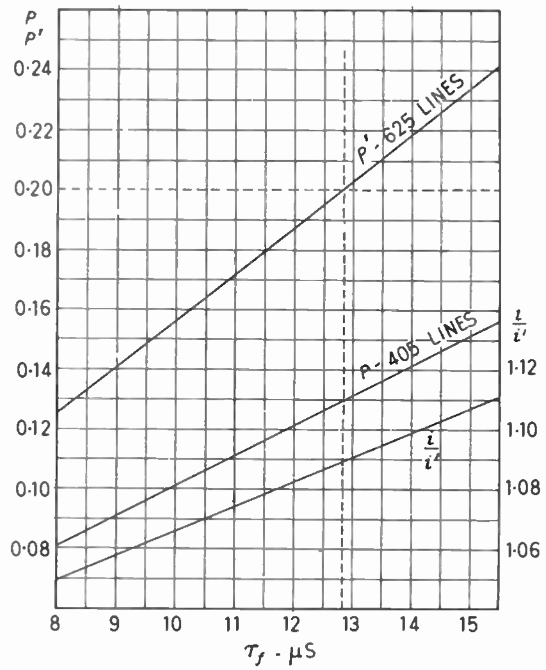


Fig. 1. p, p' and i/i' as a function of τ_f.

The main advantage of this approach is that it does not require any switching of the output transformer; all the switchings can be at d.c. Since the flyback time is constant, the third harmonic tuning is maintained automatically on both systems.

On the debit side is poor efficiency on the 405-line system. Both the anode dissipation and the average anode current are higher than on the 625-line system. Since both factors affect the size of the valves needed, this may be a consideration in addition to the higher h.t. power requirements. Unless stabilization is used, e.h.t. regulation is poor and there may be some difficulties with meeting the heater voltage limits of the e.h.t. rectifier.

Switching must be used to provide the following:

(a) in a stabilized line time-base to reduce anode dissipation on the 405-line system, and in the non-stabilized time-base to bring the end of scan anode potential into the region below the "knee".

(b) maintain constancy of the a_1 supply to the cathode-ray tube.

(c) s-correction.

Steps must be taken to reduce the change in the heater voltage of the e.h.t. rectifier when switching from one line system to another.

It will be shown later that the nature of design provides for some inequality of e.h.t. and scan on switching from one line system to another.

4.2. *Constant Flyback-to-Scan Ratio*

Examination of the synchronizing waveforms for the 405-line and the 625-line systems (see Table 1) shows that the ratio of blanking time to the total period time is the same. It is reasonable to assume that any margin of safety required by one system will also meet the needs of the other system. Thus, a constant ratio seems to be a fundamentally sound principle on which to base the design of a switchable line time-base.

With this mode of operation there is complete freedom for the selection of a suitable flyback time. A range of possible values is shown graphically as a function of p in Fig. 2. The line time-base efficiency approaches that of a single standard system, and the existing range of valves can therefore be used. It is suitable for application in stabilized and non-stabilized designs. Equality of scan, e.h.t., heater voltage to the e.h.t. rectifier, and the a_1 potential is automatically maintained.

However, the constant ratio design requires switching of deflection coils on changing from one system to another. Furthermore, there is also a need for switching of third harmonic tuning. In addition a switch is required for s-correction.

4.3. *Constant H.T. Power*

It should be possible to develop a design on the basis of constant power taken from h.t. line. Unfortunately, it would have to be a mixture of the two design methods described. The total power taken would be higher than with the constant ratio design, and the

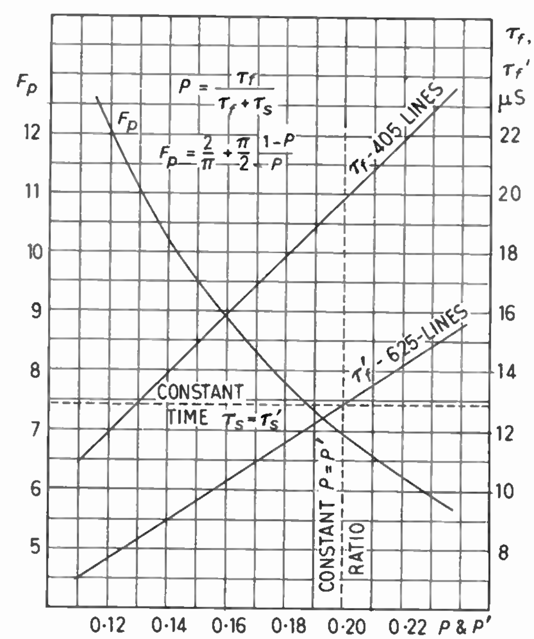


Fig. 2. F_p, τ_f and $\tau_{f'}$ as a function of p and p'.

switching arrangement would involve all the complexities of the two systems described above.

Table 1

Systems Data†

(where applicable the data refer to 50% amplitude)

	405	625
Number of lines	405	625
Line frequency, c/s	10 125	15 625
Line period ($\tau_f + \tau_s$), μs	98·76	64·0
Line sync. pulse, μs	9 ± 1	4·5 – 4·9
Front porch, μs	1·5 – 2	1·3 – 1·8
Line sync. pulse + back porch, μs	16·5 ± 0·5	—
Line blanking, μs	17·5 – 19·0	11·8 – 12·3
$\frac{\text{Line blanking}}{\text{Line period}}$	0·177 – 0·192	0·184 – 0·192

5. Some Theoretical Considerations

5.1. *Systems Data and Nomenclature*

In this paper, plain symbols are used for general considerations and also to denote operation on the 405-line system; a symbol with a dash is used to signify operation on the 625-line system.

L_y inductance of deflection coils

i peak to peak current in deflection coils

I average current (anode or booster diode)

V_{ys} average voltage across deflection coils during scan

v_{yf} peak voltage across deflection coils during flyback

f line scan frequency

n number of turns on line output transformer

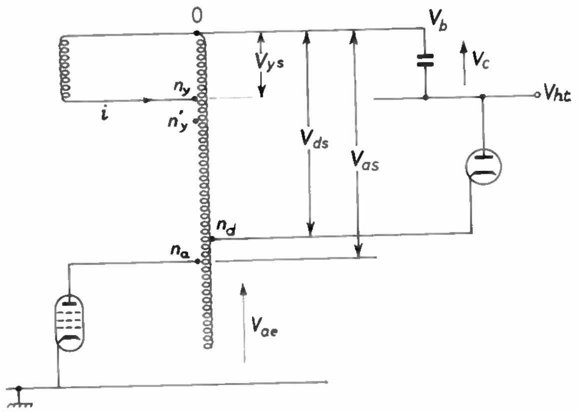


Fig. 3. Schematic representation of the line time-base.

† Information on the 405-line system extracted from B.B.C. Technical Information Sheet No. 2020/2, February 1962. Information on the 625-line system extracted from Vol. 3 of C.C.I.R. 9th Plenary Assembly Report 124, pages 167–170.

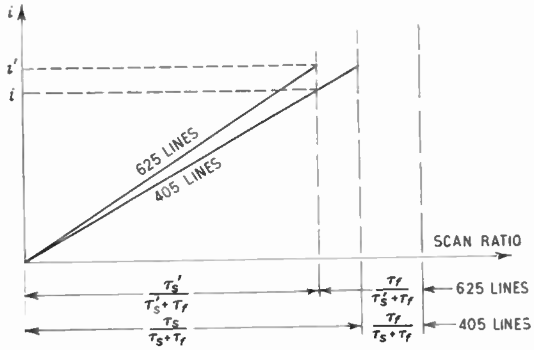


Fig. 4. Effect of constant flyback time on the deflection coils current.

With reference to Fig. 3, notation n_y implies number of turns between the taps $0-n_y$, n_d — between the taps $0-n_d$ and so on.

$$\tau_f = \text{flyback time}$$

$$\tau_s = \text{scan time}$$

$$p = \frac{\tau_f}{\tau_s+\tau_f}; \qquad p = f\tau_f; \qquad 1-p = f\tau_s$$

$$\frac{\tau_f}{\tau_s} = \frac{p}{1-p}; \qquad \frac{\tau'_f}{\tau'_s} = \frac{p'}{1-p'}$$

Note: If
$$\frac{\tau_f}{\tau_s} = \frac{\tau'_f}{\tau'_s}$$

then

$$\frac{p'}{1-p'} = \frac{p}{1-p}; \qquad p(1-p') = p'(1-p); \qquad p = p'$$

Thus constant $\frac{\tau_f}{\tau_s}$ and constant p are synonymous.

$F_p = \frac{2}{\pi} + \frac{\pi}{2}\frac{(1-p)}{p} \simeq \frac{\pi}{2}\frac{(1-p)}{p}$ — ratio of the peak voltage during flyback to the average voltage during scan.‡

In Fig. 2, τ_f, τ_s and F_p are given as a function of p and p'.

In Fig. 1, p and p' are shown as a function of τ_f.

5.2. *Constant Flyback Design*

5.2.1. Scanning current

While the peak-to-peak current in the deflection coils required to scan the useful screen area is independent of repetition frequency, some complications arise if the flyback time is kept constant. If equal scanning currents are provided on both systems, the situation is as shown in Fig. 4; at any part of the scanning cycle

‡ C. J. Boers and A. G. W. Uitjens, "Practical considerations on line time-base output stages with booster diode circuits", *Electronic Application Bulletin*, 12, No. 8, page 140, August 1951.

the current in the deflection coils (and therefore scan amplitude) is lower on 405 lines than on 625 lines. In order to bring the two scans to the same values, the following requirement must be satisfied:

$$\frac{i'}{\dfrac{\tau_s'}{\tau_s'+\tau_f}} = \frac{i}{\dfrac{\tau_s}{\tau_s+\tau_f}}$$

Therefore $$\frac{i}{i'} = \frac{f\tau_s}{f'\tau_s'} = \frac{1-p}{1-p'} \qquad \ldots\ldots(1)$$

A plot of i/i' as a function of flyback time is shown in Fig. 1. The graph indicates that the disparity in currents increases with the flyback time and at $\tau_f = 12{\cdot}8$ µs it is equal to 1·09.

5.2.2. Voltage across deflection coils

The average voltage during scan is given by the relationship

$$V_{ys} = L\frac{i}{\tau_s} = L\frac{i}{\tau_f}\frac{p}{1-p} \qquad \ldots\ldots(2)$$

and the ratio of voltages obtainable on the two systems is

$$\frac{V_{ys}'}{V_{ys}} = \frac{i'}{i}\frac{p'}{p}\frac{1-p}{1-p'}$$

If it is desired to maintain the same scan amplitudes, then i'/i should be as indicated by eqn (1). Therefore

$$\frac{V_{ys}'}{V_{ys}} = \frac{p'}{p} = \frac{f'}{f} \qquad \ldots\ldots(3)$$

Thus the ratio is a function of line scanning frequencies only, and for the 405/625 system it is 1·54.

The peak voltage across deflection coils during flyback is given as

$$v_{yf} = F_p V_{ys} \simeq \tfrac{1}{2}Li\frac{\pi}{\tau_f} \qquad \ldots\ldots(4)$$

Since the flyback time is the same for both systems, v_{yf} is a function of deflection current only.

Thus the peak voltage and therefore e.h.t. can be kept approximately the same on both line systems if the deflection current is kept the same. The latter however, disagrees with the requirements stated in eqn (1). Unless a resort is made to special switching, a compromise has to be made. Probably the best approach is to increase the peak current according to eqn (1). The overall result of such a design would be for the scan on 405 lines to be below that on 625 lines by a factor

$$\sqrt{\frac{1-p}{1-p'}}$$

and the e.h.t. on 405 lines above that on 625 lines by

$$\frac{1-p}{1-p'}$$

Poorer e.h.t. regulation on 405 lines will tend to diminish the differences under normal operating conditions. On the other hand, care must be taken not to exceed the maximum permissible value of e.h.t. for the cathode-ray tube and the maximum peak voltages applied to the scanning valves at zero beam current.

5.2.3. Boost line voltage

The voltage available on the boost line is

$$V_b = V_{ht} + V_c$$

where $$V_c = \frac{n_d}{n_y}V_{ys} \qquad \ldots\ldots(5)$$

In practice V_{ht} may be around 220 V, and V_c' about 700 V. This would result in V_c being 450 V and

$$V_b = 670\text{ V}$$

$$V_b' = 920\text{ V}$$

Normally the first anode of the cathode ray tube is supplied from the boost line. In order to keep the screen brightness on the two systems at the same level, the V_{a1} will have to be closely controlled. In the case under consideration it is obvious that switching must be used. Furthermore, if the flyback time is made short, there is a possibility that the boost voltage may be too low to satisfy minimum requirement of V_{a1}.

5.2.4. Anode potential at the end of scan

The potential is given as:

$$V_{ae} = V_b - \frac{n_a}{n_y}V_{ys} = V_{ht} + V_c - \frac{n_a}{n_y}V_{ys}$$

Substituting for V_c (eqn (5)) and rearranging

$$V_{ae} = V_{ht} - V_{ys}\left(\frac{n_a}{n_y} - \frac{n_d}{n_y}\right) \qquad \ldots\ldots(6)$$

Let $$V_{ys}\left(\frac{n_a}{n_y} - \frac{n_d}{n_y}\right) = A$$

Then $$\frac{A'}{A} = \frac{f'}{f} = 1{\cdot}54$$

And $$V_{ae} = V_{ht} - A = V_{ht} - 0{\cdot}65A'$$

$$V_{ae}' = V_{ht} - A'$$

Therefore $$V_{ae} - V_{ae}' = 0{\cdot}35A'$$

Now A' is the voltage which, when subtracted from V_{ht}, gives V_{ae} (transformer design takes care of that). Covering all practicable values of V_{ht} and V_{ae} it will be found that the end of scan anode voltage on 405 lines exceeds the voltage on 625 lines by

$$V_{ae} - V_{ae}' = 60\text{ V} \pm 15\text{ V}$$

The effect of this change in the anode potential at the end of scan will depend on the mode of operation of the time-base.

In a stabilized line time-base, when the peak anode current is controlled on the control grid by a feedback network, switching from 625 lines to 405 lines will result in the anode voltage increasing by some 60 V. Since the anode dissipation is proportional to the anode voltage, there will be a corresponding increase. Typically, it may be in the order of 5 W for 110 deg tube time-bases. Since this is not the only cause for the increase in the anode dissipation, steps will have to be taken to bring the voltage down.

In a line time-base where the peak anode current is stabilized by bottoming the valve (below the "knee" operation) the difference in the end of scan voltage is much more serious. On switching from 625 lines to 405 lines, the valve will be brought out of the "knee" with the consequent loss of control and large increase in scan and e.h.t. In order to avoid such an occurrence, the h.t. line on 405 lines must be lowered so as to bring the valve back into the controlled region (below the "knee").

5.2.5. Average anode current

In a line time-base,

$$V_{ht} I_a = P_l + P_a + P_d$$

where I_a = average current

P_a = anode dissipation of line output valve

P_d = anode dissipation of booster diode

P_l = circuit losses.

Since $P_a + P_d = I_a V_{ae}$

therefore the average current can be expressed as

$$I_a = \frac{P_l}{V_{ht} - V_{ae}} \qquad \ldots\ldots(7)$$

Rearranging eqn (6)

$$V_{ht} - V_{ae} = V_{ys}\left(\frac{n_a}{n_y} - \frac{n_d}{n_y}\right)$$

Therefore, using eqn (3)

$$I_a = \frac{P_l}{V_{ys}\left(\frac{n_a}{n_y} - \frac{n_d}{n_y}\right)}$$

If P_l contained only losses occurring during flyback, it would be proportional to frequency and

$$\frac{I_a}{I'_a} = \frac{kf}{kf'}\frac{f'}{f} = 1$$

However, P contains also losses which are largely frequency independent.

For this type of loss

$$\frac{I_a}{I'_a} = \frac{f'}{f} = 1{\cdot}54$$

Thus the true ratio I_a/I'_a must be larger than 1 and smaller than 1·54.

Losses which are frequency independent are:

- resistive losses in the deflection coils and transformer,
- losses in linearity sleeve,
- heater power of the e.h.t. rectifier,
- d.c. leakages of the e.h.t. system,
- power taken from the boost line for auxiliary services.

In the group of frequency dependent losses are:

- ferrite core loss,
- dielectric loss,
- unrecoverable losses due to leakage inductance ringing.

For a time-base using a linearity sleeve, the two groups of losses are about equal, and therefore the ratio I_a/I'_a can be taken as

$$\frac{I_a}{I'_a} \simeq \frac{1}{2}\left(1 + \frac{f'}{f}\right) \simeq 1{\cdot}3 \qquad \ldots\ldots(8)$$

A somewhat lower ratio is obtainable with a saturated reactor linearity control.

5.2.6. E.h.t. load

The considerations in the previous paragraph do not take into account the power taken from the e.h.t. supply or boost line. If any loading of this nature is imposed on the time-base, then the additional average current taken by the line output valve is

$$I_{a(eht)} = \frac{P}{V_{ht} - V_{ae}}$$

These loads are frequency independent, hence

$$\frac{I_{a(eht)}}{I'_{a(eht)}} = 1{\cdot}54$$

Therefore the total average current taken by the line time-base may be around 40% higher on 405 lines than on 625 lines. Since the power consumption of a line time-base is a major proportion of the total power in the receiver, this increase is a very important aspect of design.

5.2.7. Anode dissipation

This can be approximately expressed as

$$P_a \simeq \tfrac{3}{4} I_a V_{ae}$$

In Section 5.2.6 it was estimated that the average current on the 405-line system would be 1·4 times that on the 625-line system. This factor alone would increase anode dissipation by the same amount. Furthermore, the end-of-scan voltage can increase by about 60 V. In a stabilized line time-base this would imply about a 100% increase. Thus, if no steps are taken to reduce the h.t. line by the same amount, the anode dissipation of a stabilized time-base working on the 405-line system can be 2·8 times that on the 625-line system.

In the case of a non-stabilized line time-base the h.t. will be reduced for operational reasons.

5.2.8. Peak flux density

Using normal notation, this is given as

$$B_{(\mathrm{pk})} = B_0 + B_{\max}$$

where $$B_{\max} = \frac{\frac{1}{2}Li}{n_y A}\, 10^8 \text{ gauss}$$

and B_0 is normally given in published data as a function of

$$\frac{I(n_a - n_d)}{l}$$

where l = magnetic path length.

For ferroxcube material the maximum peak flux density is determined by the temperature of the core. In practice this means that the peak flux density for the 625-line system should be about 10% below that for the 405-line system. It has been shown above that the peak current i is about 8% larger, and the average current I about 40% larger, on a 405-line system than on a 625-line system. For this reason the design of the magnetic circuit must be based on the requirements of the 405-line system.

For a given magnetic core the design involves selection of a suitable air-gap. Large currents call for a larger air-gap and this in turn lowers incremental permeability. Since there is an optimum ratio of transformer primary inductance to the deflection coil inductance, a transformer with a larger gap must carry a larger number of turns.

When such a transformer is switched to the 625-line operation, reduction in peak and average current causes reduction in flux density well below the 10% level required. Since no advantage can be taken of this (because of fixed design), such a transformer would have to carry more turns than the transformer specifically designed for the 625-lines.

5.2.9. Heater voltage

Let v be average heater voltage during scan, $v_{(\mathrm{pk})} \sin \omega t$ heater voltage during flyback and V the r.m.s. heater voltage, then:

$$\frac{\int_0^{\tau_f} v_{(\mathrm{pk})}^2 \sin^2 \omega t\, dt + \int_0^{\tau_f} v^2\, dt}{\tau_s + \tau_f}$$

$$= \frac{\frac{1}{2} v_{(\mathrm{pk})}^2 \tau_f + v^2 \tau_s}{\tau_f + \tau_s}$$

$$= v^2 f(\tfrac{1}{2}\tau_f F_p^2 + \tau_s)$$

$$\simeq v^2 f \tau_s \left(\frac{\pi^2}{8}\frac{\tau_s}{\tau_f} + 1\right)$$

but $$v = V_{ys}\frac{n_h}{n_y}$$

where n_h/n_y is the transformation ratio between the deflection coils tap (n_y) and the heater winding n_h. Substituting for V_{ys} from eqn (2),

$$v = L_y \frac{i}{\tau_s}\frac{n_h}{n_y}$$

and $$V^2 \simeq f L_y^2 \left(\frac{n_h}{n_y}\right)^2 i^2 \frac{1}{\tau_s}\left(\frac{\pi^2}{8}\frac{\tau_s}{\tau_f} + 1\right)$$

Normally, $$\frac{\pi^2}{8}\frac{\tau_s}{\tau_f} > 1$$

therefore

$$V^2 \simeq f L_y^2 \left(\frac{n_h}{n_y}\right)^2 i^2 \frac{\pi^2}{8\tau_f} \qquad \ldots\ldots(9)$$

For the constant flyback operation τ_f is the same for both systems, hence

$$\frac{V'}{V} \simeq \frac{i'}{i}\sqrt{\frac{f'}{f}}$$

Substituting for i'/i from eqn (1)

$$\frac{V'}{V} \simeq \frac{1-p'}{1-p}\sqrt{\frac{f'}{f}} \qquad \ldots\ldots(10)$$

It is necessary to stress that the above calculations apply to an idealized case. In practice the power contributed by ringing may modify the results.

The necessary equalization of heater voltage between the two line frequencies can be obtained by means of an inductance connected in series with the heater of the e.h.t. rectifier. In practice the waveforms involved may be quite complex and applicable to the particular case only. For this reason the value of the required inductance can be best determined experimentally. Some practical advantages can be gained by using an e.h.t. rectifier having low heater voltage.

Insertion of a choke in series with the heater increases the effective impedance of the heater supply. If the choke impedance is much higher than the heater impedance, any change in the input voltage to the combination of the two results in an equal change in the heater current. For a thermionic valve the permissible heater voltage tolerances are nearly twice as wide as the heater current tolerances. Thus the above arrangement restricts the permissible tolerance of the input voltage by about a half.

5.3. *Constant Ratio Design*

5.3.1. The ratio

An immediate advantage of the constant ratio design is the fact that the peak current flowing through the deflection coils is identical for both systems. Furthermore, for any waveform on the line output transformer the ratio of peak voltage generated during flyback to the voltage generated during scan is given by

$$F_p = \frac{2}{\pi} + \frac{\pi}{2}\frac{\tau_s}{\tau_f} \simeq \frac{\pi}{2}\frac{(1-p)}{p}$$

If the same value of p is used on both standards, F_p remains constant. Thus there exists a possibility of maintaining the peak voltage on the transformer the same for the two line systems. The same peak voltage and F_p imply that the e.h.t., boost line potential and heater voltage to the e.h.t. rectifier are the same for the 405 and 625-line systems. However, these advantages must be paid for in other ways.

5.3.2. Voltages on the transformer

The average voltage developed across deflection coils during scan is

$$V_{ys} = L_y \frac{i}{\tau_s} = L_y if \frac{1}{1-p}$$

Since p is constant by definition, the voltage is proportional to frequency. However, it can be made the same for the two line systems by switching the deflection coils across a different number of turns on the line output transformer when changing from one system to another. The turns should be in the ratio (see Fig. 3) of

$$\frac{n'_y}{n_y} = \frac{f'}{f} \qquad \text{......(11)}$$

With such switching, for any tap n_x on the transformer

$$\frac{n_x}{n'_y} f' = \frac{n_x f}{n_y f'} f' = \frac{n_x}{n_y} f = K \qquad \text{......(12)}$$

Thus the peak anode voltage

$$v_{a(\text{pk})} = F_p V_{ys} \frac{n_a}{n_y} = F_p \frac{n_a}{n_y} Lif \frac{1}{1-p}$$

$$= F_p KLi \frac{1}{1-p}$$

and this is frequency independent.

Similarly the boost voltage

$$V_c = V_{ys} \frac{n_d}{n_y} = L_y if \frac{1}{1-p}\frac{n_d}{n_y} = L_y Ki \frac{1}{1-p}$$

The same applies to the heater voltage of the e.h.t. rectifier. Substituting for $\tau_f = p/f$ in eqn (9) and incorporating eqn (12)

$$V = L_y Ki \frac{\pi}{2} \frac{1}{\sqrt{2p}}$$

5.3.3. Change in flyback time

In order to meet the requirements of the constant ratio, the flyback times have to satisfy the following relationship:

$$\frac{\tau_f}{\tau'_f} = \frac{p/f}{p'/f'} = \frac{f'}{f} \quad \text{since } p \text{ is constant} \quad \text{......(13)}$$

Let the total stray capacitance at the anode tap of the line output transformer be C_t and the total transformer inductance referred to the same point be L_t. When the deflection coils are connected across n_y turns the total LC product is:

$$LC = \left[C_t + C_y \left(\frac{n_y}{n_a}\right)^2\right] \frac{L_t L_y (n_a/n_y)^2}{L_t + L_y (n_a/n_y)^2} \quad \text{...(14)}$$

Line output transformer is always designed so that $L_y(n_a/n_y)^2$ is 5 to 8 times L_t. Thus to a first approximation the inductance component of eqn (14) can be written as

$$L_y \left(\frac{n_a}{n_y}\right)^2$$

With this amount of inductance, the circuit is resonant during flyback at frequency $f_f = 1/2\tau_f$ and

$$\omega_f^2 = \frac{1}{L_y \left(\frac{n_a}{n_y}\right)^2 C_t}$$

$$C_t = \frac{\tau_f^2}{\pi^2 L_y \left(\frac{n_a}{n_y}\right)^2}$$

For any line output valve there is a maximum peak anode voltage that may be used, and normally the design is fixed somewhere near this value. The peak anode voltage with respect to chassis is given by:

$$v_{a(\text{pk})} = V_b + V_{ys} \frac{n_a}{n_y} F_p.$$

$$v_{a(\text{pk})} - V_b \simeq L_y \frac{i}{\tau_s} \frac{\pi}{2} \frac{n_a}{n_y} \frac{\tau_s}{\tau_f} \simeq L_y i \frac{\pi}{2\tau_f} \frac{n_a}{n_y}$$

Substituting for τ_f in the formula for C_t,

$$C_t = \frac{L_y i^2 \pi}{4(v_{a(\text{pk})} - V_b)^2}$$

Since L_y is proportional to the square of turns on the deflection coils, then for a particular execution of deflection coils, Li^2 is a constant. In such circumstances, the value of C_t is also a constant, dependent only on deflection VA.

For the present range of tubes and $L_y = 4{\cdot}8$ mH, i is about 1·9 A. Taking

$$v_{a(\text{pk})} - V_b = 5 \text{ kV}$$

$$C_t = 170 \text{ pF}$$

Measured self-capacitance of the deflection coils quoted above was found to be 150 pF. Since the value of C_t refers to the anode tap, the contribution due to deflection coils is 150 $(n_y/n_a)^2$. The ratio n_a/n_y is at least 4, hence the effective capacitance becomes 9 pF. This is quite small in comparison with 170 pF.

Therefore the capacitance component of eqn (14) is expressed by C_t. Thus, with some approximation, eqn (14) may be written

$$LC = C_t L_t \left(\frac{n_a}{n_y}\right)^2 = \left(\frac{\tau_f}{\pi}\right)^2$$

Therefore $$\frac{\tau_f}{\tau'_f} = \frac{n'_y}{n_y}$$

From eqn (11) $$\frac{n'_y}{n_y} = \frac{f'}{f}$$

Therefore the requirement of eqn (13) is satisfied.

5.3.4. Number of secondary turns n_y

The optimum secondary inductance L_s of the line output transformer is quoted as†

$$L_s = \frac{k_c^2}{\sqrt{1-k_c^2}} L_y$$

where k_c is the coefficient of coupling between windings n_y and n_a. In practice the ratio is limited to between 6 and 10.

In the constant-ratio design deflection coils are switched from n_y turns to n'_y. This causes a change of inductance L in the ratio

$$\left(\frac{n'_y}{n_y}\right)^2 = \left(\frac{f'}{f}\right)^2 = 2{\cdot}38$$

A compromise must be reached by aiming at below the optimum inductance on the 405-line system and above the optimum on the 625-line system. Since the former system requires less power, some loss of efficiency can be tolerated. However, on the 625-line system the number of turns will be normally in excess of that used in a transformer designed for operation on the 625-line system alone. Apart from cost and bulk, increased number of turns may bring some difficulty with ringing at the beginning of the scan and possibly with third harmonic tuning (or at least to call for third harmonic tuning).

5.3.5. Magnetic circuit

As quoted before, the peak flux density in the core is

$$B_{(pk)} = B_0 + B_{max}$$

Normally in a transformer B_0 is close in magnitude to B_{max}. If they can be made equal on the 405-line system by design, then on switching to the 625-line system:

$$B_{max} \text{ will be reduced by } \frac{n_y}{n'_y} = \frac{f}{f'}$$

and $$B_0 \text{ will increase by } \frac{I'}{I}$$

In practice $I'/I < f'/f$. Therefore the peak flux density on the 405-line system can be higher than on the 625-line system.

5.3.6. Adjustment of third harmonic tuning

It has been shown already that switching of the transformer changes also the flyback time in the required ratio. Desirable as this action is, unfortunately it seriously upsets the third harmonic tuning.

There are several methods of effecting the coupling between the main portion of the transformer and the overwind.

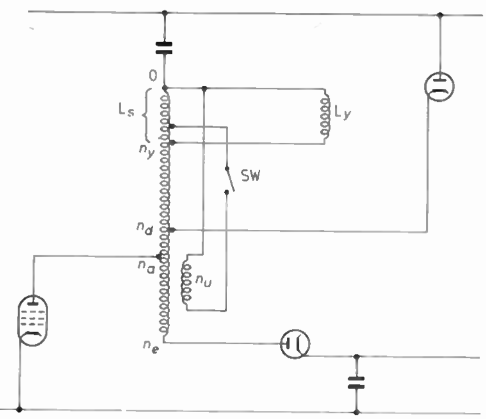


Fig. 5. Circuit diagram of third harmonic switching.

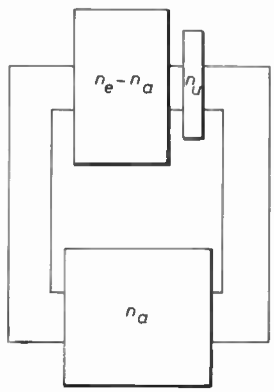


Fig. 6. Placement of coils for third harmonic tuning.

A method which is particularly adaptable to switching‡ is illustrated by circuit diagram in Fig. 5 and by a sketch in Fig. 6. An auxiliary coil, wound in the same sense as the e.h.t. overwind coil, is placed close to the e.h.t. overwind. It is connected across the primary of the transformer L_s. The number of turns n_u on the auxiliary coil and the number of turns on the transformer across which the coil is connected are made to be equal.

The e.h.t. overwind is designed to provide the third harmonic tuning on the 405-line system. On switching to the 625-line system the auxiliary coil is brought

† E. Jones, "Scanning and e.h.t. circuits for wide-angle picture tubes", *J. Brit.I.R.E.*, 12, pp. 31–48, January 1952 (App. III).

‡ J. B. Hughes and K. E. Martin, "Line Time-bases", paper 3 of "Circuit Design for Transistorized Television Receivers", (Mullard, London, 1960).

into the circuit. Its presence increases coupling and thus reduces the amount of leakage inductance. The amount of reduction required to bring back the third harmonic tuning can be controlled in the design by the number of turns n_u and the spacing between the auxiliary coil and the e.h.t. overwind.

With the amount of inductances normally involved a small difference in number of turns between the auxiliary coil and the tap on the transformer primary does not produce any noticeable loading on the transformer. It is necessary to stress that there exists a large voltage difference between the auxiliary coil and the e.h.t. overwind, and therefore a suitable insulation should be provided.

6. Adjustment of Scanning Waveform

6.1. *S-correction*

The scanning current waveform required by modern cathode ray tubes for linear scan flattens out at the extremities of scan. A close approximation to the desired waveform can be obtained by connexion of a capacitor C_s of a suitable value in series with the deflection coils. The resulting current is a segment of a sinusoid of angular frequency $k\omega$ such that

$$\frac{1}{k^2\omega^2} = C_s L_y$$

On switching from one system to another it is necessary to preserve the proportions of the sinusoid. This can be produced by switching the s-correction capacitors in the ratio

$$\frac{C_s}{C_s'} = \left(\frac{k\omega'}{k\omega}\right)^2 = \left(\frac{f'}{f}\right)^2 = 2{\cdot}38$$

6.2. *Linearity correction*

Linearity correction primarily compensates for the resistive losses of the scanning circuit, including the booster diode. For the same circuit and components, it is reasonable to assume that any compensation made on one system should hold for the other system. Linearity sleeve and saturated inductor type of controls have been successfully tried in the laboratory.

6.3. *Amplitude adjustment*

As yet there is no standard approach to this problem and for this reason not much work has been done in this direction. Since normally only a small amount of correction is required, primarily to cover circuit tolerances, any type of control can be used. Resulting discrepancies should be tolerable, particularly in view of the 4 : 5 aspect ratio of the cathode-ray tube.

7. Control Grid Waveform

This is of particular importance for stabilized line time-base operation where it is necessary to ensure that the booster diode conducts till the end of scan under all circumstances. Significant saving in the average anode current of a pentode operated below the "knee" can also be achieved by a suitable selection of the time constant. A switch would be required to provide optimum conditions for each line system.

Consideration of previous sections showed that for both modes of operation the amount of power taken from the h.t. line is not the same. For this reason it may be sufficient to adjust the line time-base for optimum on the line system requiring more power and allowing the other system to take its own course.

8. Switching

An ideal dual-standard line time-base would not require any switching. In fact, such a design is possible, if one is prepared to accept a certain amount of inconvenience. The design would follow the constant flyback time principle. Anode dissipation of the line output valve would be very high and a special valve with about 50% higher limit than is at present generally available would be required. Steps should be taken to stabilize the supply to the first anode of the cathode-ray tube. Unfortunately, s-correction in such a time-base would be only a compromise between the two line systems.

An introduction of one switch can reduce the anode dissipation of the line output valve to more reasonable proportions. Another switch, if available, could be used to provide individual s-correction for each line system. To make the line time-base fully switchable requires only one more switch—to adjust a_1 potential for the c.r.t. These switches are dealing with one function at a time, hence they can be of the single-throw variety.

Switching of the constant ratio line time-base is equally simple. Apart from the necessity to provide an additional tap on the line output transformer, the functions to be switched are: s-correction, deflection coils tap and the third harmonic tuning. By a careful adjustment of component values, it is possible to provide the necessary switching by means of two change-over switches.

Thus while switching undoubtedly is a nuisance, from the line time-base point of view there is every justification why it should be used. On the whole, the effort put into switching is well repaid in terms of line time-base operation.

The precise placing of the switches and their ratings will depend on the particular situation. It may be worth mentioning that with the slide type of switches it is easy to obtain a good voltage rating between the contacts and chassis. This may be preferable to re-adjustment of the circuit so that the switches are at earth potential.

In the past there have been difficulties experienced in production from line time-base radiation. These have been overcome largely by designing the line output transformer for as good coupling factor as possible, good physical separation between the tuner and the line time-bases, careful lay-out of the receiver, screening, chokes, etc. It is reasonable to expect that in a receiver covering a wider frequency range, the difficulties from this source may increase. Therefore one should revert to old cures, perhaps with an added degree of care.

9. VA Requirements

9.1. *Deflection Coils*

This is given by

$$\text{VA} = L_y \frac{i}{\tau_s} i = L_y i^2 \frac{f}{1-p}$$

9.2. *Valve* VA

$$\text{VA} = i_a(V_{ht} + V_c - V_{ae}) + i_d V_c$$

Peak anode voltage with respect to chassis:

$$v_{a(\text{pk})} = (V_{ht} + V_c - V_{ae})(1 + F_p) + V_{ae}$$

Peak cathode voltage (diode) with respect to chassis:

$$v_{d(\text{pk})} = V_c(1 + F_p) + V_{ht}$$

Neglecting V_{ae} in comparison with $v_{a(\text{pk})}$ and V_{ht} in comparison with $v_{d(\text{pk})}$

$$\text{VA} = i_a \frac{v_{a(\text{pk})}}{1+F_p} + i_d \frac{v_{d(\text{pk})}}{1+F_p}$$

Let $$\frac{n_a}{n_y} i_a = i_1$$

and $$\frac{n_d}{n_y} i_d = i_2$$

so that $$i_1 + i_2 = i$$

then $$\frac{i_2}{i_1} = \sqrt{\eta}$$

where η—conversion efficiency. Therefore

$$i_d = i_a \frac{n_a}{n_d}\frac{i_2}{i_1} = i_a \frac{n_a}{n_d}\sqrt{\eta}$$

Since $$\frac{n_a}{n_d} \simeq \frac{v_{a(\text{pk})}}{v_{d(\text{pk})}}$$

VA contributed by valves is

$$\text{VA} = i_a \frac{v_{a(\text{pk})}}{1+F_p}(1+\sqrt{\eta})$$

Equating this with VA requirement of the deflection coils:

$$L_y i^2 \frac{f}{1-p} = i_a \frac{v_{a(\text{pk})}}{1+F_p}(1+\sqrt{\eta})$$

It is usual to specify sensitivity of deflection coils in terms of peak magnetic energy:

$$E = \tfrac{1}{8} L i^2$$

Substituting in the previous equation and rearranging:

$$E = \tfrac{1}{8} L_y i^2 = \frac{1}{8}\frac{i_a v_{a(\text{pk})}}{f(1+F_p)}(1+\sqrt{\eta})(1-p)$$

Since $$1 + F_p = 1 + \frac{2}{\pi} + \frac{\pi}{2}\frac{(1-p)}{p} \simeq \frac{\pi}{2}\frac{(1-p)}{p}$$

therefore $$E \simeq i_a \frac{v_{a(\text{pk})}}{4\pi}\tau_f(1+\sqrt{\eta})$$

When third harmonic tuning is used, $v_{a(\text{pk})}$ is reduced by about 15%.

In practice it will be found that the conversion efficiency η is somewhat higher for the 405-line system than for the 625-line system.

Thus the peak energy delivered by the line output valve is approximately proportional to τ_f. This implies that for the same flyback time τ_f, a time-base operating on the 405-line system needs the same amount of peak energy as a line time-base operating on the 625-line system. Therefore the conversion from one system to another does not automatically imply an increase in the peak energy requirement.

9.3. *H.T. Power*

However, the average current taken from the h.t. line increases with frequency as will be shown by the following argument.

Power delivered at the end of scan $P_1 = \tfrac{1}{2} L_y i_1^2 f$

Power recovered after flyback $P_2 = P_1 - P_l$

The efficiency of the circuit is

$$\eta = \frac{P_1 - P_l}{P_1}$$

from which

$$P_l = P_1(1-\eta) = \tfrac{1}{2} L_y i_1^2 f(1-\eta)$$

Since $$i_1 = i\frac{i_1}{i_1 + i_2} = \frac{i}{1+\sqrt{\eta}}$$

Substituting $$P_l = \tfrac{1}{2} L f i^2 \frac{1-\sqrt{\eta}}{1+\sqrt{\eta}}$$

also $$\tfrac{1}{2} L i^2 = 4E$$

$$P = 4Ef\frac{1-\sqrt{\eta}}{1+\sqrt{\eta}}$$

Since from eqn (7)

$$I_a = \frac{P_l}{V_{ht} - V_{ae}}$$

Substituting for P_l

$$I_a = \frac{4Ef}{(V_{ht} - V_{ae})} \frac{(1-\sqrt{\eta})}{(1+\sqrt{\eta})}$$

In practice frequency independent losses, including the e.h.t. power, will increase the value of average current.

Therefore $$\frac{I_a'}{I_a} < \frac{f'}{f}$$

Thus the power taken from h.t. line is

$$P_{ht} \simeq V_{ht} \frac{4Ef}{(V_{ht} - V_{ae})} \frac{(1-\sqrt{\eta})}{(1+\sqrt{\eta})}$$

10. Practical Designs

The designs to be described in this Section show the switchable line time-bases, and also indicate that by using a larger flyback, they can be designed well within the limits of the existing range of valves. For the constant flyback time time-base 12·8 μs flyback time is used, while for the constant ratio design $p = 20\%$.

Deflection coils used for all the designs have the following characteristics:

Inductance at 1 kc/s = 4·76 mH

Resistance at 20° C = 5·68 ohms

Peak to peak current = 1·73 A
measured on AW47–90 at 16 kV and 39 cm scan.

In all the designs deflection coils are balanced in respect of the boost line. This arrangement helps to reduce time-base radiation on the medium wave bands, but its performance in respect of ringing at the beginning of scan is slightly inferior.

A sleeve linearity control is employed.

No provision is made for width control.

The s-correction capacitor for the operation on the 625-line system is left in the circuit all the time; on switching to the 405-line system an additional capacitor is connected in parallel.

Attention is drawn to the fact that h.t. values quoted are not the same for the 405-line system and the 625-line system. This was done deliberately in order to take into account inherent h.t. line regulation.

10.1. *Constant Flyback Design*

The circuit diagram in Fig. 7 shows a conventional configuration of the line time-base. Stabilization is used to reduce sensitivity of the time-base to mains voltage variations. This arrangement helps also to keep the heater voltage of the e.h.t. rectifier within the published data limits, and to reduce e.h.t. regulation.

In order to reduce anode dissipation of the line output valve on the 405-line system, a resistor with a suitable decoupling is switched in series with the booster diode.

Inequality of heater voltage of the e.h.t. rectifier between the 405-line system and 625-line system is equalized by means of a choke in series with the heater. The choke is small enough to be mounted inside the valve holder. Details of its construction are given in Fig. 8.

Using one switch (S3) stable voltage is supplied to the first anode of the cathode-ray tube and to the stabilization network. In this particular instance all the available voltage on the 405-line system is supplied to the tube, but the same switching configuration could be used when the voltage is higher than that.

It was found necessary to use the PY88, primarily to meet the average current rating for the 405-line system.

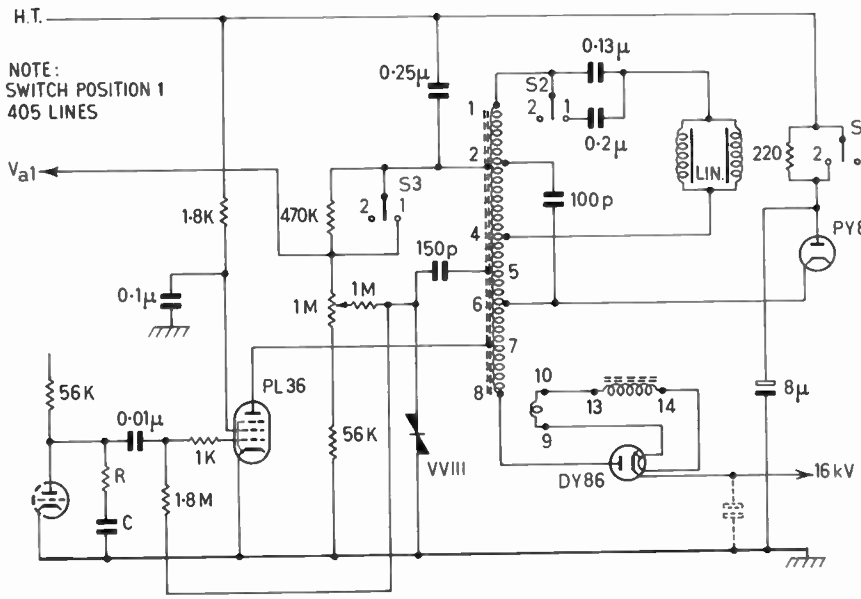


Fig. 7. 405/625 stabilized line time-base — constant flyback design.

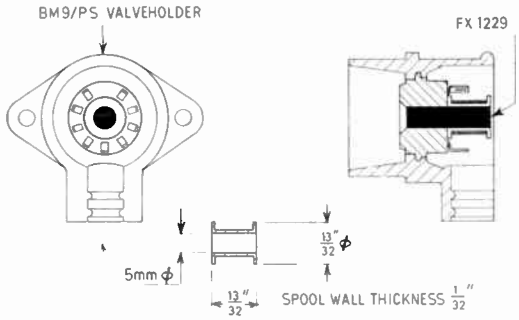


Fig. 8. DY86 mounting showing position of heater voltage compensating choke.

The results of measurements shown in Table 2 largely confirm the points discussed earlier in the paper. Of particular interest is the considerable increase in the average anode current on the 405-line system. Special effort was made to provide equal scan on both systems and this is reflected by increase in peak currents and e.h.t. on the 405-line system.

It must be admitted that the e.h.t. regulation is not satisfactory. A design was tried with a layer wound overwind with practically the same results. Using a split overwind, that is transferring one portion to the primary coil, it was possible to obtain e.h.t. regulation of 4 MΩ on the 625-line system. However, with this arrangement it was difficult to obtain a clean waveform of the anode voltage during scan.

Mechanical details of the original transformer are shown in Table 3.

10.2. *Constant Ratio Design*

Two designs are shown, stabilized in Fig. 9 and non-stabilized in Fig. 10. Apart from stabilization the circuit and the switching are identical.

The auxiliary coil (11–12) has been dimensioned so that it could be connected directly across the existing taps (1–3) on the transformer. With this arrangement one tap on the transformer is eliminated. The switching of the coil is interesting; one change-over switch S1 is used to change s-correction (405

Table 2
Electrical Measurements: Constant Flyback Time-base

		Television system 625 lines		405 lines		Published data limits
H.T. Line	(V)	220		215		
Beam Current	(μA)	0	300	0	300	
PL36						
v_a(pk)	(kV)	5·45	5·55	5·55	5·55	7
v_a(e.o.s.)†	(V)	68	63	94	73	
i_a(pk)	(mA)	235	293	250	315	320
i_{g2}(e.o.s.)†	(mA)	35	37	30	35	
I_a	(mA)	89	115	121	161	
I_{g2}	(mA)	23	25	22	24	
V_{g2}	(V)	180	177	174	170	
P_{g2}	(W)	4·15	4·42	3·82	4·07	5
P_a	(W)	4·55	5·44	8·54	8·68	9
PY88						
$v_{(h-k)}$(pk)	(kV)	4·55	4·55	4·55	4·55	6·6
i_k(pk)	(mA)	220	225	250	255	550
I_a	(mA)	89	115	121	161	220
Flyback time	(μs)	12·8	12·8	12·6	12·6	
V_{boost}	(V)	810	790	550	515	
DY86						
V_h	(V)	1·5	1·37	1·52	1·32	1·61–1·3
E.H.T.	(kV)	15·5	13·4	16·5	13·8	18
E.H.T. reg.	(MΩ)	7·7		9		
Peak Flux Density at 85°C (gauss)		2670		2930		3250
Wave-shaping Components						
R	(kΩ)	68		120		
C	(pF)	180		100		

†End of scan.

lines) and third harmonic tuning (625 lines). The other switch (S2) connects the deflection coils to the appropriate tap on the transformer.

With the constant ratio design there is no need for any adjustment of the heater voltage to the e.h.t. rectifier.

The boost voltage available from the stabilized line time-base is reasonably constant on the two systems and can be fed to the cathode-ray tube without further processing. The position is less satisfactory with the non-stabilized time-base. Since the boost

Table 3

Specification of Line Output Transformer: Constant Flyback Design

Core: 2 × FX2380.
Air-gap: 2 × 0·003 in.
Former for winding 1–7: S.R.B.P., i.d. 0·578 in., wall $\frac{1}{32}$ in.
Former for winding 7–8: S.R.B.P., i.d. 0·578 in., wall $\frac{1}{8}$ in.
Wire: Winding 1–7, Lewmex F, 30 s.w.g.
Winding 13–14, Lewmex F, 32 s.w.g.
Winding 7–8, enamelled copper d.s.c. 40 s.w.g.

Winding 1–7: 1070 turns, 70 turns/layer, tapped at 105, 210, 330 and 880 turns.
Winding 7–8: 1110 turns, wave wound, single wave, 4·1 mm wide.
Winding 9–10: 8 turns of suitably insulated wire.
Winding 13–14: 87 turns of 32 s.w.g. on a former $\frac{13}{32}$ in. long, i.d. 5 mm, $\frac{1}{32}$ in. wall thickness. Ferroxcube rod FX1229.

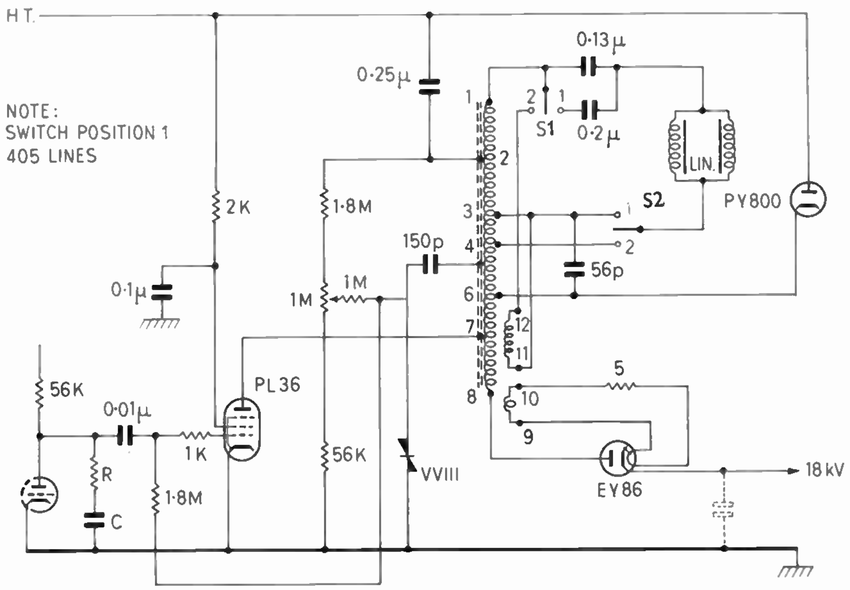


Fig. 9. 405/625 stabilized line time-base — constant ratio design.

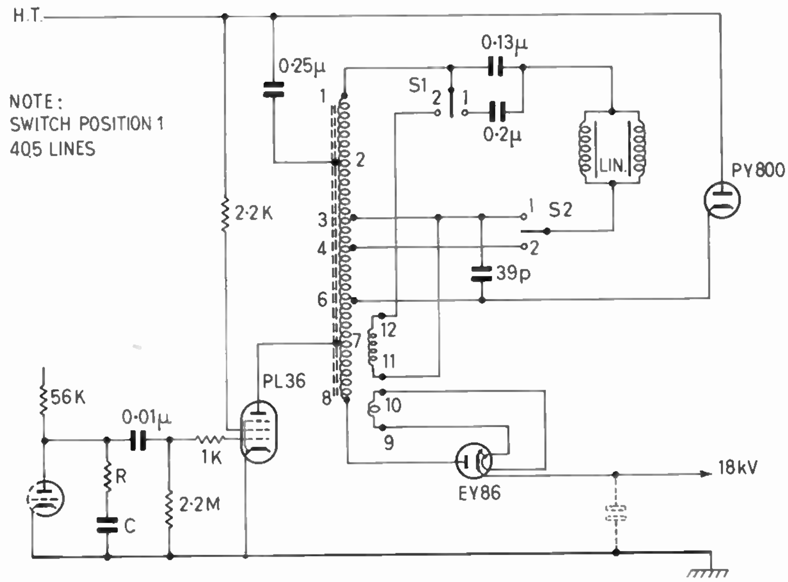


Fig. 10. 405/625 non-stabilized time-base—constant ratio design.

Table 4

Electrical Measurements: Non-stabilized Line Time-base, Constant Ratio Design

		Television system 625 lines		Television system 405 lines		Published data limits
H.T. Line	(V)	215		220		
Beam Current	(μA)	0	300	0	300	
PL36						
v_a(pk)	(kV)	5·7	5·5	5·9	6·0	7
v_a(e.o.s.)	(V)	26	33	24	27	
i_a(pk)	(mA)	245	287	176	220	
i_{g2}(e.o.s.)	(mA)	41	38	54	50	
I_a	(mA)	87	106	64	86	
I_{g2}	(mA)	23·8	21·9	31	28·5	
P_{g2}	(W)	2·8	2·8	3·1	3·0	5
P_a	(W)	1·7	2·6	1·15	1·73	
i_{g2}(r.m.s.)	(mA)	32·5	29·0	41·1	38·3	
PY800						
$v_{(h-k)}$(pk)	(kV)	4·75	4·65	4·95	5·05	5·75
i_k(pk)	(mA)	260	240	160	140	350
I_a	(mA)	87	107	64	86	150
Flyback time	(μs)	12·6		19·8		
V_{boost}	(V)	915	870	970	920	
EY86						
V_h	(V)	6·55	6·25	6·4	5·9	7·25–5·85
E.H.T.	(kV)	18·1	16·4	18	16·2	18
E.H.T. reg.	(MΩ)	6		6		
Peak Flux Density at 85°C (gauss)		2720		3300		3250
Wave-shaping Components						
R	(kΩ)	22		22		
C	(pF)	100		100		

voltage is too high, the necessary stabilization can be obtained using a voltage dependent resistor.

The electrical measurements on the time-bases are collected in Tables 4 and 6. Although the time-bases are designed for operation on 18 kV, the peak voltages and peak currents are well within the published data limits for the valves. The average currents on the 405-line system are about a half of the constant flyback design.

Mechanical details of the transformers are given in Tables 5 and 7.

11. Conclusions

It has been shown that the design of a switchable line time-base can be based either on the principle of constant flyback time or constant flyback-to-scan ratio. From a purely technical point of view, the latter mode of operation has much to offer. However,

Table 5

Specification of Line Output Transformer: Non-stabilized Line Time-base, Constant Ratio Design

Core: 2 × FX2380.
Air-gap: 2 × 0·003 in.
Former for winding 1–7: S.R.B.P., i.d. 0·578 in., wall 1/32 in.
Former for winding 7–8: S.R.B.P., i.d. 0·578 in., wall 1/8 in.
Wire: Lewmex F, 30 s.w.g. winding 1–7.
winding 11–12.
Lewmex F, 38 s.w.g. winding 7–8.
Winding 1–7: 1340 turns, 70 turns/layer, tapped at 125, 140, 250 and 1100.
Winding 7–8: 1500 turns, 55 turns/layer.
Winding 9–10: 4 turns of suitably insulated wire.
Winding 11–12: 140 turns, wound on a perspex bobbin 1/8 in. wide.
Interleaving: All windings interleaved with two layers of 0·002 in. paper.

Table 6

Electrical Measurements: Stabilized Line Time-base, Constant Ratio Design

		Television system 625 lines		405 lines		Published data limits
H.T. Line	(V)	215		220		
Beam Current	(μA)	0	300	0	300	
PL36						
v_a(pk)	(kV)	5·5	5·65	5·65	5·7	7
v_a(e.o.s.)	(V)	74	69	83	80	68
i_a(pk)	(mA)	240	286	171	218	290
i_{g2}(e.o.s.)	(mA)	40	44	34	37	
I_a	(mA)	103	132	83	114	
I_{g2}	(mA)	25·1	27·4	24·6	26·2	
V_{g2}	(V)	165	162	171	167	
P_{g2}	(W)	4·14	4·44	4·18	4·38	5
P_a	(W)	5·72	6·9	5·2	6·82	
PY800						
$v_{(h-k)}$(pk)	(kV)	4·75	4·85	4·85	4·9	5·75
i_k(pk)	(mA)	260	260	180	220	350
I_a	(mA)	103	133	84	114	150
Flyback time	(μs)	12·8		19·4		
V_{boost}	(V)	900	885	885	865	
EY86						
V_h	(V)	6·55	6·45	6·32	6·0	7·25–5·85
E.H.T.	(kV)	18	17	17·9	16·2	18
E.H.T. reg.	(MΩ)	3·3		5·7		
Wave-shaping Components						
R	(kΩ)	120		120		
C	(pF)	270		1200		
Peak Flux Density at 85°C (gauss)		2760		3250		3250

Table 7

Specification of Line Output Transformer
Stabilized Line Time-base, Constant Ratio Design

Core: 2 × FX2380.
Air-gap: 2 × 0·003 in.

Former for winding 1–7: S.R.B.P., i.d. 0·578 in., wall $\frac{1}{32}$ in.
Former for winding 7–8: S.R.B.P., i.d. 0·578 in., wall $\frac{1}{8}$ in.
Wire: Lewmex F, 30 s.w.g., winding 1–7.
winding 11–12.
Lewmex F, 38 s.w.g., winding 7–8.
Winding 1–7: 1300 turns, 70 turns/layer, tapped at 125, 155, 250, 370 and 1100.
Winding 7–8: 1480 turns, 50 turns/layer.
Winding 9–10: 4 turns of suitably insulated wire.
Winding 11–12: 155 turns, wound on a perspex bobbin $\frac{1}{8}$ in. wide.
Interleaving: All windings interleaved with two layers of 0·002 in. paper.

transformer switching must be used. This should not be an unsurmountable obstacle and no doubt the inherent inventiveness of the production staff will find a satisfactory solution.

It may be worth stressing the fact that the peak VA is approximately proportional to the flyback time. Thus, it may be advantageous to consider using as long flyback time as possible.

During the design stage of the receiver, and especially when switching is planned, some attention should be given to the possibility of interference from the line time-base. It is reasonable to expect that with planned extension of frequency range (Band 4 and 5) this problem may assume some importance.

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Radio Engineering Overseas . . .

The following abstracts are taken from Commonwealth, European and Asian journals received by the Institution's Library. Abstracts of papers published in American journals are not included because they are available in many other publications. Members who wish to consult any of the papers quoted should apply to the Librarian, giving full bibliographical details, i.e. title, author, journal and date, of the paper required. All papers are in the language of the country of origin of the journal unless otherwise stated. Translations cannot be supplied. Information on translating services will be found in the Institution publication "Library Services and Technical Information".

MEASUREMENT OF THE TRANSIENT RESPONSE OF TELEVISION EQUIPMENT

Two methods of measuring the transient response of television equipment are compared in a German paper. The measurement with a 250 kc/s square wave with a rise time of 100 mμs and the measurement with a "pulse and step function". Systems distortions are produced in the first case by the restricted bandwidth. The second method avoids this disadvantage, but it requires a large amount of measuring equipment and computation. It is shown that this outlay can be avoided by additional measurements of the amplitude and phase response.

"Measurement and tolerancing of the transient response of television equipment", H. Springer. *Nachrichtentechnische Zeitschrift*, **15**, No. 2, pp. 57–62, February 1962.

BINARY COUNTING CIRCUIT USING CORES AND DIODES

The logical design of an experimental model of a non-destructive and reversible two-digit binary counting circuit using simple toroidal cores and diodes is described in a paper in a German journal. The ferrite toroidal cores have the advantage that there is no upper limit of the applied magnetic field strength to switch a core from one state to the other. This makes the determination of the numbers of turns less critical. By using only passive elements this circuit can be considered as a preliminary step towards an all-magnetic pulse sequential circuit.

"A non-destructive and reversible binary counting circuit using ferrite cores and diodes", H. B. Liem and M. J. O. Strutt. *Archiv der Elektrischen Ubertragung*, **16**, No. 4, pp. 189–92, April 1962. (In English.)

ANALYSIS OF TRANSDUCERS

Electromechanical and electro-acoustic transducers used by communications engineers have commonly been analysed for steady state response by the device of an equivalent linear impedance intended to represent the reaction of the mechanical or acoustic forces and velocities on the electrical portion of the transducer. This procedure can overlook effects of importance. Two methods for deriving exact differential equations for electro-mechanical systems are described in an Australian paper and are based on (i) conservation of energy during an arbitrary displacement or (ii) Hamilton's principle. As an example differential equations of an electrostatic tweeter driven from a source of finite impedance are derived. Both these equations turn out to be non-linear and it is shown that to linearize them into agreement with conventional analyses using analogous linear circuits, requires a number of assumptions, of which at least one is very dubious indeed. The paper concludes that, for examination of such transducers, dynamic circuit theory provides a tool which is a good deal more powerful than conventional schemes based on the assumption that equivalent linear electric circuit elements can adequately represent the mechanical or acoustic reaction on the electric circuit.

"Dynamic circuit theory for use in communications" R. M. Huey. *Proceedings of the Institution of Radio Engineers Australia*, **23**, No. 2, pp. 69–77, February 1962.

DIRECTION METER FOR STEREOPHONY

With a very simple circuit it is possible to make more effective use of the "logarithmic" programme meters when transmitting stereophonic programmes, according to a Dutch engineer. The overload limit of both channels is continuously monitored on one meter, while the difference in level between the two channels is indicated on the other. The latter meter is directly calibrated in decibels, thus giving an immediate indication of direction in "intensity" stereophony. The unambiguous and precise indication of the centre is perhaps its most valuable feature, although other characteristics of the sound pick-up may be measured with this device.

"A direction-meter for stereophony", J. J. Geluk. *European Broadcasting Union Review*, No. 72-A (Technical), pp. 50–56, April 1962.

NOISE IN BACKWARD WAVE TUBES

Long beam tubes and more particularly "O" type carcinotrons exhibit spurious modulation effects in the generated u.h.f. signal. A French engineer has examined, on experimental tubes, the various effects which may be at the root of the spurious modulation. A first effect corresponds to modulation frequencies of about 1 to 3 Mc/s and is observed at the line input. It is the result of coupling between the primary beam and the secondary electrons generated at the line input and moving in a direction opposite to that of the primary beam. A second spurious effect corresponds to a modulation frequency of 0·2 to 1 Mc/s and appears as a plasma oscillation in the beam, this oscillation being excited by the first effect mentioned. Some technological modifications are indicated for removing spurious modulation.

"Contribution of the investigation of spurious noise phenomena in carcinotron 'O'," G. Biguenet. *Annales de Radioéléctricité*, No. 68, pp. 100–18, April 1962.