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EUROPEAN UNITY OF ENGINEERS

PLANS for the economic integration of Europe centre largely on manufacturing problems, and the future role of the British engineer demands closer association with his European confrères. Since 'A European Common Market' was the subject of the editorial in the Institution's April 1957 *Journal*, some progress has been made in promoting unification of engineering standards. But there is a need for clear understanding of national manufacturing needs and problems if European unity is, in fact, to benefit every participating country. Certainly an increasing number of qualified engineers is required in order to achieve further prosperity in Europe. Moreover, the skill and ability of those engineers must be recognized beyond national borders of education and training.

It is therefore significant that for the first time the Federation Europeenne d'Associations Nationales d'Ingenieurs (F.E.A.N.I.) has met in London. Since its formation in 1951, F.E.A.N.I. has attracted a membership of over half a million professional engineers drawn from eighteen countries of Western Europe. Some 148,000 of these members come from Great Britain, but this number is rapidly increasing through the registration of engineers with the Council of Engineering Institutions which arranges British participation in F.E.A.N.I. activities.

The Federation is consulted by the Council of Europe and has carried out work for the Organization for Economic Co-operation and Development (O.E.C.D.) and for the European Economic Community in Brussels (E.E.C.). F.E.A.N.I. also published in 1965 the first European Register of the higher technical professions. The Federation also co-operates with the International Labour Office and U.N.E.S.C.O.

F.E.A.N.I. has already assumed the role of European convenor of International Conferences of Engineering Societies. Four such Conferences have already been held, and at the October London meeting, preparations were made for the Fifth International Congress to take place in Athens from the 7th to 12th May 1967 with the active participation of O.E.C.D. The general theme will be 'The Engineer and the Economic and Social Progress of Nations'. This emphasizes awareness that the development of Europe must be based on the progress of technology.

Great Britain has as much to contribute to European prosperity as any other European nation. The individual British Engineer can only make his contribution through the Council of Engineering Institutions which is the national British body associated with F.E.A.N.I. How far the various branches of engineering will be reflected in F.E.A.N.I. activities will depend upon the contributions of individual Institutions.

Possibly in no other professional field, with the exception of finance and economics, is European unity so dependent upon the ability of engineers and manufacturing organizations. Most of the Chartered Engineering Institutions, I.E.R.E. included, are well placed to play a positive part in F.E.A.N.I. and European conventions because of a long-standing membership drawn from European countries.

The proposal to create a European Engineers' Register will alone demand co-operation and understanding in order to avoid complicated registration formalities which can be both discouraging to the individual and harmful to organizations needing the services of the specialist engineer. European unity need not mean, either in individuals or in manufacturing techniques, nondescript conformity. Perhaps in Athens in May 1967 engineers will demonstrate their ability to create both unity and opportunity for the development of new ideas unfettered by the harsh dictum of unrelenting regulations or treaties.

G. D. C.

INSTITUTION NOTICES

Annual General Meeting of the Institution

The Annual General Meeting of the Institution, at which the Annual Report of the Council and the Accounts for 1965–66 will be presented, and officers and Council for 1967 elected, will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, on Thursday, 15th December 1966. A formal Notice and Agenda was published in the July–August issue of the *Proceedings*.

Members not resident in Great Britain who do not at present subscribe to the *Proceedings* may obtain a copy of the Annual Report free of charge on application to the Secretary.

At the conclusion of the formal business of the Annual General Meeting, the President-elect, Professor Emrys Williams, will deliver his Inaugural Address. The Address will be printed in the January 1967 issue of *The Radio and Electronic Engineer*.

The Norman W. V. Hayes Memorial Medal

The I.E.R.E. acts in alternate years with the Institute of Electrical and Electronics Engineers of America as adjudicators for the award of the Norman W. V. Hayes Memorial Medal of the Institution of Radio and Electronics Engineers, Australia. This Medal is presented each year to the author of the outstanding paper published in the Australian Institution's *Proceedings*. The I.E.R.E. recommendation for the award of the Medal for 1966 is that it should be presented to Professor A. R. Billings, C.Eng., for his paper on ' Applications of Delay Line Networks having Time Dependent Impulse Response', which was published in the *Proceedings of the I.R.E.E Australia* for April 1965.

Professor Billings is at present Dean of the Faculty of Engineering at the University of Western Australia, where he has occupied the Chair of Electrical Engineering since 1959. He was a lecturer in electrical engineering at the University of Bristol from 1952 to 1959.

U.K.A.C. Annual Lecture

This year's U.K. Automation Conference Lecture will be given by Professor J. R. N. Stone, C.B.E., of the Department of Applied Economics, University of Cambridge, on Tuesday, 6th December, 5.30 p.m. at the Institution of Electrical Engineers. Professor Stone has chosen for his subject 'Our Unstable Economy—Can Planning Succeed?'

Tickets may be obtained from the Honorary Secretary, United Kingdom Automation Council, c/o The Institution of Electrical Engineers, Savoy Place, London, W.C.2.

Award of Royal Medals

Her Majesty the Queen has been graciously pleased to approve recommendations made by the Council of the Royal Society for the award of a Royal Medal for 1966 to Mr. J. A. Ratcliffe, C.B., C.B.E., F.R.S., C.Eng. (Member), formerly director of the Science Research Council's Radio and Space Research Station, for his distinguished studies in the ionosphere and on the propagation of radio waves. Mr. Ratcliffe is President of the Institution of Electrical Engineers for the current year.

Symposium on Space Technology and Science

The Seventh International Symposium on Space Technology and Science will be held in Tokyo, at the Nippon Toshi Center, from 15th to 20th May 1967.

The Symposium is planned to cover the following fields:

Propellants and propulsion; vehicles (including materials, structures, dynamics, aerodynamics and astrodynamics); spacecraft (including manned and unmanned spacecraft, scientific and communications satellites and space probes); space electronics (including telemetry, tracking, space communication and instrumentation); guidance and control; systems engineering (including operation, reliability and ground support); space science; space medicine and biology; space law; national space programmes.

Intending contributors should send abstracts of their papers (about 200 words) by 1st February 1967, to: Professor Tamiya Nomura, Chairman of the Programme Sub-Committee, ISTS-Tokyo 1967, Institute of Space and Aeronautical Science, University of Tokyo, 856 Komaba-machi, Meguro-ku, Tokyo, Japan.

Applications for registration should be addressed to: Mr. Seiichiro Nozawa, General Secretariat of ISTS–Tokyo, 1967, Japanese Rocket Society, The Yomiuri Newspaper Building, 1, 3-chome, Ginza-Nishi, Chuo-ku, Tokyo, and should arrive before 10th March 1967.

Correction

The following corrections should be made to the paper 'Pulse Response of Delay Lines: II *m*-derived Delay Lines', published in the October issue of *The Radio and Electronic Engineer*.

Page 227, Equation (4), change c to C in the denominator;

Page 229, Equation (10), right-hand side, should have plus sign after 1, thus:

$$E_n/E_0 = 1 + K.\exp(X\alpha t) + \ldots$$

Page 229, Table 2, column 'L', fourth item from top should read: 2·00000.

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Some Design Aspects of an Industrial Telemetry and Telecontrol System

By

J. D. MARTIN, M.Sc.(Eng.), C.Eng.†

Summary: The overall design philosophy of an industrial telemetry and telecontrol system is approached in a general manner, and its leading characteristics are deduced. The constituent parts of such a system are then examined in some detail. Particular attention is paid to the practical communication aspects of the system, and to the logic circuits which could be employed in its construction. The enunciated principles are illustrated by reference to a particular commercial equipment.

1. Introduction

Telemetry and telecontrol involve measurement and control of plant from a distance. During the past few years there has been a large increase in the application of such techniques to distribution networks of all kinds. Electricity supply networks have been employing telemetry and telecontrol for some time, but the trend has now spread to railway systems and lately to the gas, water and oil distribution systems. Whilst telemetry used to be applied to single quantities, it is now fairly common for a control centre to have access to several hundred telemetered quantities which are all collected by a single integrated telemeter system. This system is usually bi-directional and can therefore convey a similar number of control instructions to the outlying stations.

Figure 1 shows the typical layout of a master station (control centre), and several out-stations. The out-stations may be interlocking areas for a railway system, pumping stations for a pipeline system, or substations in an electricity supply network. Telemetered quantities include pipeline pressure, quantity flow, supply voltage and current. These will be converted into equivalent electrical signals by suitable measuring transducers. There will also usually be two-state indications from the out-stations such as 'pump running', 'valve closed', 'points thrown', etc. Controls are mostly two-state such as 'start/stop', 'up/down', etc., although there are applications where a continuous drive is required, such as voltage regulation or operation of a pipeline valve. Distances between the stations may vary from one mile to several hundred miles, the only means of communication being the communication links L; and so the crux of the telemetry problem is one of transferring data over these links. Most links in this country are rented G.P.O. lines having a restricted frequency range, and their layout is dependent on local conditions rather than the desires of the system planner. (See, for example, L6 in Fig. 1 which would more conveniently follow route L7.) Certain, or all, of the links in a system may be v.h.f. or u.h.f. radio paths, if this is economically better or if G.P.O. lines are not available at the particular locations. The system designer is therefore constrained at the outset by the existing communication links and their complex layout, and his main task is to ensure that reliable communication is carried out over these sometimes indifferent circuits.

The smaller systems, ranging from perhaps one to twenty telemeters, employ many different operating principles. The most well-established of these use electro-mechanical components; for instance, the electricity supply industry's telemetry system.[1] Although analogue techniques can be used for these small systems, it becomes desirable to use digital representation of data in large systems. Electro-mechanical devices do not have the speed or life for the continuous operation required by a large system,

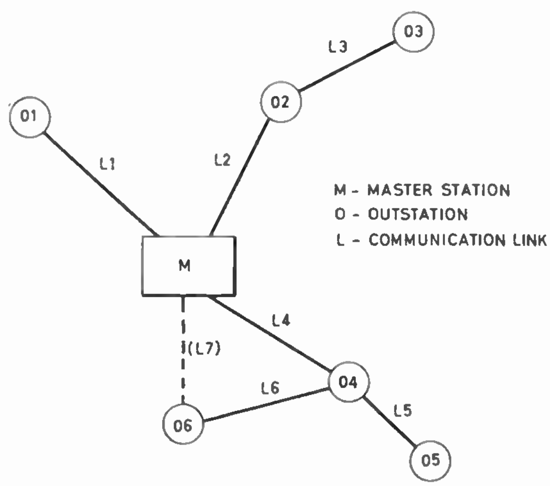

Fig. 1. Typical telemetry and telecontrol system.

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and so solid-state devices are widely used. Delanoy and Richardson have described a system employing cold-cathode tubes and transistors,[2] while the operation and use of typically large telemetry systems can also be found.[3–5] An appraisal of electro-mechanical and solid-state systems has been given by Coley.[7]

The object of this paper is to review the major issues involved in the design of large solid-state systems, and to indicate the reasons for certain preferences where a choice of technique exists. The discussion is illustrated by reference to the 'Westronic' Style F Remote Control and Indication System which was developed initially by the Westinghouse Brake & Signal Co. Ltd. for railway use, but has subsequently been widely used by electricity, water, gas and oil undertakings, as well as railway authorities. The Style F equipment is the latest in a succession of 'Westronic' equipments, the first of which was installed for railway traffic control in 1958–9.[6]

In the subsequent discussion, the term telemetry will be used to describe both telemetry and telecontrol functions.

2. System Specification

The production of a suitable telemetry system for a particular application involves close co-operation between the designer and the prospective user, since each approaches the problem from a different background, and with different ends in view. Although this is true of any electronic equipment, it is particularly so in this case owing to the large difference in outlook between the two parties.

The user's main concern is that the telemetry system should provide adequate supervision of the remote plant. All the control functions and telemeters should be available for use with a response time one order better than that of the actual plant. Since the operator is only concerned with plant operation, it is important to make the control system as unobtrusive as possible. In the control of voltage regulators and similar devices where the operator is controlling some remote apparatus and simultaneously watching the results of his action on a telemetered indication, it is most important that he retains the 'feel' of the process as if he were actually at the remote location. The electronic equipment should not interfere with the sequence of control operations, but should automatically deal with the data to be transmitted without any assistance from the plant operator.

Reliability in all senses is most important to the user, since the public utilities provide a twenty-four hour service to the community, and breakdowns of communication can be very disturbing or even dangerous. Consequently, the equipment must operate continuously for long periods without attention at all, and in the event of a rare breakdown the fault must be quickly found and put right. This fault-finding procedure is most important since it is probable that the searcher will be only partially skilled in electronics, and will not be very familiar with the procedure since faults occur infrequently. It is only possible permanently to employ an electronic technician for first-line maintenance where the installation is large or where other electronic equipment is in use. Fault-finding is thus often carried out by an ordinary electrician, or other maintenance personnel. Breakdowns are bad, but data inadvertently routed to the wrong location may be more embarrassing than a complete stoppage, since wrong data-transfer may not be detected. Data errors of this type due to circuit faults may be reduced to a minimum by careful circuit and equipment logic design. Particular attention has to be paid to the communication aspect of the system, because of data errors caused by external random interference. These cannot be entirely eliminated, but their probability of occurrence without detection can be reduced to an acceptably low value. It is desirable to regard a telemetry system as non-fail-safe, and to ensure safe operation by local interlocking, e.g. between pump motor and delivery pressure. In some cases this is not possible, e.g. electrical substation control.

The equipment designer usually has a limited knowledge of the particular problem of the user, and his main aim while complying with the functional requirements and required reliability of the equipment, is to construct equipment which has the utmost flexibility. Operational procedures vary from industry to industry, and even within one industry each installation is very much an individual problem since numbers of controls, and numbers and dispositions of out-stations all vary from scheme to scheme. For instance, a railway scheme may require the continuous display of several hundred items of information from several out-stations, while keeping the time between initiation and display of any indication to one second. Alternatively, a gas pipe-line grid may require any alarm condition to be registered within five seconds of initiation, a continuous display of conditions at one selected out-station, a typed log of all telemetered quantities every thirty minutes (without disturbing the alarm scanning), and the facility of continuous inching control of any of the pipe-line throttle valves in the system. The same basic hardware and system organization should be used for all types of application so that manufacturing and servicing may be carried out most efficiently.

The keynote of reliable, flexible equipment is simplicity, and the designer's major task becomes the production of an equipment based on simple circuits, with straightforward logical layout. Not only are faults less likely, but they can be located much more

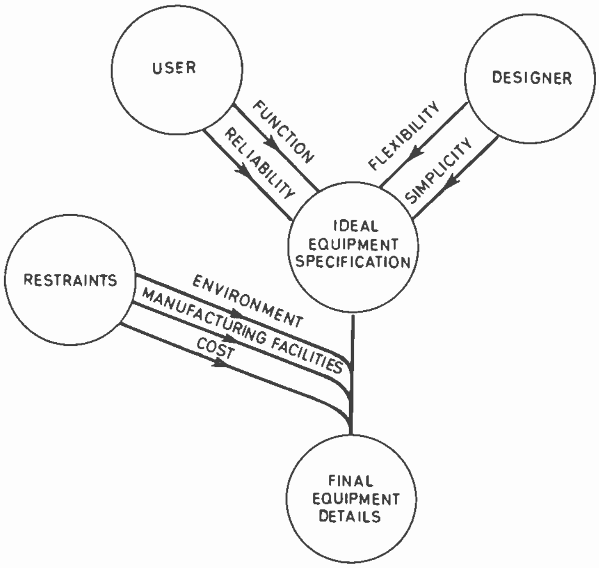


Fig. 2. Equipment design process.

quickly and new system arrangements can be evolved which will widen the scope of system applications without degrading its performance. A pictorial view of the above processes is shown in Fig. 2.

A telemetry system is often placed in a hostile environment which is far removed from the sheltered conditions in a development laboratory. Many equipments are placed in buildings which are occupied by plant personnel and so the temperature and humidity are controlled but when the remote plant is completely unattended, the telemetry equipment often faces wide extremes of temperature and humidity. Power supplies are often poorly regulated and sometimes shared with other equipment which feed back voltage pulses and spikes into the common supply. The associated relays, transducers, and indicators which are connected directly to the telemetry equipment often introduce large interference signals, and sometimes dangerously high voltages under fault conditions. The communication path normally introduces signal distortion which is liable to change through re-routing of the line and which can be considerable if fast working is required. Interfering signals are injected into the line circuit by sundry pieces of plant, and can result in wrong information being received.

3. System Analysis

Owing to the vast quantity of varied information that must be transferred between stations, a digital method of representing data is desirable. For small numbers of on/off data channels, a frequency-division-multiplex or tone-signalling system may be used as in v.f. telegraphy.[7] Transmitted data are shared between several out-stations, and so it becomes convenient to divide the data digits into groups or blocks. Figure 3(a) shows that in a simple system there might be m blocks each having n digits. Normally this process is repetitive, and so the system continuously cycles through mn data digits.

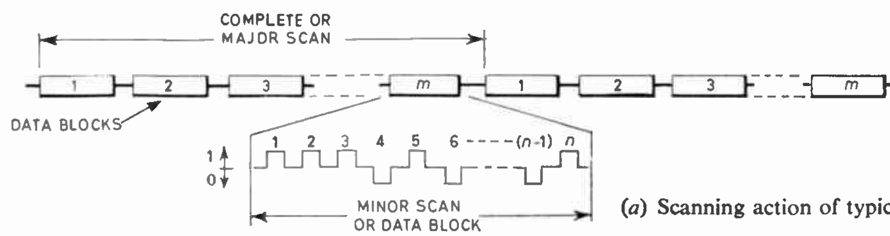
Figure 3(b) shows the block diagram of a typical continuously-scanning system. All major telemetry systems will approximate to this principle. The significance of not scanning through all available data digits will be considered later.

Starting from the communication link, the first item is the time-division-multiplex (t.d.m.) unit. In operation, this unit takes n parallel binary data inputs and converts them into an equivalent serial n-pulse binary train which is sent out over the link by means of the 'modem' (modulator/demodulator). The modulator converts the d.c. train of pulses into a suitable form of modulated carrier signal for transmission, and the demodulator performs the converse operation for carrier signals arriving from some other station. Incoming trains of n pulses are sorted and presented as n simultaneous output signals. Between any two stations therefore, the t.d.m. arrangement provides n data channels.

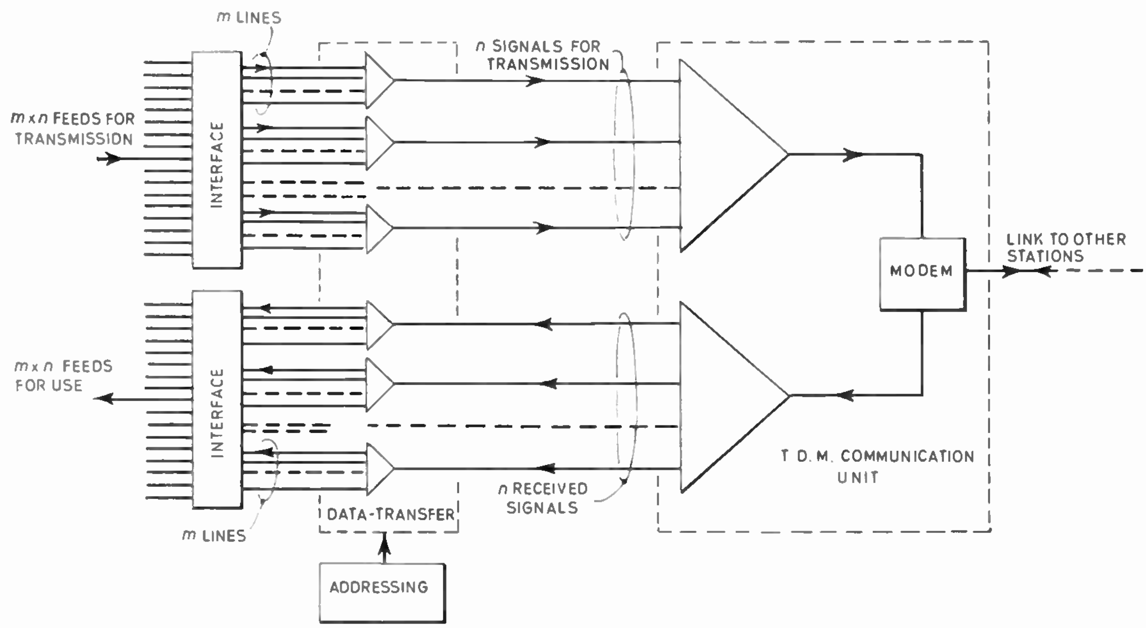
A t.d.m. terminal unit requires a source of timing pulses at the data pulse rate, and synchronizing signals which define the start and finish of each train of n pulses. The signals are generated continuously, irrespective of the data actually being transmitted at the time, and serve to maintain the various out-stations in time correspondence. Consequently, the design of the t.d.m. terminal unit is basic to the problem of communicating between the various stations.

Reference to Fig. 3(a) will show that successive groups of n pulses carry different information, and in general will represent data from different stations. A further stage of multiplexing is therefore required, which sorts data on a scan rather than a pulse time-scale. This function may be called data-transfer rather than t.d.m., since in a complex equipment it is not necessary that the different blocks of data will be presented in strict rotation. Incoming and outgoing data-transfer are separate and depend in detail on the sources and stores of data, as will be indicated later.

Control of data-transfer operations is vested in the addressing unit. Blocks of data travelling around the system must be identified to enable them to be routed to their correct destination. Identification is often carried out by means of an address tag consisting of a special binary code. Hence the generation of address codes and routing functions from them is of the utmost importance in transferring blocks of data from station to station.



(*a*) Scanning action of typical system.



(*b*) Simplified master station for *m* addresses.

Fig. 3.

The interface between the local telemetry station and the associated external plant is a vulnerable point for the entry of interference. The actual form of the interface is determined by the storage and input devices in use, as will subsequently become clear.

4. System Design

Although much of what follows is completely general and describes the problems facing any designer of telemetry and telecontrol equipment, some of the conclusions reached refer specifically to the Westronic Style F equipment.

4.1. *Overall System Operation*

Before a detailed analysis of the individual units can be commenced, the overall system must be considered and its macroscopic mode of operation decided.

Simplex operation of the communication channel (i.e. transmission in one direction only at any one time), is found to be optimum for a general system. Figure 4 shows the simplex operational sequences. Not only does this enable various items of equipment at the terminal stations to be shared between transmitting and receiving functions, but it also removes the fundamental restriction on 'go-and-return' signal delay which exists when simultaneous transmissions are used. More fundamentally still, an out-station must receive an incoming signal completely before its address can be decoded. Until the address code is received and decoded, the out-station cannot transmit, so the address acts as a call-sign, and the action is inherently simplex. Some radio circuits are simplex anyway, and even on line circuits it is often an advantage to be able to use the same frequency channel in both directions of transmission.

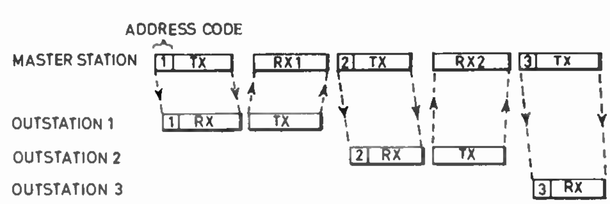


Fig. 4. Transmit/receive sequences for simplex working.

Simplex operation is consequently basic to this type of system. If the communication path can allow duplex traffic, then this may be provided by doubling part of the master station equipment and allocating the out-stations so that the master station will be simultaneously transmitting and receiving, but each out-station will be working on the normal receive/transmit sequence (Fig. 5). The arrangement effectively becomes two simplex systems sharing the same communication path but interlocked so that when one half of the master station is receiving the other half is transmitting. This is achieved in practice with very little extra equipment over the standard simplex arrangement, but with operation at twice the data rate.

Another decision is required as to whether the system should continuously scan around all out-stations and functions, or whether it should operate on an on-demand or 'single-shot' basis. Information for controls or for indication of plant operation tends to occur in a spasmodic fashion. Consequently it appears that an optimum telemetry system would be one which is available to transmit any new item of random information exactly when it occurs. The converse of this, which is the continual scanning and transmitting of all data sources, would seem to waste time before a change could be detected and transmitted. However, it is highly impractical for a master station to receive randomly occurring data from a number of out-stations, since there is the likelihood of two or more out-stations transmitting together. There are methods for ensuring that this case is catered for, but the overall system then tends to be rather sensitive to small faults and malfunctions. The best solution appears to require the master station and out-stations to be in continual contact. This is achieved by allowing the master station to transmit to each out-station in turn, and allowing that out-station only to reply. In this way, the integrity of the system is continually monitored, and an alarm can be raised in the event of a failure. The various stations are now permanently in time synchronism with one another. It is not necessary for data to be transmitted in this state, but the basis is laid for alarm or control information to be transmitted over the system when required, with only a short delay. Thus, independently of the data being transmitted, the t.d.m. communication units of master and out-stations are kept continually in synchronism and are always actively awaiting the insertion of data.

While considering the basic time relationship between master and out-stations, it is pertinent to introduce the idea of eavesdropping operation. Referring to Fig. 1, suppose that the telemeter indications from out-station O6 are required to be displayed at out-station O5. The sequence of actions

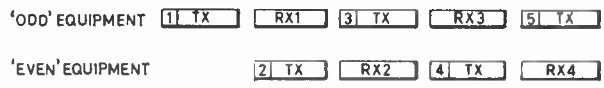


Fig. 5. Operation of duplicated master station to provide duplex working.

would be as follows. The master station transmits a calling sign to out-station O6. In the next block time interval, out-station O6 transmits its telemeter indications back to the master station, out-station O5 also receives this information, and consequently is said to eavesdrop on the exchange of data between O6 and M. Notice that in the normal sequence O5 would be either quiescent or transmitting at this part of the system time cycle (Fig. 6). In fact, the call-sign which instructs O6 to transmit also prepares O5 to receive extraordinary data on the next scan. Observe that in the system layout shown in Fig. 2, a transmission path must be provided between O5 and O6 over the links L5 and L6, which is not always automatically provided by radio links.

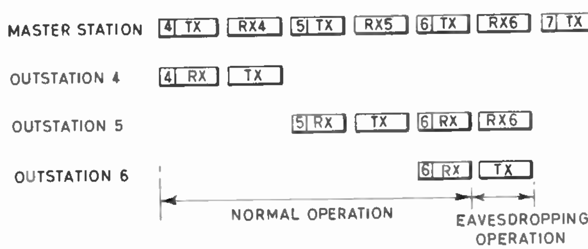


Fig. 6. Eavesdropping operation.

4.2. *T.D.M. Communication Unit*

There are two basic design parameters which must be initially settled. These are the data pulse-rate, and the number of pulses per block, n. There is no general method of allocating these, so the choice depends upon a detailed review of each particular system.

The most direct limitation on data pulse-rate is the bandwidth available on the communication link. A channel of bandwidth B Hz, can theoretically pass a pulse-train of $2B$ pulses/second. This limit can be approached in practice, but a more usual practical limit is B pulses/second, and even this depends upon the method of modulation which is used.

When approaching this limit severe distortion of the signal takes place, so that the received pulse is more of a raised cosine shape rather than the usual square form. The detector function is then to enable the original two-level signal to be recovered, and clearly involves logical decision-making. If an audio-frequency carrier is used, there may be only one or two cycles of carrier in the duration of each pulse, and the detection process is complicated still further. James has given a good general summary of the factors involved in pulse transmission.[8] Most industrial telemetry and telecontrol systems are mainly limited by other factors, but bandwidth is liable to be a major limitation, if a 3 kHz channel is available which must be shared with speech.

A high data pulse-rate also introduces other problems. The circuit is more susceptible to impulsive interference, and because many of the available line circuits have poor frequency characteristics, equalizing networks are often required which add to the overall cost and require readjustment if the line route is changed.

The major constraint imposed on industrial telemetry systems is time. A maximum limit is commonly imposed on the time interval between the initiation of an action and its fulfilment. Although this is an overall system parameter, it does have a first-order effect on the data pulse-rate. Initially, therefore, it seems that the pulse-rate needs to be as high as the channel bandwidth will allow, in order to achieve the fastest overall system response. However, there is a minimum time limit imposed on the length of each block of data by the input/output devices of the system which frequently are electromagnetic relays (the reasons for this are discussed in Sect. 4.3). If these relays are the P.O. 3000 pattern or similar, they may require 100 ms to release when driven from semiconductor circuits. Thus, incoming data must be stored for this period to allow the relays to operate satisfactorily. To avoid excessive buffer storage on electronic circuits, it is therefore convenient to keep the time length of a data block greater than 100 ms. Using a fixed block time means that the number of data pulses n is inversely proportional to the pulse-rate. Hence the advantage of working at a fast pulse-rate is largely destroyed, because the number of pulses in the block is increased and the t.d.m. unit becomes larger. A cost/speed compromise is therefore necessary. However, pulses extra to those representing data in each block may often be used for code checking, and hence result in more reliable data; system reliability can thus be increased at additional cost.

The nature of the transmitted data also affects the number of pulses in the block. Most schemes require the transmission of a number of telemeters each of which will usually consist of 12 binary digits (3 decades in binary-coded-decimal (b.c.d.). Unless the system is very large it is not economic to send two telemeter readings in one block, so that the number of different data blocks transmitted from a given out-station is usually equal to the number of telemeter readings. Suppose a station has 6 telemeters and 48 on/off indicators. Then each data block could conveniently be made up of 12 telemeter bits, plus 8 indication bits, plus as many check bits as are thought to be necessary. Each station in the system must transmit the same number of bits in each data block, and so the final figure for n must be determined after considering the needs of each station, and arranging the data blocks to give maximum utilization of the system capacity.

A further decision which is fundamental to the detailed design of the t.d.m. communication unit is the provision of time references. Consider an out-station receiving from the master station a pulse-train containing control data. Viewed in isolation, the pulse-train appears just like a random sequence of signals at the '0' and '1' levels (Fig. 7). Two operations are necessary before the original binary code can be recovered. First the individual bits must be detected, which requires a knowledge of the pulse-rate, and of its correct phasing; i.e. if the waveform is sampled by a series of timing pulses at the correct rate, these intervals should occur around the centre of each pulse. The second operation is to determine the start and finish of each block of pulses, which is essential in order to define the binary word so formed. Both these operations must be carried out reliably at all times, under adverse line and ambient conditions. Since they are so essential to the conveyance of information, the techniques used should be rugged, reliable and as simple as possible.

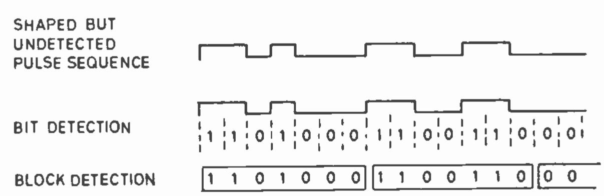


Fig. 7. Bit and block detection.

The crux of the problem may be stated thus. At the receiving end of any transmission, means must be provided for determining the transition between one bit of data and the next, and also for determining the ends of each block of data.

There are two basic methods of effecting bit timing. The first employs a local timing oscillator which is synchronized to the incoming data. This local oscillator must be stable and very close in frequency to the remote one. It must have the capability of phase adjustment to correspond to the incoming data, but it should ignore spurious and noisy signals when making this adjustment. This method is preferred for conventional data transmission where large quantities of data are transferred at each pass, and plenty of time is available for phase adjustment. In industrial telemetry data are transferred in small quantities allowing only a short time for phase adjustment, and the requirement of low cost for the relatively large number of out-stations often precludes the use of these advanced timing arrangements.

The second method of bit timing allows the timing information to be sent along with the data signal. Bandwidth limitations and non-linear delay characteristics preclude the use of a separate channel for timing information. The solution adopted with Westronic Style F employs bipolar 'return-to-zero' (r.z.) coding (Fig. 8). A frequency-shift (f.s.) type of modulation is used, with three allowed shifted frequencies, each corresponding to one of the signal levels. The r.z. signal will require at least twice the bandwidth of the more usual 'non-return-to-zero' (n.r.z.) signal, but the effect of this bandwidth redundancy is to reduce the error rate by improving the signal/noise ratio, and to give a simpler, more rugged system with only one timing source operating at a time. This source is therefore uncritical in its performance. When noise and general line interference are combined with the incoming signal, the overall effect is to distort the shape of the data pulses. Many pulses which would otherwise be lost when using fixed reference timing, can still be correctly detected by this system, since the time reference is determined by the incoming pulses themselves. The only necessary safeguard is the rejection of the extra-short pulses generated by noise.

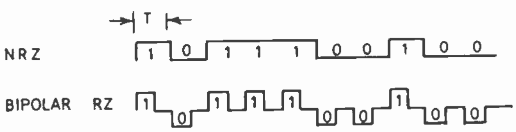


Fig. 8. N.r.z. and r.z. pulse sequences.

Information giving the start and finish of each block of data must be transmitted with the block itself. The start signal resets the receiving station and prepares it to accept the following data signal. The finish signal is not essential if all blocks have the same length, but it can be used in conjunction with a counter at the receiving station to check the number of received pulses in the block. This provides quite a powerful but simply operated check against certain classes of signal mutilation.

Start-finish data can be transmitted by a special character signal such as an extra-long pulse, or by a suitable binary code. The former is generally simpler to implement and is adopted in Westronic Style F. The basic length of this pulse is $2T$ where T is one complete pulse interval (Fig. 8). In practice, generous tolerances are allowed on the circuits which generate and detect this pulse, and the nominal transmitted pulse is somewhat greater than this. The start and finish pulses are of the same length, but of different sense. '0' denotes a start or reset and '1' denotes a finish or sync.

The t.d.m. communication unit may be constructed in several different ways, but the details of its operation depend upon a number of system variables, such as whether the location is a master or out-station, whether simplex or duplex operation is required, etc. By way of example, Fig. 9 shows the principle features of a typical t.d.m. communication unit for a master station working simplex. Three data bits only per block are shown, for simplicity.

Data presented for transmission are applied in parallel fashion to the upper set of gates, which are then sequentially interrogated by pulses from the clock-pulse distributor. The combined outputs form a sequential pulse train which is passed on to the modulator for transmission. Timing information as represented by clock, start and sync pulses is combined with the data signal before transmission.

Incoming signals are demodulated, and a parallel set of static outputs is obtained by applying the incoming data pulse train to the lower set of gates and sequentially interrogating these gates by outputs from the clock-pulse distributor. The clock pulses in this case are derived from the incoming signal, and routed to the distributor by the transmit/receive bistable (TR/RX). Other actions are necessary to check the incoming signal, and to insert checks into the outgoing signal.

4.3. *Data Transfer*

Data presented to and accepted from the t.d.m. communication unit are handled by data-transfer. Transfer is carried out under control of the addressing unit, which inserts and detects addresses as appropriate. The function of this unit is to take one group of n data bits out of an aggregate of p such groups say, and present this data-block for transmission. When an incoming group of n bits has been checked and passed as satisfactory, then this incoming data-block must be safely routed to a particular group of n stores, out of a possible number p. The action is thus basically that of an n-pole p-way switch, controlled as to position by the address. The precise physical form will depend upon the nature of the devices at the equipment interface (Fig. 3(b)).

Consider first the data stores which will be fed by the incoming line data to the station. While electronic stores may be used for driving an electric typewriter or a local compact display panel, it is desirable to use electro-mechanical relays as stores where the

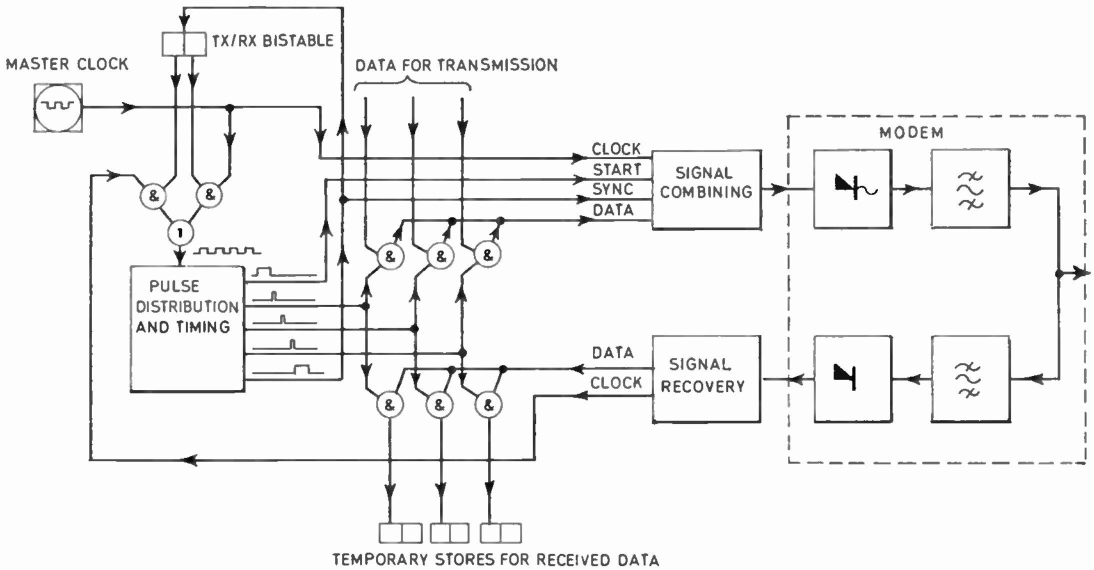


Fig. 9. Simplified operation of t.d.m. communication unit, for a 3-bit block, double-frequency-simplex master station.

outputs are interconnected in a complex manner or where they are required to operate a wide variety of voltage supplies. Relays have the outstanding advantage over semiconductor stores that several electrically isolated change-over contacts are available in each store. Because of the electrical isolation, large voltage surges on outside plant are less likely to interfere with the internal operation of the equipment. P.O. type 3000 relays are naturally the first choice in this matter, but reed relays are often used now that change-over contacts are more readily available. Reed relays will operate faster than the P.O. 3000 relay and consequently will allow the whole system to be speeded up (Sect. 4.2), but at present they still show a cost disadvantage.

In the Westronic Style F equipment, P.O. 3000 relays are the normal storage medium, although semiconductor and reed relay stores are used where their special properties are required. Figure 10 shows a typical small relay store. The storage relays are double-wound with 'hold' and 'operate' windings, and each group of relays has its two sets of windings controlled by a common relay, called a delivery relay. When data are ready for storage the delivery relay for the appropriate storage group operates, taking off the hold feeds and putting a supply on to the operate windings which are permanently connected to outputs of the t.d.m. unit. Each relay operates according to its own data line and when the delivery relay releases, these data remain held in the storage relays. Since storage and switching are performed by relays, this part of the equipment which is electrically very noisy can be kept well away from the sensitive electronic circuits. Mercury-wetted-contact relays are used for the delivery function, because of their long and trouble-free life. The action of these relays is essentially make-before-break, which prevents the storage relays from dropping out during the changeover interval.

Control of indication data from the external plant is in the form of a switch contact. The incoming connections are also very noisy and must be isolated or filtered before coming close to the electronic circuits. For Westronic Style F, R-C filters are used on all the incoming lines. In cases where dangerous

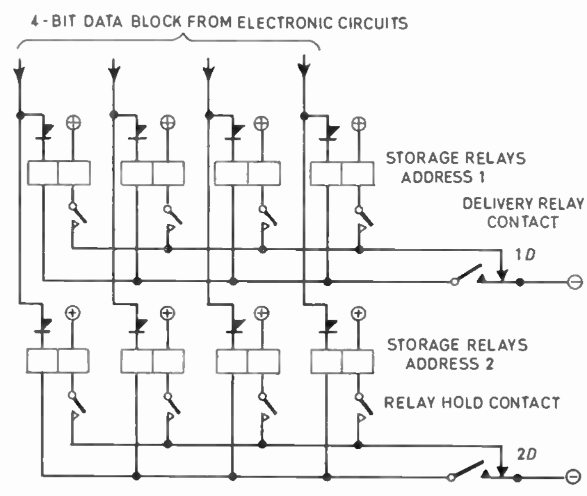


Fig. 10. Typical relay storage circuit for system having 2 addresses and 4 bits per block.

voltages might appear on these lines (e.g. 110 V d.c. in an electricity substation), isolating relays are sometimes used. The necessary switching action is very similar to that at the storage end of the system except that no hold feature is required, but it is performed by similar common relays, called entry relays.

4.4. *Addressing*

Generation of address codes controls the whole transfer of data throughout the system. The master station usually generates all addresses and thus retains complete control of all out-stations. In Westronic Style F, the master station generates an address, and then transmits it with any appropriate control data to the out-stations. All out-stations receive this transmission, but only the one that recognizes the address takes any action. This out-station transmits back to the master station the indication data and telemeter demanded by the received address, but does not transmit the address itself. The master station receives the data and stores it according to the address that was last transmitted. Since much more data flows from out-station to master station than vice-versa, this use of the address economizes on transmission capacity in the right way.

For sequential scanning through all functions, the addresses are generated by binary counters. In many cases, an arbitrary scanning sequence is required. For example in a large system of nearly identical out-stations, it may be desirable from the display aspect to gather data from only one station over quite a long period of time. Further, if data-logging on an electric typewriter is required, then the generation of addresses needs to be geared to the data requirements of the typewriter if excessive storage is to be avoided. It is quite easy to allow the basic scanning to progress sequentially, and allow special addresses to be inserted on demand from the data-log control equipment, or manually from the control desk. However, logging of telemeters or locking the display on to a single out-station must not prevent alarm conditions from being registered.

In Westronic Style F this is accomplished by interleaving address sources. Suppose that a continuous display is required from out-station 16, and that the data-log typewriter is simultaneously logging all out-stations. Then two address generators cater for these requirements, while two more scan through all out-stations looking for alarm conditions. The final address scanning sequence is then as shown in Table 1. Rapid scanning and indication of alarm conditions is therefore achieved while data-log and display requirements are dealt with independently. Reception of an alarm condition at the master station automatically causes the appropriate out-station to be scanned in more detail to ascertain the nature of the alarm.

Inching control is sometimes required for governors and regulators of various kinds. Halting the address generator on this particular address automatically provides continual two-way communication between master and out-station, with a time resolution limited only by the data block length itself.

4.5. *Control Encoding*

Control information is in general more important than indication or telemeter information. While errors in the latter can be very embarrassing, faults in the former could result in considerable damage to the controlled plant. In cases where there is little or no local interlocking between plant controls, the control data must be securely coded to prevent maloperation. Single equipment faults are usually covered by these extra precautions, but the main area of protection is in transmission faults.

A number of special codes can be used, applying the check at t.d.m. communication level, but these often further complicate the electronic portion of the equipment. A relatively simple arrangement describes all controls by decimal numbers,[4] so that to transmit 6 for instance, 6 '1' pulses are sent followed by 4 '0' pulses, and the total number of received pulses must be 10 if the code is to be accepted. Each decade of the original descriptive number is treated separately.

Table 1

Typical scanning sequence for combined logging, display and alarm functions

Transmission scan No.	Address generators: alarm scan (I)	data-log	alarm scan (II)	display
1	1			
2		1		
3			2	
4				16
5	3			
6		2		
7			4	
8				16
9	5			
10		3		
11			6	
12				16
13	7			
14		4		
15			8	

Westronic Style F does use simple parity checks at this level, but the main form of extra security is applied at the storage relay level, i.e. at the outgoing interface (Fig. 3(b)). While being simpler to implement, this approach also helps to protect against certain circuit faults. A typical example uses three bits to describe each two-state control. One is used as a common selection signal, and the other two indicate the nature of the two-state command. Suppose the three storage relays for these bits are A, B and C. Then A and B operated together will pick-up the interpose relay which performs the function, while A and C operated together will release the interpose relay. All other combinations of these three relays cause no action at all. Between control commands, A, B and C are left in their released condition, and the interpose relay latches itself.

Other coding arrangements which can be readily realized are those which use *m* bits out of *n*, e.g. the 2 out of 5 code. Suppose this binary code word is received on five relays A, B, C, D, E, then for a code to be accepted, only two relays must be operated. This check is easy to apply to relay circuits, and adds to the security of transmission. More elaborate arrangements are possible, e.g. a 3 out of 12 code which has been used for extreme security.

4.6. *Logic Circuits*

When considering the apparatus needed to realize the functions which have been discussed above, the keywords must be simplicity and reliability, as mentioned in Section 2. The circuits must be simple, well-tried versions of the many different kinds available. Their interconnection should result in predictable sequential working, not dependent upon critical timings or amplitudes. No *in situ* adjustments should be necessary, and preferably no adjustable devices should be used. Temperature and supply voltage ranges over which the circuits will operate satisfactorily should be greater than those required in the foreseeable future. Components must be carefully selected from established manufacturer's ranges, be conservatively rated, and for everyone's convenience should employ the smallest feasible number of different values and types.

Silicon transistors have been adopted for Westronic Style F, partly because of their great reliability and good ambient temperature range, but also because silicon transistor switching and logic circuits can be made to work with only one voltage supply.[9] Consequently the circuits are very tolerant to supply-voltage fluctuations, since all currents are derived from the same source instead of from two separate sources as in conventional switching circuits. As an example, circuits designed to operate from an 18 V supply with good safety factors, have worked successfully from a 5 V supply. Since the supply voltage is so uncritical, there is no necessity for an expensive electronically-stabilized power unit, and a conventional rectifier set is used for the electronic circuits as well as some of the ancillary items such as lamps and relays.

Large factors-of-safety are adopted in the design of the transistor circuits and worst-case design procedure is adopted throughout. The circuits are designed to operate over a temperature range of −30°C to +80°C, with full factors-of-safety in most cases.[9] This large safety margin ensures that the circuits are extremely tolerant to excessive component drift, or unforeseen ambient conditions. A good resistance to interference is obtained by this method and also by the low impedance level of all circuits.

Digital systems inevitably consist of a large number of basic circuits which are used as functional building blocks. One most important choice is that of the basic circuit block which will be flexible enough to economically produce all the desired system functions. Closely allied to circuit detail is the packaging problem of how large or small the basic mechanical unit should be, and how it should be secured in the equipment.

It is clear from broad discussions of the system, that extreme flexibility is required, not only according to a pre-determined range of types, but also in an unforeseen manner as determined by one particular customer's requirements. Ultimate flexibility is achieved by using a single transistor stage as the basic electronic and mechanical unit. There can be simple variations on this basic unit, such as a bistable assembly or a power-driver stage, but the fundamental advantage of being able to produce large quantities of these few basic units very cheaply, is retained. Construction of the system consists of fixing and wiring these basic blocks together. Replacement costs are minimal, since a faulty unit can be replaced and thrown away, thus avoiding the heavy burden of repair. A good example of this technique in use is described by Keeling and Kermack.[4]

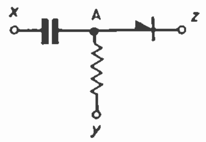
Although this approach has certain attractions, there are several disadvantages. A vast amount of inter-circuit wiring is required, and since this is done by hand its cost is high. The economic alternative is printed circuit wiring in large quantities, but any groups of circuits wired in this manner would have to be common to most systems in order to justify the initial cost.

When using individual circuit blocks, the large number of connections rules out the use of plugged blocks, and so fault-finding must proceed carefully and relatively slowly to enable the faulty block to be located first time. It is possible to group circuit blocks together to form functional units such as a 4-stage binary counter, and have the peripheral connections plugged. In the event of a failure now, the faulty unit can be located rapidly, replaced, and removed for inspection. If the failure symptoms are vague or misleading as they often are, then a unit can be replaced on suspicion, to see if it is faulty, and a firm decision reached much more quickly than when following a detailed search to determine the exact nature of the fault *in situ*. Since the searcher is often only semi-skilled in electronic techniques, this is an important aspect of system planning. Although industrial telemetry and telecontrol equipment usually has a mean-time-between-failures (m.t.b.f.) in excess of 6 months, and so easy fault-finding may not seem to be a vitally important design factor, yet the equipment is continuously performing a vital role in a complex plant or public service network, and it is essential that faults are rectified in the minimum time. If this unit approach is adopted, care must be exercised in choosing the unit sizes, or else an excessive quantity of spares will be necessary.

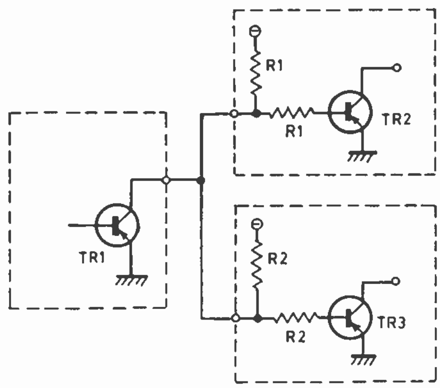
If a system is built up from a few given types of circuit block, then in most cases more components will be used than if circuits were specially built for the specification in hand. This arises because certain inputs and outputs are not always used, and because extra buffer stages have to be added in places where re-design of the stage using different resistor values would render a buffer stage unnecessary. The latter effect is most prevalent around the periphery of an equipment, where special indicators or stores have to be driven, or particular inputs accepted. Since telemetry and telecontrol equipment is concerned essentially with the routing of signals between locations, a large proportion of the circuits are dealing with components outside the electronic logic environment, and this effect is pronounced.

From the above discussion it becomes clear that there are definite advantages in arranging circuits in pluggable units with printed wiring, designed for particular functions. This approach has been adopted in Westronic Style F. A study of the system requirements enabled the functions of certain basic units to be decided. These circuit units are mounted on 9 in × 5 in printed-circuit boards with integral plug. Although the circuit configurations of gates, bistables, etc., are standardized, the component values used in each case are chosen to suit the local conditions. Care has been exercised in designing these basic circuits to allow the full flexibility which is required by the equipment.

All this logic operates at a fixed voltage level, nominally −9 V for '0' state, 0 V for '1' state. A fixed voltage level simplifies the design of individual stages, enables the maximum gain to be realized, and simplifies testing and fault-finding. The most important reason for choosing a fixed voltage-level scheme lies in the use of capacitor-coupled trigger-gates (Fig. 11(a)). This most useful circuit produces a positive output voltage spike at z, when a positive-going voltage step appears at x, provided that y is at zero potential. If y is sufficiently negative-biased then the output is inhibited. Thus the trigger-gate may be used for triggering bistable and monostable circuits, and it also provides a gating function. Consider now the importance of maintaining a fixed logic voltage-level in a system freely using many of these circuits. For the trigger-gate to inhibit the output completely, the negative bias at y must be greater in magnitude than the positive-going step at x. Now in conventional logic circuits, the potential at an 'off' transistor collector, i.e. the logic voltage-level, depends on the number of loads connected to it, and it is not possible to drive the x and y inputs to a trigger-gate from any randomly selected pair of logic signals if the gate is to inhibit. If all logic levels are the same, then there is no difficulty.



(*a*) Trigger gate.



(*b*) Principle of constant voltage-level logic.

Fig. 11. Logic circuits.

Since it is difficult to make all logic levels actually identical, two thresholds have been introduced into the trigger-gate. A voltage threshold applied at point A (Fig. 11(a)), and a current threshold introduced by a resistor in series at z, which limits the pulse current. This last also precisely defines the current load on the driving stage at x, and enables its gain factor-of-safety to be calculated.

The constant voltage-level is attained by providing equal resistance paths to the supply rail and to the common rail at each drive/load point. Figure 11(b) illustrates the principle. A pair of resistors are usually included also at the collector of TR1. No matter how many loads of this nature are connected to TR1, the 'off' collector voltage will be always -9 V, and trigger-gates may be safely driven from this point. As the number of loads is increased, however, TR1 'on' collector current increases too. Thus the upper loading limit is imposed by the collector current which can be handled by TR1. Tables can be produced for each output/input showing the current capacity vs. drain and the interconnection of circuits carried out accordingly.

Bistable circuits are also well served by this arrangement. Provided sufficient excess collector current is allowed, additional loads may be added to a bistable, without reducing the internal feedback in any way. Thus a good degree of flexibility is obtained, coupled with simple circuits which do not require large numbers of buffer stages.

The general mechanical arrangement of this equipment is well illustrated by Fig. 12, which shows a typical master station cubicle. The left-hand bay is occupied by the data-logging and electric typewriter drive circuits, while the right-hand bay largely contains the telemetry equipment.

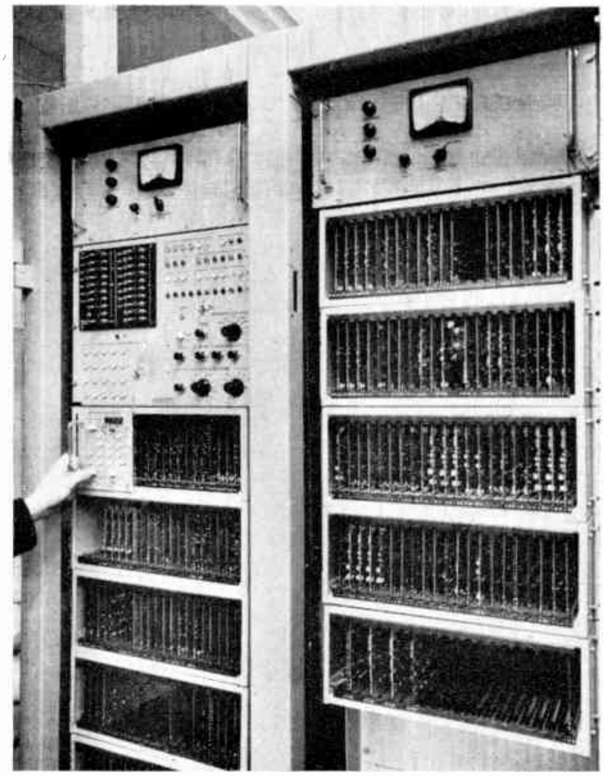


Fig. 12. General view of master station equipment cubicle.

4.7. *Fault Finding*

Before a fault-finding sequence can commence, there must be an alarm indication that a fault exists. The t.d.m. communication unit, and the addressing unit can be monitored fairly readily, but individual functions can only be alarmed by the system operator himself when he notices unusual occurrences.

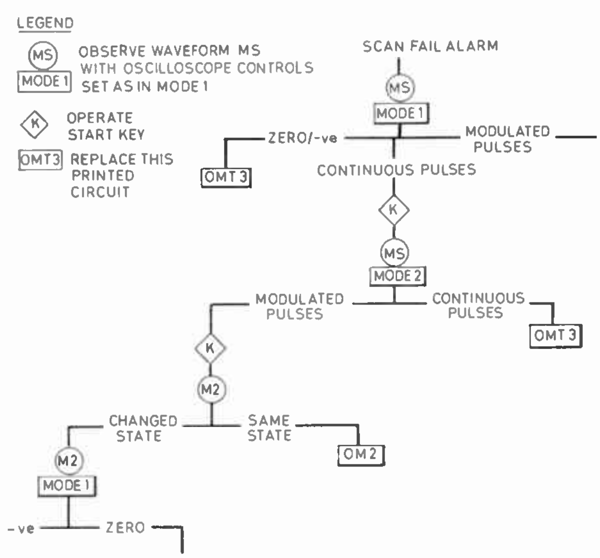


Fig. 13. Part of typical fault-location chart.

Alarms in Westronic Style F are based upon the action of 'delivery'. This functions to transfer incoming data into permanent stores only after all the available checks are applied to the received data. Since the out-stations are often unattended, the main alarm operates in the master station. Transmission from an out-station is conditional upon delivery of the received data being satisfactorily completed. Thus, when the master station registers a failure of delivery, this event could be caused by either a master or out-station fault, and the whole equipment is covered by this one alarm. The delivery alarm has a preset binary counter which registers the number of deliveries actioned in each major scan of all addresses, and a faulty delivery for three successive major scans brings up the alarm. In case the system should halt altogether, an independent timing circuit monitors the continued 'ticking' of the system.

Once an alarm condition is registered, a series of simple tests are sufficient to locate the fault in a given station, and finally in a given printed-circuit board. Guidance for these tests is given in the form of a systematic chart (Fig. 13). Each test position indicates a particular waveform to be observed, the result of the observation being of the yes/no variety which then takes the searcher on to the next step. The information is simply coded so that the chart is as free from unnecessary information as possible, but all the necessary data are close to hand. Access to the various waveforms is by a plug-in monitoring unit.

5. Conclusion

Some design aspects of an industrial telemetry system have been considered in some detail. Consideration of the basic philosophy of such a system has indicated that the design must be extremely flexible, and that the operating principles must be simple. Since the problem is largely one of communication between scattered out-stations and a controlling master station, good design of the t.d.m. unit which controls this communication is vital. Quite apart from the method of modulation which is chosen, the most important factor is the maintenance of time synchronism between terminal stations.

Design of the addressing and data-transfer facilities can often follow conventional logic, but provision must be made for random access to specific controls and telemeters. Control functions must be securely coded in most cases, and several simple alternatives exist which can also guard against a single circuit failure. Relay stores may be used with advantage in this context.

The choice of logic circuits and packaging techniques has been discussed at length, and the basis argued for a particular choice. Fault-finding, which is a very important aspect of this type of equipment should be considered early on in the design process, and particular provision must be made for semi-skilled fault-searchers.

6. Acknowledgments

The author wishes to thank the Westinghouse Brake & Signal Co. Ltd., for permission to publish the details in this paper. Many of the ideas mentioned in the paper owe a great deal to several of the author's former colleagues at the Company, and their assistance in many stimulating discussions is gratefully acknowledged. A number of these ideas are currently the subjects of Patent Applications.

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Two Anniversaries . . .

Thirtieth Anniversary of World's First Television Service

The 2nd of November 1966 marked the thirtieth anniversary of the opening of the first public high definition television service in the world, launched by The British Broadcasting Corporation at Alexandra Palace. The equipment for this first service was designed and supplied by the Marconi-E.M.I. Television Company Ltd., a joint venture in which the transmitting equipment and aerials were developed by Marconi and the studio and signal modulating equipment by Electric and Musical Industries.

This initial service was largely experimental, and for the first three months, the Marconi-E.M.I. system shared alternate weeks with the Baird mechanical, 240-line system. In February 1937, the Marconi-E.M.I. electronic scanning system was adopted permanently for the London station, and the standards used by this system, 405-lines, 50 frames interlaced to give 25 complete pictures per second, have remained to this day. This choice was recommended by the Postmaster General's Television Advisory Committee, largely because the complex mechanical equipment was obviously reaching its peak, whereas the electronic system was potentially more flexible, and capable of much further development.

The original Marconi transmitting equipment and aerials remained in operation at Alexandra Palace until 1956, with the exception of the war years, after which the B.B.C. London television station was moved to Crystal Palace.

The anniversary was commemorated by the B.B.C. on the evening of 3rd November by the first showing of 'The Discovery of Television', a 50-minute film produced for Mullard Ltd. and the B.B.C. using much authentic original material and information. The film relates a fascinating story, clearly bringing out the great controversy surrounding the Baird mechanical system but equally clearly showing the overwhelming advantages of an electronic system despite the technological problems which had to be solved. Most of the solutions in fact came from the industrial research group at E.M.I. headed by Isaac Shoenberg and staffed by such pioneers as Condliffe, McGee, White, Cairns, Broadway, Browne and Blumlein, the last two of whom were killed during the War while conducting radar experiments.

The concluding remarks by the commentator are particularly appropriate:

'Baird died in 1945, experimenting through the War on colour and stereoscopic television. Public houses, streets, memorial tablets perpetuate his name. It was he who obtained the first pictures. He made the going, but the system used throughout the world is not his. It was Campbell-Swinton who proposed the theory, Zworykin who pioneered and patented the electronic system; the E.M.I. team who created the modern television system. It was they who made television, as we know it, happen.'

Services Electronics Research Laboratory—21st Anniversary

This year the Services Electronic Research Laboratory (S.E.R.L.) celebrated its coming of age. Having as its chief ancestor the Co-ordination of Valve Development organization (C.V.D.), S.E.R.L. has been either the source of, or a close partner in, the development of many of the most vital electronic equipments provided for the British armed services.

The story of the Laboratory goes back before World War II when the Admiralty co-ordinated the development of valves for all three Services, thus anticipating by nearly thirty years the present-day unification of the supporting services to the armed forces. The C.V.D. organization was set up to provide for the necessary research and development in industry and the Universities. The rapid growth in the importance of radar added greatly to the requirements especially in the microwave field. A small research team directly responsible to the Admiralty already existed at Portsmouth and was thus able to act as the nucleus of this organization. After the war this was amalgamated with others to form the Establishment at Baldock. (An extension was opened a few years ago at Harlow.)

The special field of the Laboratory rapidly extended to cover the whole range of active electronic components. This nowadays includes gas discharge tubes, transistors and many other related solid-state devices which have revolutionized electronics. Quite a substantial proportion of the effort of the Laboratory at present is concentrated on lasers and it has become the centre for Government work on the two main types—the gas discharge, and semiconductor lasers.

Work initiated for military purposes often turns out to have non-military applications and it may be economical to follow these up. In this way projects are sometimes carried out for civil organizations. As the centre of a consortium for defence electronic devices which includes most of the important firms in the U.K. electronics industry and the Universities, the Laboratory has always maintained a close liaison with these colleagues and it is this co-operation which will continue to be cultivated in future in line with Government policy.

At 'open days' during October the Laboratory showed lasers of several different kinds in operation, microwave tubes of formidable complexity weighing more than a hundredweight, crystal lamps smaller than a pin's head, neutron tubes and many other electronic devices, both solid state and vacuum.

U.D.C.621.382.23

Trajectory of the Operating Point in a Binary Circuit using One Tunnel Diode

By

M. NALINIMOHAN RAO, Ph.D.†

Summary: The working principle of a simple binary circuit employing one tunnel diode is explained qualitatively using the trajectory of the operating point during its to and fro motion between the two stable states of equilibrium by the application of rectangular current pulses of either polarity. Trajectories in either case were constructed by the graphical method using the principle of Ussui.

1. Introduction

It has been suggested by Kaenel[1] and Whetstone and Kounosu[2] that a simple binary circuit can be constructed using one tunnel diode, in which the persisting nature of inductance has been made use of to switch the system from one stable state to the other and vice versa by the application of successive unipolar pulses. Study of the path traced by the operating point, as the circuit switches to and fro between its stable states, will help understanding of the processes involved in the action of the system as a binary circuit. In this paper such trajectories of the operating point have been plotted for the full cycle of switching using the principle of Ussui.[3]

2. Binary Circuit

A simple binary circuit can be constructed by connecting an inductance L, load resistance R and bias voltage E all in series with a tunnel diode (T.D.) as shown in Fig. 1. The applied voltage, E, and the resistance R, should be chosen in such a way that the load line intersects the volt-ampere characteristic of the tunnel diode at two points of stable equilibrium '1' and '0', as shown in Fig. 2, on the low and high voltage branches respectively. Inductance L in the circuit is necessary for the binary counting action to take place. Rectangular current pulses to be counted from a generator with large internal resistance are applied directly to the tunnel diode and output is taken from the same point. If various parameters are properly chosen, then this circuit switches back and forth between its stable states when current pulses (of any polarity) are applied and at the output one pulse will be obtained for every two input pulses.

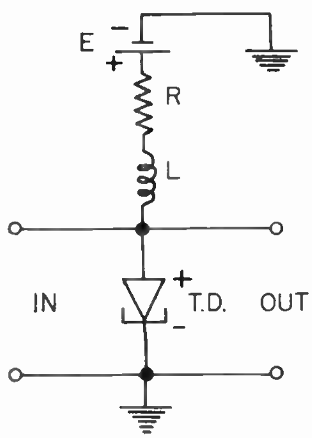

Fig. 1. Binary circuit using one tunnel diode.

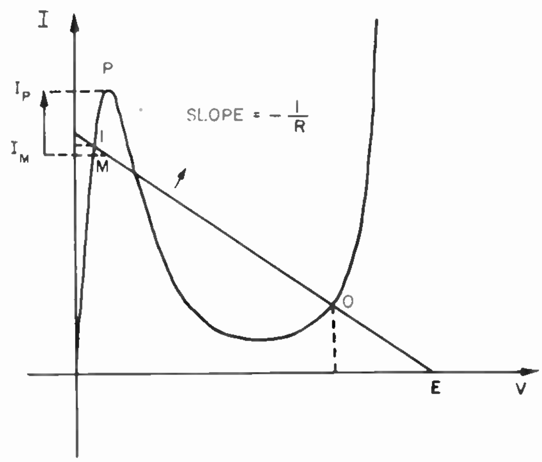

Fig. 2. Volt-ampere characteristic of the tunnel diode.

3. Working Principle of the Binary Circuit

Let positive pulses be applied to the input of the circuit for counting and let the operating point be initially situated at point 1 (see Fig. 2). Then the first positive input-pulse of amplitude larger than $(I_P - I_M)$ 'lifts' the load line above the peak of the characteristic curve and the operating point 'rises' up from point 1 and quickly 'jumps' to the high voltage branch.

The trajectory of the operating point in this case is as shown in Fig. 3. (The method of constructing this trajectory will be treated in the next section.) As is seen from Fig. 3, the operating point reaches point B on the high voltage branch. If there were no

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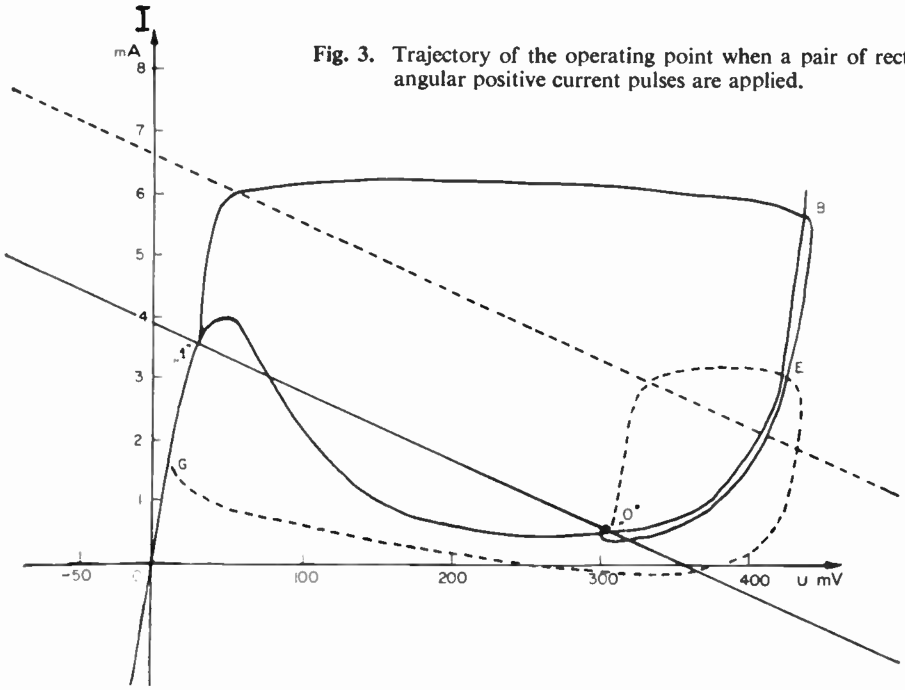


Fig. 3. Trajectory of the operating point when a pair of rectangular positive current pulses are applied.

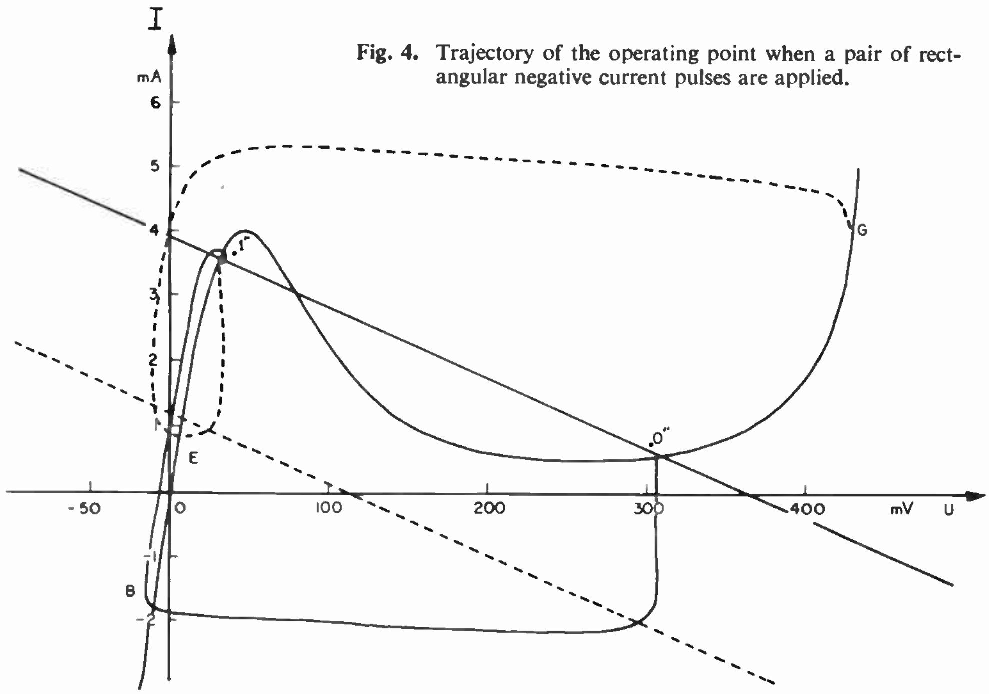


Fig. 4. Trajectory of the operating point when a pair of rectangular negative current pulses are applied.

inductance in the circuit, the path traced by the operating point would have simply coincided with the 'lifted' load line shown by dashes in Fig. 3. However, due to the presence of inductance, the operating point moves in a curvilinear path as shown by the continuous line in Fig. 3. During the duration of the input pulse the operating point will be held in the vicinity of point B because of the persistence of the inductor current. When the pulse ends, the operating point tends towards the second stable point O″, during which process it does not follow the characteristic curve, but takes a spiral-like path round the point O″ due to the presence of inductance. Abrupt ending of the pulse will be accompanied by a negative undershoot of voltage across the inductance. Parameters of the circuit should be so chosen that the amplitude of this undershoot is small enough not to drive the operating point back to the low voltage branch of the characteristic. The system then settles at the stable point O after the transients die out and stays there for any length of time.

The second positive input-pulse will first 'lift' the operating point from point O″ to point E (see Fig. 3), where it will be held during the duration of the pulse. When this pulse ends, again there develops a negative undershoot of voltage across the inductance, which will be sufficiently large (provided the parameters of the circuit are properly chosen) to shift the operating point to point G on the low voltage branch. From this point G the operating point finally moves back to the static equilibrium point 1 when the transients die out.

Thus, for two input pulses the system completes one full cycle and one pulse will appear at the output.

In a similar way, the same circuit shown in Fig. 1 works as a binary counter even if negative pulses are applied to the input. The trajectory of the operating point in this case is shown in Fig. 4.

4. Graphical Construction of the Trajectory of the Operating Point in the (i, U) Plane

The equivalent circuit of the binary of Fig. 1 is represented in Fig. 5, where

i_t = amplitude of the trigger current pulse
$i(U)$ = current through the tunnel diode, i.e. volt-ampere characteristic curve of the tunnel diode
C = capacitance of the tunnel diode
i_c = current through the capacitance C
i_L = current through the inductance L
R = load resistance
E = bias voltage.

The equations given below follow from Kirchhoff's laws for this circuit:

$$L\frac{di_L}{dt} + Ri_L + U = E \qquad \text{......(1)}$$

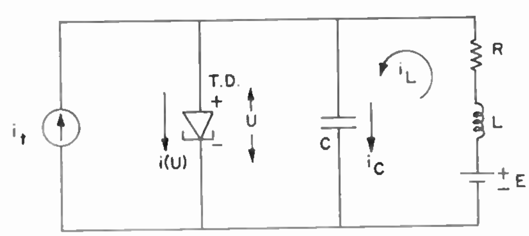


Fig. 5. Equivalent circuit of the binary counter.

$$c\frac{dU}{dt} = i_L + i_t - i(U) \qquad \text{......(2)}$$

Dividing eqn. (1) by eqn. (2),

$$\frac{di_L}{dU} = \left(\frac{c}{L}\right)\frac{E-U-Ri_L}{i_L+i_t-i(U)} \qquad \text{......(3)}$$

is obtained. Equation (3) can be used to construct the trajectory of the operating point in the (i, U) plane using Ussui's principle.[3] According to Ussui, the numerator of the right-hand side of eqn. (3) is proportional to the distance along the voltage axis from any selected point in the (i, U) plane to the straight line described by the equation

$$U = E - Ri_L \qquad \text{......(4)}$$

whereas the denominator is proportional to the distance of the same point along the current axis up to the curve described by equation

$$i_L = i(U) - i_t \qquad \text{......(5)}$$

This curve is nothing but the volt-ampere characteristic of the tunnel diode, bodily shifted along the current axis by an amount $-i_t$ and straight line (4) is just the load line.

The straight line represented by eqn. (4) is shown in Fig. 6. Curve (2), drawn in a continuous line, corresponds to eqn. (5), whereas the broken curve (1) is described by the equation $i_L = i(U)$.

At a given instant of time, let the operating point be situated at some arbitrary point A in the (i, U) plane and it is now required to find out the path of its future motion.

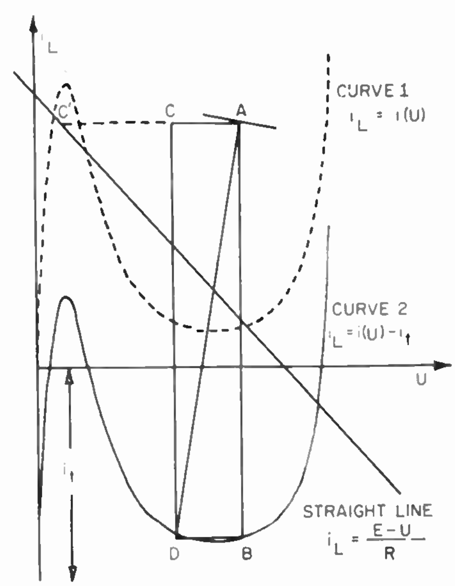


Fig. 6. Method of determining the direction of the normal to the trajectory of the operating point at any arbitrary point.

Now, from point A draw two straight lines—one parallel to the current axis and the other parallel to the voltage axis—until they intersect curve (2) at point B and the load line at point C′ respectively. Mark point C on the straight line AC′ such that

$$\frac{\mathrm{AC}}{\mathrm{AC'}} = R_0^2 \cdot \frac{C}{L}$$

where

$$R_0 = 1 \text{ ohm}$$

The segments AC and AB will be proportional respectively to the numerator and denominator of the right-hand side of eqn. (3). Coefficients of proportionality will be equal if the scales of representation along voltage and current axes are equal. Then, if a vector is constructed with components AB and AC, its direction will be perpendicular to the tangent to the trajectory of the operating point. The next position of the operating point in motion, say A_1, will be situated along this tangent and the closer one selects the point A_1 from the initial point the closer will be the constructed trajectory to the real one. Again starting from point A_1 the next position A_2 of the operating point can be found by a similar procedure and so on.

It must be mentioned that this method of plotting the trajectory is correct only when the scales of representation along the voltage and current axes are equal. That is, for example, 1 mV and 1 mA should correspond to 1 cm of distance along both the axes. However, in this case, the volt-ampere characteristic will be 'compressed' along the current axis and AC/AC′ will be very small, since in practice $C/L \ll 1$. Consequently, the accuracy in constructing the trajectory will suffer considerably.

This difficulty may be overcome by expanding the scale along the current axis, by a factor m, which is equivalent to representing the quantity $m.i_L$ instead of i_L along the ordinates. Equation (3) will take the new form,

$$\frac{\mathrm{d}mi_L}{\mathrm{d}U} = \left[\frac{m^2C}{L}\right] \frac{E-U-Ri_L}{mi_L+mi_t-mi(U)} \qquad \ldots\ldots(6)$$

Therefore

$$\frac{\mathrm{AC}}{\mathrm{AC'}} = \frac{m^2C}{L}$$

Now let us construct the trajectory of the operating point when a positive current pulse is applied to the binary circuit. For this purpose let us refer to Fig. 7 where the abscissae represent U and the ordinates represent $m.i_L$. The solid straight line in this figure is the load line described by eqn. (4). The broken curve (1) is represented by the equation:

$$m.i_L = m.i(U) \qquad \ldots\ldots(7)$$

and it denotes the initial position before the advent of the trigger pulse. The continuous curve (2) running parallel to curve (1) and shifted by an amount $m.i_t$ is represented by the equation

$$mi_L = mi(U) - mi_t \qquad \ldots\ldots(8)$$

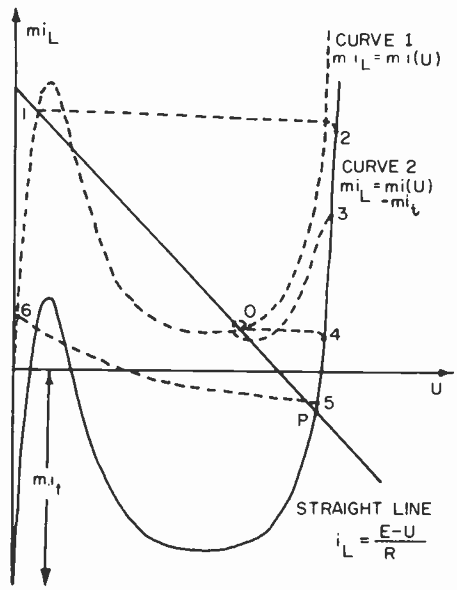


Fig. 7. Construction of the trajectory of the operating point in the (i_L, U) plane.

Here it should be noted that the effect of the application of the current trigger pulse, which is of course to move the operating point up or down depending upon the polarity of the pulse, can as well be achieved by shifting the axes in the opposite direction without shifting the operating point. Then, for constructing the trajectory of the operating point by the method of Ussui, one has to use the solid curve (2) of Fig. 7 described by eqn. (8) whenever the step current is acting, whereas when that step current is withdrawn the broken curve (1) described by eqn. (7) should be made use of.

Since the operating point is situated at point 1 before the advent of the first positive current pulse, point 1 itself will be the starting point for constructing the trajectory of the operating point and its motion is dictated by eqn. (6).

Starting from point 1 the trajectory of the operating point runs as shown in Fig. 7 up to point 2. During the interval t_w of the input pulse, the operating point 'slides' along the curve (2) and reaches a certain point 3 which can be determined using eqn. (9), given below. This can be derived starting from eqns. (1) and (2).

$$V_3 = \frac{[(G+a_1)V_2+(a_0-i_t-GE)]}{(G+a_1)} \exp\left[-\frac{(G+a_1)}{Ga_1L}\cdot t_w\right] - \left[\frac{a_0-i_t-GE}{G+a_1}\right] \quad \ldots\ldots(9)$$

a_1 and a_0 are constants in the equation $i_d = a_0+a_1 . V$ which approximates the upgoing branch of the characteristic of the tunnel diode. V_2 and V_3 are voltages corresponding to points 2 and 3 in Fig. 7, and $G = 1/R$.

Equation (9) does not take into account the time taken by the operating point to move from point 1 to point 2. This assumption is quite justified since this time is negligible in comparison with the time taken by the operating point to slide from point 2 to point 3.

When the input pulse ends, then for the further construction of the trajectory starting from point 3, the broken curve (1) and the load line of Fig. 7 should be made use of. Starting from point 3 the operating point moves in a spiral path round the static equilibrium point 0 and finally coincides with it.

If the width t_w of the pulse is very large when compared to the inductor time-constant (L/R), then the operating point may 'slide' down the curve (2) so much that after the removal of the input pulse the operating point may not reach the point 0 and may 'slip' back to the low voltage branch of the characteristic. Evidently, this régime cannot be used for a binary circuit. Thus for a given inductance L, width t_w of the input pulse has a higher limit.

When the second positive input current pulse is applied, for further construction of the trajectory, load line and curve (2) of Fig. 7 should be used. Starting from point 0, the operating point moves to point 4. For the duration of the pulse the operating point 'slides' along the curve (2) up to a certain point 5 which can be determined again by using eqn. (9) except for the difference that V_2 and V_3 are replaced by V_4 and V_5 respectively.

When the second pulse ends, the load line and the broken curve (1) of Fig. 7 should be used for further construction of the trajectory. Starting from point 5, the operating point moves up to point 6 on the low voltage branch of curve (1) and from there to the stable equilibrium point 1.

If the pulse width t_w is too small in comparison with the inductor time-constant, then for the duration of the second pulse, the operating point might not have moved very far off from the initial point 4 and when the pulse suddenly ends, the operating point may again spiral round the stable point 0 and finally coincide with it, instead of reaching the low voltage branch of the characteristic. Evidently, this régime also will not do for our purpose. That is, there exists a lower limit also for the allowed width of the input pulse for a given inductance.

It is easy to reconstruct the trajectory of the operating point in the (I, U), plane using the relation

$$i_L+i_t = i(U)+i_c = I \quad \ldots\ldots(10)$$

For this purpose the parts (1–2–3) and (0–4–5) of the trajectory should be 'lifted' up by an amount mi_t, since these parts correspond to increase in current due to the application of the current step, whereas the parts (3–0) and (5–6–1) remain where they are since these parts correspond to the removal of the current step. When these transformations are made the final trajectory looks as shown in Fig. 3.

Similarly, one can construct the trajectory of the operating point when negative current pulses are applied to the input of the binary circuit. One example of such a trajectory is shown in Fig. 4.

Figures 3 and 4 are drawn for a particular case of the following set of circuit parameters.

$$L = 10^{-5}\ \text{H}$$
$$C = 250\ \text{pF}$$
$$m = 100$$
$$R = 90\ \text{ohms}$$
$$E = 360\ \text{mV}$$
$$t_w = 0{\cdot}2\ \mu\text{s}$$
$$i_t = 2{\cdot}7\ \text{mA}$$

5. Shape of the Output Pulse

The shape of the output pulse can be predicted using the trajectory of the operating point. A pair of positive input pulses is shown in Fig. 8(a) and the shape of the output pulse corresponding to this pair is shown in Fig. 8(b).

In Fig. 8(b) the steep line (0–1) corresponds to the quick jumping of the operating point from stable point 1 to point 2 on the high voltage branch in Fig. 7. The duration of the line (1–2) is equal to the pulse width during which period the operating point is held up around the point 2. The slight slope is due to the fact that the operating point slowly 'slides' down along the high voltage branch of the characteristic from point 2 to point 3. The sudden dip (2–3) at the time of the cessation of the first input pulse corresponds to the sudden motion of the operating point from point 3 to the static equilibrium point O. The small depression at point 3 in Fig. 8(b) is due to the negative undershoot of voltage developed across the inductance. The flat line (3–4) corresponds to the existence of the operating point in the equili-

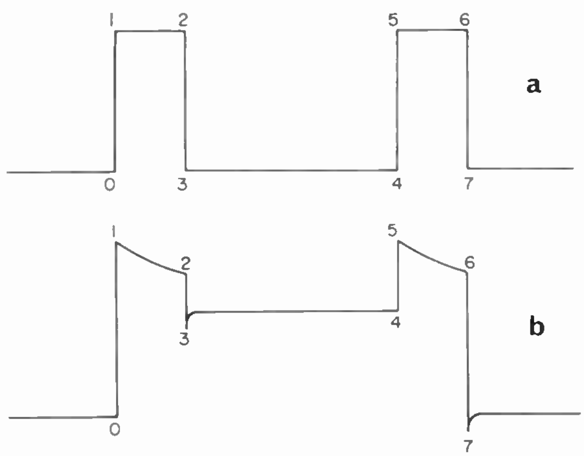


Fig. 8. (a) A pair of input rectangular positive current pulses.
(b) Shape of the output pulse.

brium position O. The sudden increase (4–5) corresponds to the jump of the operating point from stable point O to point 4 (i.e. point E in Fig. 3). The duration of the line (5–6) corresponds to the width of the second pulse and the slope is due to the slow motion of the operating point from point 4 to point 5 for the duration the second pulse. The steep line (6–7) corresponds to the switching of the operating point from point 5 to point 6 on the low voltage branch. The slight dip at point 7 and subsequent recovery to the zero level correspond to negative undershoot of voltage developed across the inductance after the sudden cessation of the second pulse and subsequent motion of the operating point to the initial point 1.

6. Experimental Results

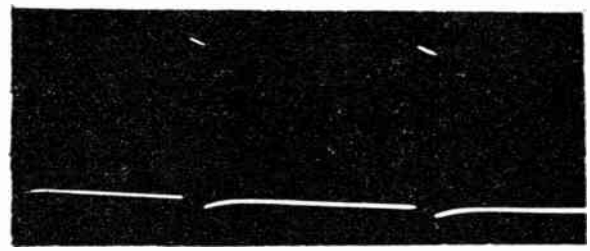
Oscillograms of a pair of input pulses and three types of output signals which correspond to minimum, intermediate and maximum values of inductance for a given value of pulse width are shown in Figs. 9(a), 9(b), 9(c) and 9(d) respectively. They closely resemble the predicted output pulse. Existence of maximum and minimum values for the pulse width for a given set of circuit parameters is confirmed.

Graphical construction of the trajectory of the operating point is helpful in qualitative understanding of the process involved in the binary action of the circuit. But for quantitative analysis of the circuit this method is very laborious and so it is not attempted here.

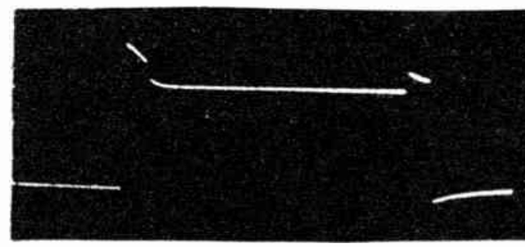
7. Acknowledgment

The work was done in the Department of Oscillations of the Faculty of Physics of the Moscow State University. I owe my thanks to Dr. K. S. Rzhevkin for his many valuable suggestions during this work.

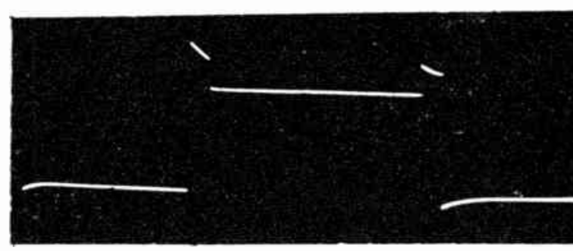
Fig. 9. Oscillograms showing:
(a) A pair of input positive pulses.



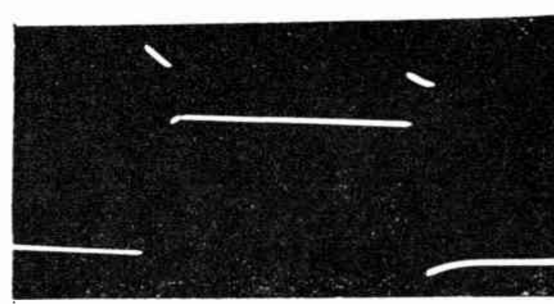
(b) Output pulse when the inductance in the circuit is a minimum.



(c) Output pulse when the inductance in the circuit is of intermediate value.



(d) Output pulse when the inductance in the circuit is a maximum.



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I.L.S. Monitoring—A Statement of the Problem

By

S. S. D. JONES, M.A., C.Eng.†

Presented at the Radar and Navigational Aids Group Symposium on 'Monitoring of I.L.S. Ground Equipment for Automatic Landing' held in London on 4th April 1966.

Summary: A brief description of the Instrument Landing System is followed by a detailed survey of its specific vulnerabilities to various forms of interference and the conclusion is drawn that, if I.L.S. is to be used as the guidance component of an automatic landing system for all-weather operations, an effective monitoring system must be provided.

1. Introduction

The purpose of this introductory paper is to set the scene for the detailed papers presented in this Symposium and to survey the overall technical background against which the work they describe has to be conducted.

The World's airlines are working towards progressive reduction of weather minima for landings. Their aim is to be able to achieve safe landings in zero visibility by the end of the present decade. It must devolve upon the national administrations (for example the F.A.A. in U.S.A. and the Ministry of Aviation in U.K.) to provide the necessary electronic ground equipment which has the performance suitable for successive lowering of weather minima towards 'zero-zero' visibility.

It is evident that the hands of the national administrations are not free; international operation is the essence of civil aviation and the electronic ground facilities provided by the various national administrations must therefore be mutually compatible. There must be an agreed international standard and the International Civil Aviation Organization has approved the Instrument Landing System as the standard approach-and-landing aid[1] and, moreover, has protected this system until 1975.

The Instrument Landing System (I.L.S.) comprises a v.h.f. split-beam sub-system as the azimuth guidance component ('localizer') and an u.h.f. split-beam sub-system as the elevation guidance component ('glide-path'). Like all radio-based systems, I.L.S. is subject to instrumental malfunction, drift of specified parameters, the effects of disturbances in the propagation path and 'interference'. A generation of meticulous engineering development has reduced the probability of instrumental malfunction to a very low level but this low level is, in part, achieved by the use of monitoring techniques; in the context of the safety standards with which civil aviation is concerned propagation disturbances and interference are still of serious magnitude and effective monitoring is needed to maintain the standard.

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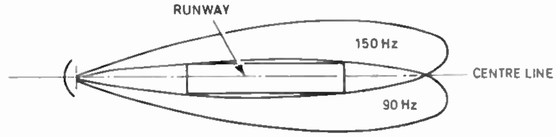
2. System Description

I.L.S. has been described fully elsewhere[2,3,4] but, for completeness and to serve as the basis of the statement of the need for monitoring, a brief system description is presented here.

Figure 1 shows the principle of the localizer. A transmitting aerial is located on the axis of the runway in use, in the overshoot area about a thousand feet beyond the end of the runway. It radiates a split beam, each of the two lobes being labelled by an individual modulation tone. The modulation frequencies are 90 and 150 Hz with a tolerance of $2\frac{1}{2}$%. The modulation depth is 20% with a tolerance of ± 2%. This low modulation depth was specified in order to accommodate a speech channel. The airborne component comprises an aircraft aerial which feeds a superhetrodyne receiver which has a very tight a.g.c. loop. The output of the second detector is proportional to depth of modulation. The receiver is followed (after the second detector) by a pair of



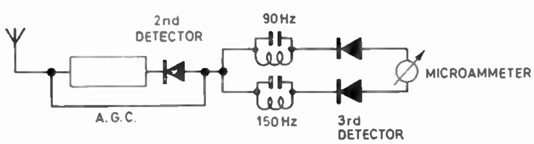


Fig. 1. Principle of I.L.S. localizer.

narrow-band audio-frequency filters which are tuned to 90 and 150 Hz respectively. These filters must have sufficient bandwidth to accommodate the tolerance in tone frequency. Each filter is followed by a third detector, and the differential output from the pair of third detectors, which is proportional to the difference in depths of modulation is applied to a central-zero microammeter of specified f.s.d. (± 150 μA) and resistance. The difference in depth of modulation is proportional to angular displacement from the course-line. This differential output, suitably slugged, can also be applied to the autopilot. The band 108–112 MHz is allocated to I.L.S. localizers. Channel spacing is 100 kHz and alternate 50 kHz channels are allocated to low-power V.O.R.

Modern localizers often have an azimuthal beam width which is too narrow for easy beam-joining.[5,6] This narrow beam width is desirable for reasons which become apparent later in this paper. Directional localizers are usually supported by an auxiliary beacon known as the 'clearance' system (Fig. 2). The clearance system is identical in principle with the localizer but the azimuthal radiation directivity pattern is designed to give coverage over 360°. The clearance system is displaced by a few kilohertz in frequency from the localizer so that, within the scale of the filter band-width, the two radiation diagrams do not add coherently and interference minima are avoided. Even so, the modulations corresponding to the two radiation diagrams do add coherently so that towards the limits of the coverage of the localizer pattern the clearance system captures the receiver without discontinuity.

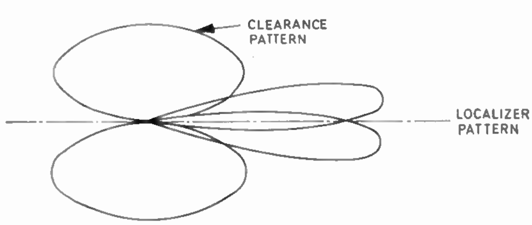


Fig. 2. Localizer and clearance arrays.

In the event of the modulation depth falling below a specified value or of excessive degradation of signal/noise ratio a warning flag is displayed on the I.L.S. indicator. The flag circuit is 'fail-safe' in that the flag is displayed when the circuit is not energized.

The glide-path component is identical in principle with the localizer, but the operating frequency is now in the band 329–335 MHz (Fig. 3). The glide-path origin is located about 1000 ft up-wind from the threshold of the runway and is displaced about 700 ft to one side. The equi-signal line which defines the glide-path is pre-set at an angle of about 3° with the horizontal. There are several forms of glide-path aerial but in all the ground in the vicinity of the aerials plays a substantial part in the formation of the two beams.

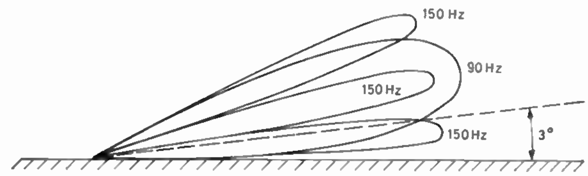


Fig. 3. I.L.S. glide-path.

I.C.A.O. I.L.S. installations are divided into three categories according to the weather limits in which they may safely be used by an appropriately equipped aircraft (Table 1).

Table 1

I.C.A.O. I.L.S. Categories

	Decision height (feet)	Runway visual range (metres)
Category I	200	800
Category II	100	400
Category III	zero	zero

3. The Operational Need

The operational need, so far as I.L.S. as the guidance system is concerned, is to define a course-line with sufficient accuracy to ensure that course-line errors, when added to control errors, aerodynamic perturbations, etc. result in a very low probability (of order 10^{-7}) of the landing aircraft missing the runway and of sufficient reliability to ensure that the guidance signal has a comparably low probability of failure during the last and vital 30 seconds or so of the landing operation. In this context the terms 'accuracy' and 'reliability' are not separable in that 'reliability' is synonymous with 'freedom from fault' and a deviation beyond the specified tolerances for a system parameter, however caused, is a fault. The term 'integrity' is commonly used to describe this necessary combination of accuracy and reliability.

Reliability of electronic equipment is commonly specified quantitatively by its mean time between failures (m.t.b.f.). If 'failure' is defined as a departure from specified parameters, whether caused by electronic malfunction, r.f. interference or any other cause, m.t.b.f. can also be used to specify 'integrity'. Assuming that a failure during the last 30 seconds of the landing operation is fairly likely to result in an accident to the aircraft and aiming at a risk of fatal accident of 1 in 10 million, multiplication of these numbers suggests that an m.t.b.f. of some five years would be needed. This figure, even though it might be marginally within the state of the engineering art for simple electronic equipment, is clearly unachievable in the case of a complex electronic system such as I.L.S. which operates in an open environment and is vulnerable to man-made and natural interference from radiating and reflecting objects. It might thus appear at first sight that safe automatic landing is not possible. The situation is, however, not so black as it might seem in that the danger lies not so much in failure as in undetected failure. It has been well established that in the event of a failure being detected promptly a landing aircraft can overshoot safely from a surprisingly low height which may be of order 50 ft but which of course depends on the type of aircraft. Below this minimum safe overshoot height it is safer to proceed with the landing in spite of subsequent failure of the guidance system. It is with the prompt detection of failure that monitoring, the subject of this Symposium, is concerned and it is upon the effectiveness of monitoring that the whole feasibility of the operation depends.

The range of the operational demands on the monitoring system is perhaps best defined by considering a hypothetical flight. Before take-off the captain of the aircraft will need to know the likely meteorological conditions at his estimated time of arrival at his destination and at the possible diversionary airfields. In the event of less than unusually good visibility being forecast he will also need an assurance that the I.L.S. at his destination and at the diversionary airfields is likely to be working at a category appropriate to the visibility at his estimated time of arrival. At this stage of the operation the monitoring needs are very modest and such a statement as 'Equipment out of action but likely to be repaired within three hours' might be sufficient to justify the decision to proceed with the execution of the flight-plan. On approaching the terminal area he will want to know the cloud base and runway visual range and also the category of I.L.S. service which is offered. Upon this information he will base his decision as to whether to attempt a landing at his intended destination or to divert. On joining the glide-path at about six miles from threshold he will want reassurance that the previously stated I.L.S. category has been maintained and on crossing the outer and middle markers at about 1000 and 250 ft respectively he will be monitoring the approach on his I.L.S. display. If, during this final approach, the flag indicators show a malfunction of the ground equipment he is likely to divert unless he has by this time established visual contact with the ground. At a height of about 100 ft the control signal from the I.L.S. glide-path will be disengaged so the pilot will be less concerned from this point with glide-path failure as indicated by his flag indicators or by v.h.f. telephony from Air Traffic Control. He must, however, continue to monitor the azimuth channel. When he has reached a height of about 50 ft he is virtually committed to a landing and any automatic disengagement which is triggered by the monitor will simply have the effect of causing the landing to be completed on 'memory'.

It will be seen from the above that the demands on the monitoring system becomes progressively more stringent as the execution of the flight-plan proceeds. The earlier stages are executed by means of environmental monitors and human judgement; the output of the environmental monitors, which can define the general status of the facility, has a warning rather than an executive function. On the other hand, the equipment must fail-safe immediately if its failure is the only indication of the need to break off the landing at a late stage. Consequently, the environmental monitor need not have an automatic executive function but instrumental monitors which detect equipment malfunction rather than the general trend of the operating environment must have a quick reaction and automatically switch off the equipment so that it fails-safe. The warning flag in the aircraft system will respond to the switch-off.

Some of the details of the application of the monitoring system have yet to be worked out. For example, even though it has been said that in some circumstances the equipment must be automatically switched off in the interests of maintenance of the 'fail-safe' facility, it could well be much more dangerous to switch the equipment off on its reversion from Category III to Category II standards than to allow it to continue at Category II level, so that an aircraft already committed to a landing could continue at a slightly reduced safety standard; following aircraft would be warned of the degeneration of the facility. On the other hand, a severe beam-shift which could cause a committed aircraft to swing off the runway should cause an immediate automatic disconnection so that a committed aircraft would have to continue a landing operation on 'memory'. Much consideration, which must be based on operational experience during fair-weather automatic landings, must still be given to the fixing of decision thresholds of the monitoring systems.

4. Specific Vulnerabilities

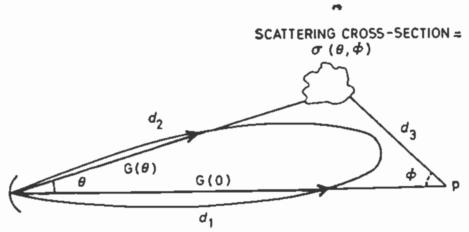
Like all electronic systems, I.L.S. has its own special-to-type vulnerabilities; it is against these vulnerabilities that the monitoring system must be planned.

The basic equipment must be as good as careful engineering can make it and the most modern I.L.S. ground and airborne equipment currently in use does achieve the necessary instrumental accuracy and stability. The necessary m.t.b.f. in terms of the reliability of the electronic equipment can be achieved, in the case of airborne equipment by duplication or triplication of the receivers, and in the case of ground equipment by duplication with automatic reversion to a stand-by transmitter in the event of specified tolerance of beam-position, control stiffness, power output, modulation depth or tone frequency being exceeded. The automatic reversion to the stand-by equipment, and automatic switch-off in the event of the stand-by equipment failing, is effected by means of the internal and near-field monitors to be described during this Symposium.

Like all radio equipment which operates in a quasi-free-space environment, I.L.S. is vulnerable to the effect of interference from man-made or natural sources. It is specifically vulnerable to certain types of interference which approximate closely to the desired signals.

4.1. '*Beam-bends*'

The I.L.S. localizer aerial has a beam width of several degrees and is located a mile or more from the touch-down point and several miles from the point of joining the glide-path. Consequently buildings and topographical features of the ground in the vicinity of the landing area are illuminated. Energy which is scattered from these objects adds vectorially to the field which corresponds to the primary radiation diagram of the aerial, resulting in distortion of the indicated course-line and perturbations throughout the region in which smooth control is demanded. The subject of 'beam-bends' has been investigated theoretically by Mercer[7] and has been the subject of much experimental flying. Theoretical analysis of an idealized case (Fig. 4) is simple but the beam distortion in a real environment is often very complex in character. The operating wavelength is about 3 metres and scattering angles vary over 360° so the periodicity in space of the disturbances can vary between a scale of very many wavelengths (up to kilometres) to a wavelength or so. In a real environment the scattering is likely to be derived from many sources and the vector sum of the contributions from these sources results in the pattern of perturbations being noise-like. The analogy to noise becomes even closer when one considers the signal received by the equipment in a landing aircraft which is making its approach at a velocity corresponding to twenty wavelengths per second or so. It is found that the observed beam noise



$$\frac{\text{Field strength of direct signal}}{\text{Field strength of scattered signal}} = \sqrt{\frac{\sigma(\theta, \phi)\, d_1^2\, G(o)}{4\pi\, d_2^2\, d_3^2\, G(\theta)}}$$

Fig. 4. Analysis of beam bends.

has a spectrum which ranges from several cycles per second to a very small fraction of a cycle per second. The amplitude of the beam noise varies from being trivial in ideal sites (for example, the Royal Aircraft Establishment at Bedford) through being potentially embarrassing at more typical sites (for example, London Airport) to a level which makes I.L.S. approaches impossible in certain mountainous areas. The contribution of the various components of the beam-noise spectrum varies according to the frequency of the components. The landing aircraft can respond to the very low-frequency components so these, according to their amplitude, add to the lateral scatter of the touch-down position. Components of somewhat higher frequency (of order 1 cycle/10 seconds) can be more serious in that they are of the order of the natural period of yaw-oscillation of large aircraft and may, in extreme cases, initiate a Dutch roll of increasing amplitude. Components of higher frequency still (of order 1 cycle/second) are significant in that, even though the control system of the aircraft cannot follow these high-frequency components, rate-stabilization is used in the control system and the rate is derived by differentiation of the displacement term. Consequently this high-frequency noise, exaggerated by the process of differentiation, is applied to the control system and may well drive it into the range of non-linearity. Acceptable limits of beam-noise have not yet been fully specified for I.L.S. for automatic landing, but it may be said that peak-to-peak amplitudes which correspond to displacements of order 10 ft are potentially troublesome. Even though the beam-noise pattern in space is quasi-static the amplitude can change in the medium term (for example, according to the number of parked aircraft in the vicinity) and it is incumbent upon the monitoring system to detect trends in changes of beam-noise structure.

4.2. *Interference from Over-flying Aircraft*

The localizer aerial, installed as it is in the overshoot area, amounts to a potentially dangerous obstruction and must be designed to present a relatively low silhouette. This means that the radiation-centre can only be a wavelength or so above the ground with the consequence that the maximum of the elevation polar diagram is elevated at an angle of several degrees (of order 15°). Consequently, the landing aircraft is flying well below the angle of maximum radiation and other aircraft on the circuit or in the vicinity may be illuminated by the ground transmitter at a much higher level than the landing aircraft. It has been established[8] that the scattering cross-section of large aircraft may, at certain bistatic angles, amount to many hundreds of square metres so that appreciable signals may be received by the landing aircraft through the medium of scattered energy from overflying aircraft. The operational significance of this phenomenon is increased by the fact that the varying geometry of the dynamic situation can result in Doppler shifts of the same order of magnitude as the tone-frequencies of 90 and 150 Hz, though the dominant effect is that of oblique incidence scattering which does not cause significant Doppler shift. An experimental data-gathering programme which is being conducted by B.E.A. on behalf of R.A.E. is showing that this phenomenon is of considerably more than merely academic significance. Figure 5 shows how, in a typical case, the components of an interfering signal derived from an over-flying aircraft may traverse the filter-bands. Detection of this phenomenon presents a particularly difficult monitoring problem in that the significance of the contribution of scattered energy depends critically on the geometry so that the aim must be to investigate the situation as presented to the landing aircraft.

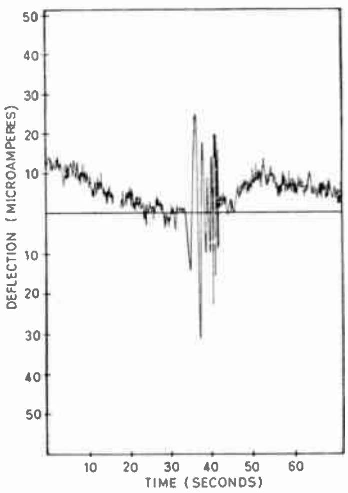


Fig. 5. Interference from overflying aircraft.

4.3. *Interference and the Receiver*

Even though this Symposium is dealing primarily with the examination of the electro-magnetic environment which is created by the ground equipment, it is necessary to consider the receiver because it is upon the receiver characteristics that the tolerable limits of various types of interference must depend.[9]

The receiver has an i.f. band-width of about 50 kHz and the second detector is followed by the 90 and 150 Hz tone-filters which have a Q of the order of 10. Consequently any signal which falls within the i.f. pass-band is a potential source of interference but a signal which differs in frequency by less than the tone-frequency bandwidth from the received carrier frequency is a much more potent source of interference in that the heterodyne between the desired and interfering signal will be accepted by the tone-filter.

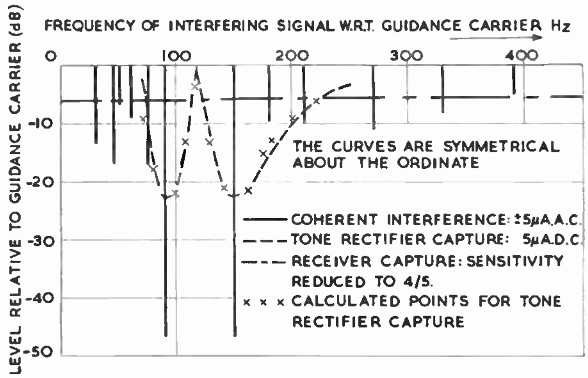


Fig. 6. V.H.F. localizer receiver interference thresholds.

Figure 6 shows, in synoptic form, the minimum significant levels of interference. The figure relates to a specific receiver but the results are very similar to those observed on other receivers and very close to theoretical. An interfering signal within the i.f. pass-band but different from the carrier frequency by substantially more than the tone-filter pass-band will, in the limit, operate the a.g.c. circuit, thereby suppressing the desired signal. A degradation in sensitivity of 1 dB will be produced by an interfering signal some 6 dB below the level of the desired signal.

Near-coherent interference at much lower levels is significant. Interfering signals 46 dB below the desired signal can, if they result in difference-frequencies near to the peaks of the tone-filter passbands, result in significant interference. This interference takes the form of an oscillatory variation in the indicated deviation from course-line, the period of oscillation being the difference frequency between the interfering signal and the carrier or side-band. Truly coherent interference of course produces a d.c. shift, but such interference is usually of short duration because of the differential Doppler shift between interfering and desired signals.

At higher levels of interference, tone-rectifier capture may become significant. This results from the interfering signal, beating with the desired carrier, and producing an a.c. signal which is passed by the tone-filter and rectified by the post-filter (tone) rectifier. Even though the significant level of interference (−22 dB in the worst case) is much higher than for coherent interference the effect can still be significant with heterodyne frequencies up to 200 Hz or so. Unlike near-coherent interference which produces a time-varying sinusoidal displacement error, tone-rectifier capture results in a d.c. shift.

4.4. *Instrumental Instability*

I.L.S. was originally conceived as an approach aid rather than as a landing-guidance system. Consequently the guidance parameters have been so chosen as to facilitate beam-joining at a distance of several miles from the runway threshold. At a range of, say, 10 miles the localizer provides a substantially linear control-law out to a distance of about 2000 ft on either side of the centre-line. The full excursion of the linear control law corresponds to a microammeter deflection of 150 μA. At threshold, where the greatest guidance accuracy is needed, the f.s.d. of 150 mA corresponds to a lateral displacement of 350 ft which means that if we are aiming at a lateral scatter of some 10 ft or so the receiver centre-zero must be held to within about 1% of f.s.d. Adequate stability of centre-zero is in fact achieved in modern receivers, but it is evident that test-gear of high quality must be provided for setting-up and checking.[10,11]

5. Conclusions

I.C.A.O. v.h.f. I.L.S., an approach aid which was conceived 25 years ago, can be developed to act as the landing-guidance component in an automatic landing system but only if suitable monitoring techniques and suitable test-gear can be developed. The supporting papers presented at this Symposium explain how the needs are dictated by propagation in the environment and how these needs can be met.

6. Acknowledgments

While the views expressed in this paper are those of the author, extensive use has been made of the work of colleagues in the R.A.E. as source material. I am particularly indebted to Mr. J. Benjamin, Mr. J. M. Jones and Mr. A. A. McCurrach.

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STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations, in parts in 10^{10}, from nominal frequency for **October 1966**

October 1966	24-hour mean centred on 0300 U.T.			October 1966	24-hour mean centred on 0300 U.T.		
	GBZ 19·6 kHz	MFS 60 kHz	Droitwich 200 kHz		GBZ 19·6 kHz	MSF 60 kHz	Droitwich 200 kHz
1	− 299·5	− 300·9	+ 1·7	16	− 298·5	− 300·3	− 0·1
2	− 297·6	− 301·0	+ 1·4	17	− 298·8	− 300·8	− 0·3
3	− 301·8	− 301·1	+ 1·3	18	− 298·0	− 299·9	+ 0·5
4	− 301·6	− 300·7	+ 0·4	19	− 299·2	− 300·0	+ 0·7
5	− 300·7	− 300·2	− 0·6	20	− 300·5	− 300·1	+ 0·7
6	− 298·9	− 300·4	− 1·3	21	− 299·0	− 300·4	+ 1·1
7	− 300·6	− 300·9	− 0·6	22	− 299·0	− 300·5	+ 1·3
8	− 300·6	− 300·6	− 0·4	23	− 298·7	− 300·3	+ 1·6
9	− 301·7	− 299·6	—	24	—	− 300·0	+ 1·8
10	− 301·7	− 299·3	0	25	− 297·9	− 300·1	+ 2·3
11	− 301·1	− 299·8	+ 0·1	26	− 297·9	− 300·0	+ 2·2
12	− 299·6	− 299·7	+ 0·7	27	− 299·0	− 299·6	+ 2·1
13	− 299·7	− 299·7	+ 0·5	28	− 299·5	− 299·9	+ 1·7
14	− 300·0	− 300·2	+ 0·3	29	− 299·8	− 300·0	+ 1·7
15	− 298·0	− 300·3	+ 0·4	30	− 301·2	− 300·2	+ 1·5
				31	− 300·0	− 300·5	+ 1·1

Nominal frequency corresponds to a va'ue of 9 192 631 770·0 Hz for the caesium F,m (4,0)–F,m (3,0) transition at zero field.

U.D.C. 621.391.812 : 621.396.933.23

The Character of the Received I.L.S. Signal and its Relation to Monitoring

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Presented at the Radar and Navigational Aids Group Symposium on 'Monitoring of I.L.S. Ground Equipment for Automatic Landing' held in London on 4th April 1966.

Summary: The paper reports an investigation of the general problem of radio propagation influences on the performance of the azimuth channel of the conventional Instrument Landing System. The main text is an attempt to define the characteristics of I.L.S. signals that are more directly relevant to monitor design.

To support the theoretical studies a localizer was installed at a disused airfield and the azimuthal guidance course was comprehensively monitored by means of signal probes erected in the required flight path. Geometrical-optical and diffraction theories were employed to assess the effects of possible re-radiating objects. The validity of the theoretical methods employed was checked by measurements at the disused airfield and at London Airport (Heathrow).

The difficult task of assessing the probability of harmful interference between localizers operating in temperate climates on the same nominal frequency has been attempted by analysing the published data and augmenting it by recording anomalously propagated signals.

Many flight trials have been undertaken in order to investigate the effect of overflying aircraft on the performance of the azimuth channel. Experimental records have been analysed in amplitude and frequency and compared with theoretical predictions. Results obtained with a 'stationary' landing aircraft have been used since the case of a real landing aircraft is at least no more critical in terms of the protection required, and an assessment has been made of the volume of air space which must be kept clear in order that there shall be a very low probability of interference from overflying aircraft.

1. Introduction

The authors have been associated with a study of the electro-magnetic medium both as a part of the environment and as an essential link in the operation of an I.L.S., with a view to assessing the limitations it imposes on the performance of I.L.S. as a self-sufficient blind-landing facility.

Although not all the parameters studied have a direct bearing on the design of I.L.S. monitoring facilities, it is considered that a broad understanding of the problem in its entirety is essential for a correct formulation of the functions of such monitors. For this reason accounts of the authors' studies of three main types of interference are given as appendices, while the main text of the paper deals with aspects that are more directly relevant to monitor design.

The results are mainly concerned with the lateral guidance channel (i.e. localizer), but, whenever it has been considered appropriate, extension to the vertical guidance channel (i.e. glide-path) has been attempted.

As in all communication problems we are concerned here with ensuring for a sufficiently high percentage of time a prescribed ratio between the wanted signal and unwanted components. The latter have various degrees of coherence with the former and may be described as 'interference' or 'noise'; either term will be used indifferently in the following text.

Whilst the assessment of such ratios for commercial communication systems is a standard exercise which the propagation expert solves by relying on comparatively well established experimental data integrated with a theoretical knowledge of the radio propagation characteristics, a similar exercise in the I.L.S. case requires a special study, mainly because the time percentage during which these protection ratios are allowed to fail is extremely small. Typical reliability figures are 10^{-3} for a commercial communication circuit and 10^{-5} for a Category III monitored I.L.S.

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2. Characteristics of Noise Affecting the Performance of I.L.S.

This section contains a broad survey of the main parameters of the interference types studied. These are reflections from overflying aircraft, reflections from ground sited obstacles and co-channel interference. The parameters to be considered may be summarized under the headings: level, spectral content, duration and height-gain. More details will be found in the appropriate Appendices.

2.1. *Level*

The peak level of interference produced by reflections from overflying aircraft can be large. Thus an error current of up to 100 μA in the difference in depth of modulation (d.d.m.) or guidance circuit output of a landing aircraft's receiver can result from an aircraft overflying the airfield at a height of 1000 ft, but the level measured in practice will depend on the time-constant of the d.d.m. circuit and the overflying aircraft's height and position. It is possible to define a volume of airspace such that the expected noise level is below a threshold of 5 μA when the overflying aircraft is outside this volume.

Very high levels of interference can be reflected from obstacles such as large buildings with vertical flat surfaces. For example, measured prior to the tone filters, a disturbance of ± 86 μA was produced by the B.E.A. Engineering Base in the early roll-out region of a westerly approach to No. 5 runway at London Airport (Heathrow). Similarly measured, the disturbance produced by the Control Tower was ± 16 μA. In the first instance the spectral distribution of the disturbance is not significant since it is outside the bandwidth of the tone filters; however, in the second case the signal is little attenuated since it falls within the pass band of these filters. Further protection at the guidance point is provided by the output time-constants relative to this point, but with the time-constants normally employed the Control Tower disturbance would be close to the limit generally accepted as ± 5 μA.

It should be pointed out that at London Airport the runways are flat. If a runway were humped, as is the case at many airports, reflections produced by similar obstructions would give rise to larger error currents with spectral distributions extending to the dangerous lower frequencies not evident at London Airport.

Regarding co-channel interfering signals, the emphasis is somewhat shifted. A very large level of interference is possible, in principle, by conceivable though rare tropospheric-propagation mechanisms, and the problem is then to assess the percentage of time during which dangerous interference levels are exceeded. It is predicted that the specified protection may fail amplitude-wise for as much as 1% of the time, the actual percentage in a specific situation depending on co-channel separation, path geometry and other factors. Although in most anomalous propagation events frequency separation will appreciably reduce the chances of an error signal of dangerous spectral distribution in the guidance output, the uncertainties with which such predictions are fraught suggest that the condition should be monitored.

Finally, it should be mentioned that in the distant undershoot at London Airport, d.d.m. output fluctuations reaching and exceeding 5 μA were observed. These random components have been attributed to the resultant of a multiplicity of scatterers. Scale-model investigations are being undertaken in order to characterize statistically the effect to be expected from a large number of scatterers, but in the meantime we can only draw attention to the existence of this type of fluctuation which is obviously relevant to monitor design.

2.2. *Spectral Content*

The spectral analysis of error signals caused by reflections from overflying aircraft shows that there is appreciable power in the error current at very low frequencies. In most cases the power is fairly evenly distributed between band 1 (0·01 to 0·04 Hz) and band 2 (0·04 to 0·2 Hz). Higher frequencies are also present.

For an overflying aircraft following a straight course, the minimum Doppler beat between wanted and unwanted carriers generally occurs at the landing aircraft when the overflying aircraft is approximately midway between the localizer and the landing aircraft. For the type of straight courses considered in the experimental study the rate of frequency change observed was always in excess of 0·1 Hz per second.

Considerations of reflections from static obstacles indicate that as a landing aircraft approaches from the distant undershoot region, the Doppler-beat frequency in the error current due to such a reflection increases steadily from a value near zero tending to a maximum which depends on the aircraft speed; for example, at 115 knots (i.e. 200 ft per second) a Doppler beat frequency of up to 44 Hz is produced. This figure applies to the azimuth channel and will be approximately three times greater for the elevation channel.

It should be pointed out that not only the absolute value but also the rate of increase of the Doppler beat frequency is small in the distant undershoot area. A compensating feature is the low level of reflected signal to be expected in this region. It is expected that our proposed scale-model experiment will throw light on the significance of this effect.

In the case of co-channel interference the frequency spectrum of the unwanted component is virtually identical to the frequency spectrum of the wanted component, except for a random frequency separation resulting from the signals not being locked. This separation may be sufficiently large to make the conditions safe, but there remains the need for continuous monitoring to ensure that the separation does not, as a result of frequency drift of either localizer transmitter, become small enough to create a significant disturbance. This suggests an advantage in a monitoring system incorporating spectrum analysis, which would enable a potentially dangerous situation to be identified before it became actually dangerous.

2.3. *Duration*

The duration of the disturbance from an overflying aircraft depends on where the threshold is set. A typical figure for a threshold set between 1 and 5 μA is 20 seconds. This figure is representative of an overflying aircraft following a straight course at a height of 1000 ft. In theory courses may be envisaged for which the disturbance lasts much longer but there is small likelihood of such courses being maintained. It should be pointed out that when the spectral line interference is produced not by the beat between wanted carrier and unwanted sidebands but between the wanted and unwanted carriers, then the overflying speed is much greater and the duration of the disturbance proportionally shorter.

An appreciation of the duration of disturbances caused by obstacles may be gained in two ways. From actual course bend records obtained at No. 5 runway, London Airport, disturbance of measurable level that can be defined as coherent is centred on the touchdown point and is traversed in approximately 10 seconds. On the other hand a calculation based on a hypothetical obstacle located 1000 ft from the runway yields the following figures for a typical aircraft approach speed of 200 ft per second. For beat frequencies of 0·01 Hz, 0·04 Hz and 0·2 Hz the aircraft to obstacle distances are respectively 33 000 ft, 16 500 ft and 7500 ft. From this it is inferred that the expected duration of noise in band 1 would be at most 83 seconds and in band 2 at most 46 seconds.

However, the corresponding levels of reflections from a single obstacle at these distances will undoubtedly be very small so that disturbances in bands 1 and 2 are more likely to be caused by a multiplicity of scatterers than by a single large obstacle. Of course if the emphasis is shifted to the multiplicity of scatterers the possible distance for obstacles from the runway can well be smaller than 1000 ft. For example, scatterers located at approximately 500 ft from the runway would produce the same frequencies as above at half the distances and for half the times.

Interference resulting from anomalous propagation can last up to several hours. Under these conditions signals are relatively steady and any superimposed fading will be very slow, perhaps of the order of several minutes.

2.4. *Height Gain*

As is usual in propagation studies the term height-gain is here used to describe the variation of signal strength with height. This factor is of primary importance in assessing the possibilities of predicting, from measurements made at or near the ground, the level of signal entering the receiver of a landing aircraft. For the purpose of assessing the presence of interference a knowledge is required of the height gain of both wanted and unwanted signals.

The height gain of the wanted signal is simply determined by the vertical radiation pattern of the I.L.S. transmitting aerial, which itself results from the interference pattern between the direct and ground reflected signals. The prediction from theory is a straightforward exercise, at least in the case of a flat runway, and even when the runway is humped can in any case be obtained from measurements. Once measured it can be relied upon to remain stable to within a few decibels under practically all meteorological conditions.

It is predicted from theory and supported by experiment that the reflected signal from overflying aircraft will be on average of the same order of magnitude at a landing aircraft at a height of 200 ft as at a possible monitor aerial at a height of, say, 70 ft. Since at these two heights the wanted signal varies in the ratio 3 to 1, the relative importance of the disturbance will be larger at the lower height. From measurements of overflying aircraft disturbances at two different aerial heights, it seems probable that the average magnitude of the effect experienced by a landing aircraft could be predicted to within ±10 dB.

The height gain of the reflected signal from obstacles exhibits a complicated lobed pattern resulting from the complex interplay of components generated by ground images. As a result of this it is extremely unlikely that any given reflected signal, as seen by a landing aircraft, can be reliably predicted from measurements carried out by a single ground-based aerial. It is conceivable however that, by using a plurality of ground-based aerials, a reliable prediction of the obstacle reflection present at the aircraft might be obtained. It is also possible that a single monitoring aerial will suffice for the statistical assessment of the long term behaviour of a multiplicity of obstacle-reflected signals.

For an anomalously propagated signal over a distance of, say, 160 miles (typical of separation between co-channel I.L.S.), it can be predicted from theory and has been confirmed by experiment, that the height gain approaches a linear behaviour over a height range from a few feet to about 40 ft at a frequency of 100 MHz, and over a somewhat smaller height range for a frequency of about 300 MHz. Above a height $h_0 = 50\lambda^{2/3}$, which in diffraction-propagation theory is referred to as the track width, the height gain is merely dependent on the variations in the obstructed-path length and is thus practically 'flat' over a height range of a few hundred feet. The track width is approximately 300 ft at 100 MHz and 150 ft at 300 MHz. In the intermediate region between the linear and 'flat' portions of the height gain curve, the signal variations are governed by too many parameters to permit a prediction in general terms. This is unfortunately the very region where information is most sought.

Even if the height gain of an anomalously propagated signal were adequately measured (e.g. by balloon), such measurements would only yield useful information regarding the average height gain, and the prediction on this basis at any given instant would still be affected by the uncorrelated and unpredictable amplitude fluctuations with time.

3. Limitations set by the Characteristics of Noise on the Performance of Possible Monitors

It would appear that the problem of confirming the integrity of I.L.S. controlled blind approach and landing requires two types of monitoring facilities with basically distinct functions.

One type, which can be called a 'performance monitor' has the primary task of collecting data over a long period of time, and such data would not normally be expected to be used during a particular approach. The collecting and analysis of data will be necessary for assessing the long term performance of the installation and also to test any assumption used in the previous certification of the installation.

The other type of instrument, which may be referred to as a 'warning' or 'executive' monitor, has the more difficult task of monitoring all flights individually and of giving warning when dangerous conditions develop. A dangerous condition arises when simultaneously a threshold level of interference is exceeded and its frequency lies within specified regions of the wanted spectrum and remains there for a specified period of time (e.g. 5 seconds).

It emerges from the information detailed in Section 2, in particular under the heading of 'Height gain' (Sect. 2.4), that the instantaneous prediction of an interfering field on the basis of measurement of the same field by means of a possible ground-based aerial cannot be accomplished with sufficient reliability. To allow for the variability in the height gains of all types of interference studied, the thresholds would have to be set considerably lower than specified by the response of the aircraft's instrumentation plus the automatic flight-control system, and consequently the resultant false-alarm rate would be unacceptably high. This rules out the possibility of using such a single ground-based aerial as an 'executive' monitor. Other aspects supporting this conclusion are the different spectral content of the interfering field as seen by the moving aircraft and a ground-based omnidirectional aerial, as well as the finite horizontal beamwidth of obstacle-reflected interference, although the last feature would be mitigated if more than one aerial were used.

If the task of a single ground-based aerial is restricted to the monitoring of comparatively long-term changes, such as the setting-in of anomalous propagation conditions giving rise to probable co-channel interference, then the above-mentioned height-gain variability would be smoothed out in the accumulated data, and the function of a 'performance' monitor would seem feasible. Such an installation would also provide an additional check to the 'near field' monitor on the instrumental performance of the localizer transmitter, aerial and feeder system. Since we are concerned with long-term changes, disturbances due to overflying aircraft would not be observed as a result of the large time-constants of necessity used. Slow drifts of the observed course line due to topographical changes, such as might be caused by the build-up of snow drifts or the installation of new power lines, would probably be observed as an error current, but the associated course bend pattern would not be revealed.

For the purpose of predicting the amplitude and frequency of low frequency noise, a multiplicity of aerials aligned on the extension of the runway would be necessary. The prediction on this basis of the spectral content of an individual interfering signal as seen by the landing aircraft would still be fraught with uncertainties. However, the information gained by scanning at a suitable rate the outputs of such a line of aerials could be used to provide long-term statistics directly interpretable as noise statistics in the aircraft receiver except for a conversion factor depending on aircraft speed.

The long-term monitoring of co-channel interference would basically require only one aerial and one receiver scanning the channel. Once the information collected by the monitor has been processed and statistics of levels and frequencies derived, these level statistics can be applied to the height of a landing aircraft by using an average height gain. In terms of frequency, the statistics obtained will presumably show a rectangular or bell-shaped distribution over the frequency-difference axis as a result of the frequency drifts of wanted and unwanted I.L.S. transmitters.

The frequency scale of such a distribution can safely be expected to be of the order of several kilohertz, so that this distribution is directly applicable to the case of the landing aircraft, the Doppler frequency components (of the order of hertz) being for this purpose negligible.

Another suggested type of ground based monitor relies on reflections from the landing aircraft to provide information on the purity of the guidance information as seen by the landing aircraft. Ideally such a monitor should provide on the ground a perfect copy of all signals entering the landing aircraft's receiver, the relative amplitudes and frequencies of all components being preserved.

A possible limitation of such a monitor, even assuming the direct signal from the wanted I.L.S. can be adequately attenuated, would seem to be the inevitable difference between the radiation patterns of the aircraft receiving aerial and of the aircraft itself as a reflector. It is very important that not only the direct wanted I.L.S. transmission but also any reflections of it produced by fixed or moving obstacles, other than landing aircraft, be also attenuated. The break-through signal impinging on the monitor either directly from the I.L.S. transmitter or directly from fixed obstacles could be filtered out for a known aircraft speed as long as this is less than 160 knots. However, such a useful feature would fail in the case of break-through signals from moving obstacles, e.g. overflying aircraft. These limitations might well result in the reflection monitor either missing the occurrence of a dangerous disturbance or giving a false alarm. It might be possible to mitigate such limitations by tailoring a directional aerial for each individual installation but this may be considered impracticable.

In view of these difficulties the favoured 'executive' monitor version is one in which the actual signal received by the landing aircraft's receiver is used to modulate an airborne transmitter and is transmitted to the ground for processing. As far as can be seen, such a monitor would overcome all the difficulties associated with the other types of monitors discussed.

It may be argued that the required air-ground communication link is unlikely to be agreed internationally. If this is so, then of all the ground-based monitors proposed, the reflection monitor appears to be the best suggestion to date for an 'executive' monitor.

4. Acknowledgments

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6. Appendix 1

The Effects of Overflying Aircraft on the Performance of an I.L.S. Localizer

It is well known that unwanted signals reflected from overflying aircraft could cause malfunctioning of the auto-control system of a landing aircraft automatically coupled to an I.L.S. localizer guidance signal. Theoretical and experimental investigations have been carried out into the magnitude and frequency distribution of this interfering signal, and an estimate has been made of the volume of air space which must be kept clear, in order that the interference should remain within tolerable limits.

6.1. *Theoretical Approach to the Prediction of the Expected Error Signal*

Spectral line interference due to reflection from an overflying aircraft is theoretically expected when the Doppler difference frequency between the direct and reflected signals assumes one of the values 0 Hz, 90 Hz, or 150 Hz, with smaller effects at sub-harmonics of the principal tones.[1] The maximum amplitude of the error signal under these conditions may be computed from the vertical radiation pattern of the localizer, the bistatic echoing area of the overflying aircraft, and the geometry involved.

There are two large uncertainties in this calculation, namely the radiation pattern of the localizer and the bistatic echoing area of the overflying aircraft.

An equi-signal localizer of a type in world-wide operational service was used in the experimental work. Since no comprehensive measurements of the radiation pattern of this localizer were available, a theoretical estimate was made on the basis of a simplified model. The localizer consists of two offset horizontal dipoles with a cylindrical reflector behind. One of the dipoles is fed with a signal modulated at 90 Hz and the other at 150 Hz.

The radiation pattern was computed from the effects of the dipoles and its image in the ground plane. In the forward direction, for rays which by geometrical optics would be reflected from the reflector, the pattern was assumed to be enhanced by an estimated gain at 14 dB.

The bistatic echoing area of the Vickers *Varsity* aircraft used in the flight trials was assumed to be 5000 ft^2. Extrapolation from experimental measurements available for models of *Comet* aircraft suggest that this value would be exceeded only for a small range of angles.

The difference in depth of modulation between the beams at the position of the overflying aircraft must also be taken into account in calculating the errors for zero Doppler difference. This is zero on the centre line and rises to a value approaching 0·4 at about 7° to the centre line. However, for errors at Doppler difference frequencies of 90 Hz or 150 Hz the difference in depth of modulation does not matter because the error is produced by the beating of the wanted and unwanted carriers.

6.2. *Experimental Measurements*

The experimental work associated with this study was undertaken on a disused airfield at Great Saling, Essex. A high mast was erected on the centre line of the runway, about 9000 ft from the localizer. Dipole aerials, which fed conventional I.L.S. receivers, were supported by this mast at heights of approximately 170 ft and 50 ft above localizer level (187 ft and 66 ft above ground level). The higher aerial sampled the field that a landing aircraft would receive. The overflying aircraft was a Vickers *Varsity*. The guidance signal deviations caused by this aircraft flying at about 130 knots at various heights up to 16 000 ft and on various courses, were observed.

The following error current recordings were made:

(i) An analogue recording from the 187 ft aerial with a smoothing *RC*-product of 0·15 seconds.

(ii) The same from the 66 ft aerial.

(iii) Analogue and digital recordings from the 66 ft aerial with a smoothing *RC*-product of 1 second.

It is believed[1] that a dangerous effect on the autopilot could be caused by an error signal of duration about 5 seconds and magnitude about 2½ μA in the frequency range up to 0·04 Hz or about 5 μA in the range 0·04 Hz to 0·2 Hz. For the purpose of obtaining an idea of the mean error currents over a period of 5 seconds, the records were split arbitrarily into 5 second periods and the mean and r.m.s. error currents computed for each. An example of a record processed in this way is shown in Fig. 1.

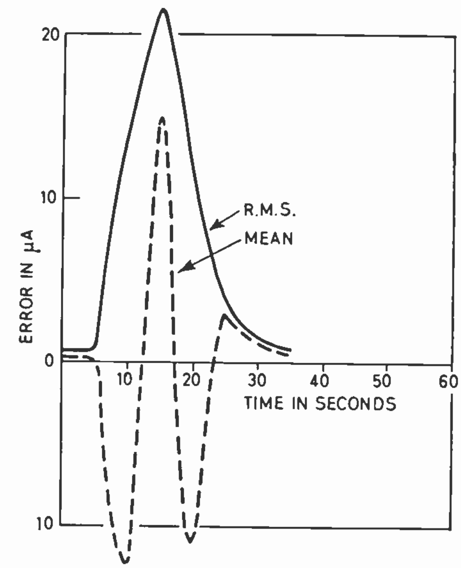

Fig. 1. Mean and r.m.s. error currents observed at mast. Overflying aircraft ground speed 130 knots, at height of 1000 ft on a course at 2° to runway.

All measurements made with *RC*-product of 1 second across output of receiver. This results in attenuation of frequency components above 0·16 Hz.

In order to compute the frequency distribution of the error current a power spectrum was computed for each run by first computing the auto-correlation function, and then taking the Fourier cosine transform of auto-correlation function. It should be noted that the digital recordings were made with a smoothing *RC*-product of 1 second, so that frequencies above 0·16 Hz are increasingly attenuated. An example of a computed power spectrum is shown in Fig. 2.

The differences between mean and r.m.s. error currents over 5 second periods are in general found to be small, suggesting that most of the power in the smoothed error signal is at very low frequencies. This is confirmed by the power spectra which show that the power distribution between frequencies above and below 0·04 Hz is roughly even.

Considered as a function of the height of the overflying aircraft, the mean error current falls off sharply at first as the height increases from 1000 ft to 2000 or 3000 ft. Thereafter the fall is much slower, values of about 1 μA being recorded at various heights between 5000 and 10 000 ft. Even at 16 000 ft ½ μA was recorded. The variation with height is shown in Fig. 3

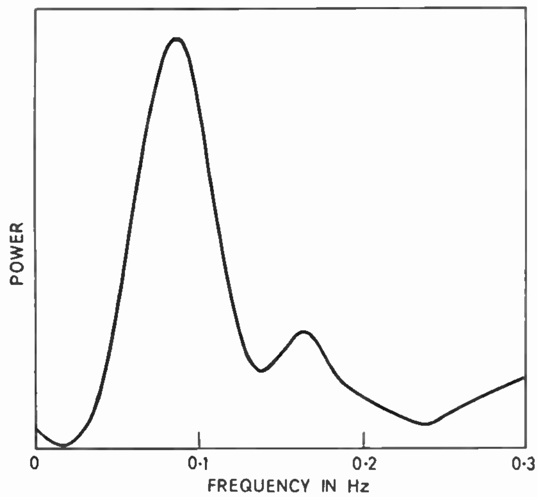

Fig. 2. Power spectrum of error current observed at mast. Overflying aircraft ground speed 130 knots, at height of 1000 ft on a course at 2° to runway.

All measurements made with *RC*-product of 1 second across output of receiver. This results in attenuation of frequency components above 0·16 Hz.

Curve is normalized to unit total power.

The more lightly smoothed error current is of interest in that the maximum value provides an indication of the values that might be reached under very exceptional conditions of zero Doppler persisting for a time of the order of 1 minute. The ratio of the more lightly to the more heavily smoothed values between 3 : 1 and 5 : 1 depending on the course and height of the overflying aircraft.

The more lightly smoothed values provide information on error currents at higher frequencies up to 1 Hz. These become serious only if the amplitude exceeds 5 μA for longer than 5 seconds.[1] This condition only exists at heights less than 5000 ft.

Besides the principal maximum at zero Doppler, subsidiary maxima were observed before the overflying aircraft reached the mast and after it passed the localizer. The Doppler difference frequency at these maxima was 30 Hz, and although this is not one of the principal spectral lines to which the receiver is sensitive, the large amplitude when the overflying aircraft is near the mast or localizer causes an appreciable effect. The interference region at 60 Hz Doppler is never reached with an aircraft flying at 130 knots.

6.3. *Comparison between Theory and Experiment*

The Doppler frequencies computed for the conditions of the flight trials have been compared with values measured from the analogue recordings and the agreement is found to be good. However, the theoretical error current amplitudes are smaller than those experimentally observed by factors between 3 and 5. This discrepancy is probably due to uncertainties in the localizer radiation pattern, particularly in view of certain terrain irregularities which could affect the radiation pattern at small angles of elevation.

6.4. *Extrapolation to the Real Case*

The situation with a real landing aircraft differs from the experimental arrangement in the height of the landing aircraft. In particular, Doppler difference frequencies of 90 Hz or 150 Hz might be possible, and could conceivably be sustained for a substantial period of time, although this is extremely improbable. In the worst conceivable situation of sustained 90 Hz Doppler from a large aircraft a correction factor of up to 10 might have to be applied to the experimental results, but in normal conditions the correction factor would be about unity.

The frequency distribution of the error current is also important, and this depends largely on the rate of change of Doppler. Calculations show that with an aircraft landing at 100 knots and an overflying aircraft on a straight course with a speed greater than 200 knots, the rate of change of Doppler is unlikely to be smaller than with a 'stationary' landing aircraft and an overflying aircraft at 130 knots, as in the trials. It is therefore, if anything, slightly pessimistic to assume that the frequency distribution is similar in the real case to that in the trials.

6.5. *Volume of Air to be Kept Clear*

In the flight trials dangerous error signals, as defined previously, were observed with overflying aircraft at heights up to 5000 ft, the worst effects being

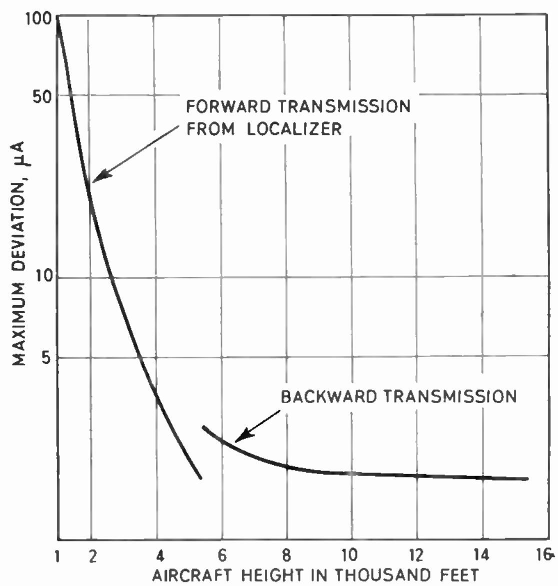

Fig. 3. Maximum deviation against height of overflying aircraft.

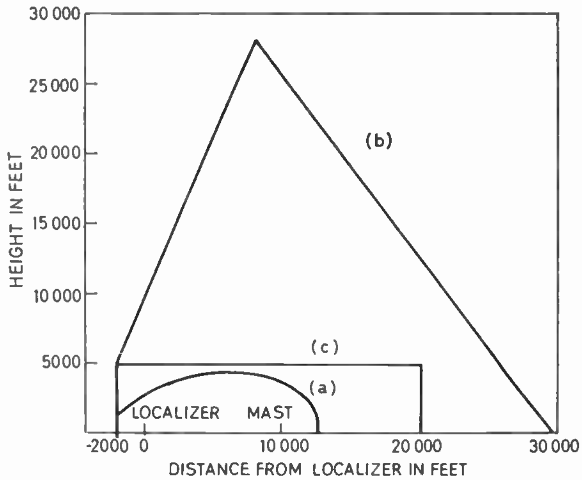

Fig. 4. Suggested protected air space.

(a) Air space in which significant noise was observed in the flight trials.

(b) Approximate volume in which significant noise might be experienced in the worst possible case. (Exact height uncertain since trials not available above 16,000 ft.)

(c) Suggested volume where there is an appreciable probability of significant noise in the real case.

with the overflying aircraft between mast and localizer. This volume must certainly be kept clear (see Fig. 4, curve a).

Calculations show that similar error currents might be obtained with the overflying aircraft further from the localizer (up to about 20 000 ft away) if Doppler difference frequencies of 60 Hz or 90 Hz were obtained in this region. Although these Doppler frequencies are certainly possible, no calculation has been made of the probability of their occurrence.

If a large aircraft were to fly in a curve producing a Doppler frequency of 90 Hz for a sustained period of 5 seconds, large correction factors up to 10 would have to be applied to the experimental results. This would lead to a large restricted volume of air as indicated by curve (b) in Fig. 4. However, the chance of this is very slight, although a numerical estimate has not been attempted. If the possibility can be disregarded, and if control procedures can eliminate the possibility of sustained zero Doppler interference from aircraft flying at less than 200 knots, then large correction factors need not be applied. Curve (c) in Fig. 4 would then indicate the volume of air space to be kept clear.

It is necessary to restrict the immediate vicinity of the localizer owing to the possibility of interference from side or back lobes at Doppler frequencies of 60 Hz or 30 Hz (as observed in the flight trials).

No direct evidence was obtained on the azimuth limits of the volume to be kept clear, but calculations suggest that for curve (c) a width of about 2000 ft on either side of the runway is adequate, so that the normal spacing of runways is therefore such that no interference should result from adjacent runways.

No consideration has been given to the effects of several overflying aircraft in the vicinity of a landing aircraft, or to the probability of occurrence of the very unfavourable conditions mentioned above which could lead to the much larger restricted volume (b). Moreover, consideration has only been given to a system using one type of equi-signal localizer, although it is believed that the radiation pattern of a null-reference localizer is such that interference from overflying aircraft would be at any rate no worse with this system.

7. Appendix 2

Theoretical and Experimental Investigations of Course-Line Disturbances from Flanking Objects

Large objects adjacent to an I.L.S. runway which lie within either of the localizer beams may scatter or reflect spurious radiation towards a landing aircraft, thereby causing deviations to the localizer's course line. Objects worthy of consideration include buildings, oil tanks, cranes, power lines, parked and taxiing aircraft and other vehicles. Acting singly, any of these objects may produce a simple periodic pattern of course bends, but the combined influence of numerous scattered objects may lead to random course deviations having some of the characteristics of 'noise'. The latter effects clearly merit further experimental and statistical study, but important as they are, they are outside the scope of this Appendix, where discussion is confined to the main features of course bends produced by a single reflecting object.

7.1. *Simple Features of the Error Signal*

Consider a reflector flanking an I.L.S. runway which is illuminated by one of the localizer beams, as in Fig. 5. An airborne I.L.S. receiver following an ideal (straight) course line will be subjected in the neighbourhood of the reflector to a spurious signal of progressively changing relative phase, which can be shown to produce a nearly sinusoidally varying

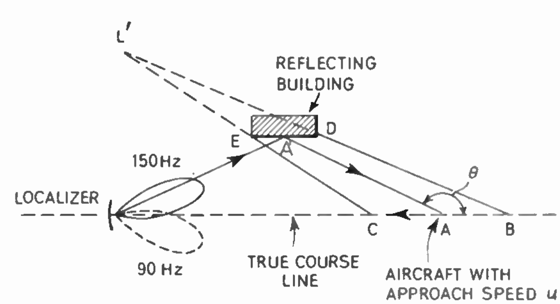

Fig. 5. Geometry of building reflection.

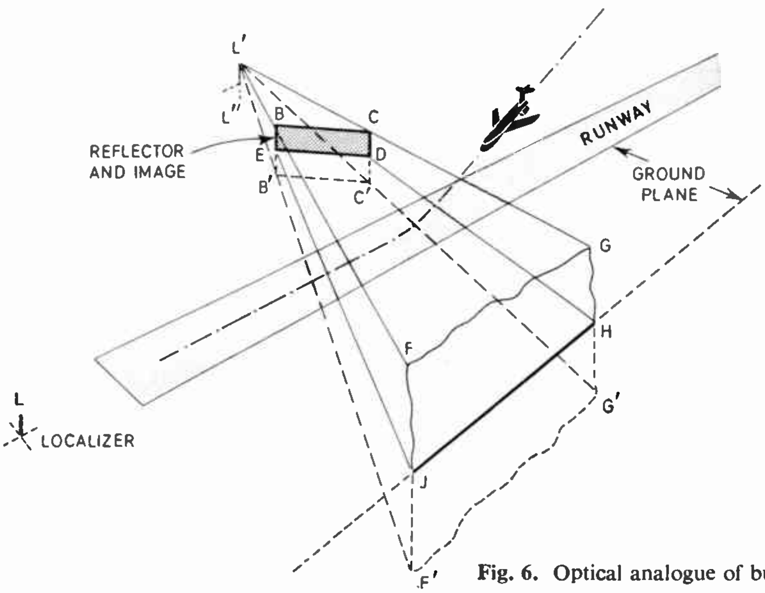

Fig. 6. Optical analogue of building reflection.

error signal in the course meter providing the disturbance is small in relation to the wanted localizer signal. Moreover, the frequency of this alternating error signal is simply the Doppler difference between the direct (wanted) and the reflected (spurious) frequencies received in the moving aircraft, which is given by

$$\frac{u}{\lambda}(1+\cos\theta) \qquad \ldots\ldots(1)$$

where, as shown in Fig. 1,

u is the aircraft approach speed,

λ is the localizer wavelength,

θ is the bearing of the reflector from the course line as viewed from the aircraft.

Turning now to the *magnitude* of the error signal, this is conveniently expressed in terms of a disturbance factor, γ, defined by

$$\frac{\text{level of spurious 150 Hz (or 90 Hz) modulated carrier}}{\text{level of wanted on-course 150 Hz (or 90 Hz) modulated carrier}}$$

On this definition it can be shown that an airborne I.L.S. equipment maintaining an ideal (straight) course would show a sinusoidal course error defined by a maximum difference in depth of modulation (d.d.m.) of

$$\pm 20\,\gamma\,\%\text{ approximately} \qquad \ldots\ldots(2)$$

so long as γ is sufficiently small and on the further (justifiable) assumption that the flanking reflector is significantly illuminated by only one of the localizer beams (Fig. 5).

Recalling the I.C.A.O. specification that a d.d.m. of 15·5% shall correspond to a left-right indicator current of 150 μA, the course error given by (2) may alternatively be expressed as

$$\pm 20\gamma \times \frac{150}{15{\cdot}5}\ \mu\text{A}$$
$$= \pm 194\gamma\ \mu\text{A} \qquad \ldots\ldots(3)$$

7.2. *The Disturbance Factor Relevant to Flanking Buildings*

An important facet of the wider problem involves the behaviour of the disturbance factor associated with a flanking reflecting building, idealized as a conducting rectangular sheet erected on flat, perfectly reflecting ground. The general behaviour of γ in these circumstances will be appreciated after a brief review of the main features of the resulting pattern of reflected radiation.

Consider for example a reflecting building illuminated by the left-hand (150 Hz) localizer beam as shown in plan in Fig. 5, illumination from the skirts of the 90 Hz beam being ignored. According to geometrical optics, the resulting disturbance along the course line is restricted to the region CAB defined by the localizer's image L′, where A′ indicates the point of geometrical reflection. However, diffraction causes appreciable radiation to spill over into the geometrically shadowed regions outside EC and DB, to an extent which is calculable by Fresnel methods.

Figure 6 shows a perspective view of an optical analogue which expresses the problem in terms of diffraction while taking account of the reflecting ground. L is the actual localizer considered as a point source 6 ft above ground level. The vertical reflector BCDE produces an image L′ (also at 6 ft a.g.l.) previously located in plan according to Fig. 5. There is also a secondary image L″ in the reflecting ground 12 ft below L′. The vertical reflector, itself

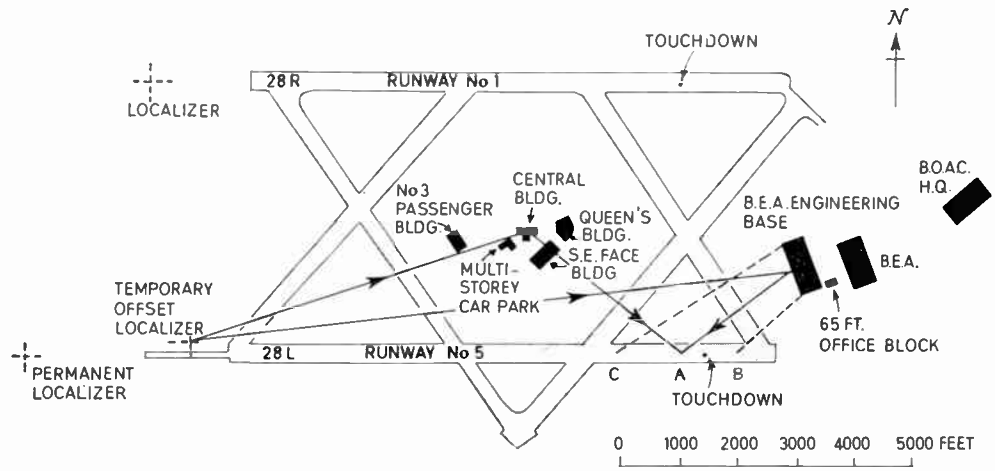

Fig. 7. Prominent buildings at London Airport.

mirrored in the ground, can now be replaced by a rectangular aperture BCC′B′ in the plane of the reflector which is illuminated by the image sources L′ and L″.

Following geometrical optics, radiation from L′ is confined to a tapered box which converges on L′ and is defined in Fig. 6 by the cross-section FGG′F′. Likewise radiation from the source L″ is restricted to a second tapered box converging on L″, which is omitted from Fig. 6 for the sake of clarity. Due to the small vertical separation between L′ and L″, the two boxes just specified are nearly coincident in the neighbourhood of the reflector, and hence the single box sketched in Fig. 6 may be considered as representative of the volume jointly illuminated by the respective sources L′ and L″. These twin sources, it should be emphasized, will interfere to create a vertical lobe pattern within the volume specified. The physically accessible portion of this volume is of course restricted to the upper half of the box lying above the ground plane EDHJ.

In broad terms, a landing aircraft is likely to suffer larger course-line disturbances within the specified box than outside it. Moreover, because the height of the box approximates to that of the reflecting building when the aircraft is relatively close to the latter, it is seen that an aircraft thus situated is more likely to experience significant course bends when it is below building height than at an earlier stage in its let-down. It should be stressed, however, that when the reflecting surface is sufficiently small or the distance between aircraft and reflector is sufficiently great, diffraction will spread significant radiation far outside the box prescribed by geometrical optics.

To calculate the reflected field-strength at a point on the track of the landing aircraft, Fresnel diffraction theory can be applied to radiation from the respective sources L′ and L″ through the virtual rectangular slot BCC′B′ (Fig. 6), the resultant field being derived from the vector addition of the individual contributions. A comparison with the easily computed 'wanted' field-strength at the airborne receiver then yields the required disturbance factor, γ, due allowance being made for the localizer's non-uniform horizontal radiation pattern (Fig. 5). The corresponding spurious d.d.m. and course error in microamperes are then quickly derivable from formulae (2) and (3).

7.3. *Course-Line Disturbances at London Airport*

An initial topographical study of London Airport (Fig. 7) showed that existing I.L.S. installations could be expected in some cases to be subject to course errors of the type described, although it must be emphasized that the disturbances subsequently recorded in the course of our investigations would not be detrimental to automatic approach and landing. The additional possibility of confirmatory flight tests encouraged the computation, on the lines outlined above, of the disturbance patterns attributable to selected prominent buildings lying within the localizer beams. Attention was first directed to the B.E.A. Engineering Base, whose long south-west facade lies well within the 150 Hz beam of the 28L localizer serving runway 5 (Fig. 7).

7.3.1. Theoretical behaviour of the disturbance factor from the B.E.A. Engineering Base

To conform with the condition of subsequent flight tests Fig. 8 was computed to show the theoretical variations of the disturbance factor as the aircraft approaches the 28L localizer on a true course at a fixed height of 50 ft. Scales of corresponding d.d.m. and course error in microamperes are also included. The well defined shoulders of the curve near the geometrical shadow boundaries stem from the fact that the building's 1000 ft long facade accommodates many Fresnel zones. It is seen nevertheless that

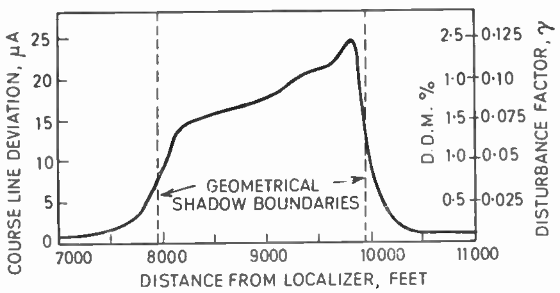

Fig. 8. Theoretical disturbance from B.E.A. Engineering Base.

significant disturbances occur outside the 2000 ft long geometrically illuminated region.

7.3.2. Theoretical and measured Doppler behaviour due to the B.E.A. Engineering Base

Application of formula (1) indicates that within the length of runway CAB 'illuminated' by the reflecting facade of the building (Fig. 7), reversals of course error at a frequency of 40–50 Hz can be expected for aircraft approach speeds of 115–145 knots. Although this frequency band is well outside the range causing impairment to an I.L.S. system, the effect was nevertheless investigated experimentally for the sake of the light it might throw on the general problem of course bends.

The above-mentioned frequencies and their associated intermodulation products are suppressed by the I.L.S. receiver output filters, and hence to disclose their presence it was necessary to by-pass these filters by recording the detector output. Although destroying the facility for direct d.d.m. measurement, this arrangement could be expected to reveal the rapid fluctuations in carrier level due to the flanking building. The localizer's 90 and 150 Hz modulations were temporarily suppressed, as they were superfluous in this application.

For the purpose of these tests, a high-speed chart recorder was connected to the i.f. detector output of the airborne I.L.S. receiver in association with a 0·05 second time-constant. The overall electro-mechanical response of this circuit was such that, with a 50 Hz input, the recorder indicated only 5% of its d.c. deflection. Thus a large correction for falling frequency response at high Doppler frequencies had to be applied when subsequently analysing the recorded detector output traces.

During May 1965 a Percival *Prince* aircraft of the Ministry of Civil Aviation Flying Unit, Stansted, carrying I.L.S. equipment modified as described, was flown visually, at a fixed height of 50 ft, as closely as possible along the true course line of the temporary off-set 28L localizer at that time serving runway 5 for westerly approaches (Fig. 7). This course was held at a steady speed from a distance of about 20 000 ft until the aircraft passed over the localizer. Several runs at speeds ranging between 110 and 140 knots gave consistent recorded traces of which Fig. 9 is typical.

The Doppler frequencies attributable to the B.E.A. Engineering Base in conjunction with an aircraft speed of about 110 knots are duly detectable on Fig. 9 as small-amplitude ripples, but they are unexpectedly superimposed on slower-period oscillations ranging in frequency between about 2 and 6 Hz, testifying to interference from a second and initially unsuspected source which will be considered later.

Confining attention for the present to the rapid fluctuations attributable to the B.E.A. Engineering Base, Table 1 lists the Doppler frequencies measured at four salient points on records of the detector output which were secured during four successive aircraft runs along the localizer's true course line at the speeds specified. Considering inevitable uncertainties in aircraft position and speed, the measured Doppler frequencies agree satisfactorily with the theoretical figures at the foot of Table 1.

Table 1

Measured and theoretical Doppler frequencies attributable to the B.E.A. Engineering Base

Date of test 16th May 1965; Frequency 110·5 MHz ($\lambda = 8{\cdot}90$ ft); Aircraft height 50 ft.

Aircraft distance from localizer	Comments	Measured Doppler frequencies† (Hz) at aircraft speeds specified			
		112 knots (Run 5)	110 knots (Run 3)	146 knots (Run 6)	143 knots (Run 4)
10 200 ft.	Start of coherent pattern	39	38·5	48	49
9600 ft	Region of maximum disturbance	38·5	39	48	49
9000 ft	Geometrical reflection point	39	39·5	48·5	48·5
8000 ft	End of coherent pattern	40·5	40	48	49
Theoretical values of Doppler frequency at 9000 ft from localizer		37·9	37·4	49·6	48·7

† One-second-mean values recorded during 4 successive flights.

Table 1 shows little change in measured Doppler frequency as the localizer distance shortens from 10 200 to 8000 ft, and this behaviour can be shown to be consistent with the path geometry. The combined evidence leaves little doubt that the observed effect indeed stems from the B.E.A. Engineering Base.

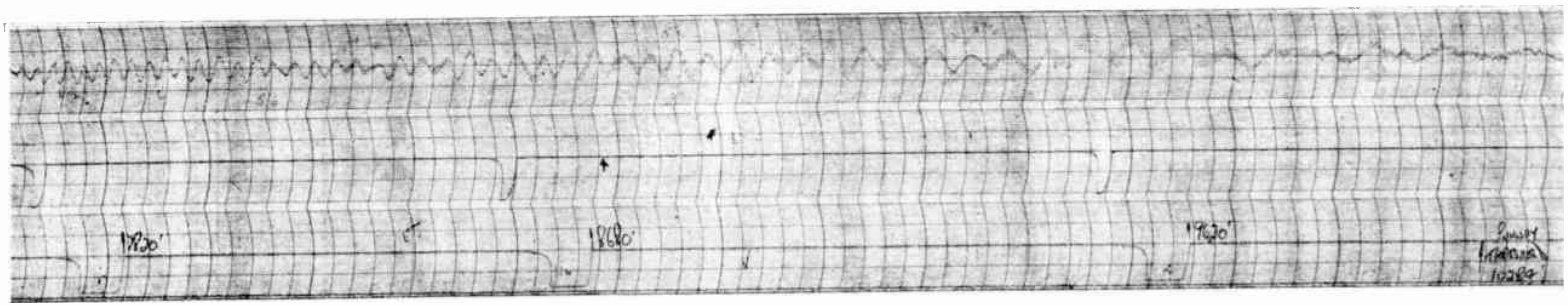

Fig. 9. Detector output recording. Composite disturbance, run 5, 16th May 1965. Aircraft speed 112 knots, I.L.S. frequency 110·5 MHz.

Fig. 10. D.D.M. recording. Control tower disturbance, run 2, 7th April 1965. Aircraft speed 100 knots, I.L.S. frequency 110·5 MHz.

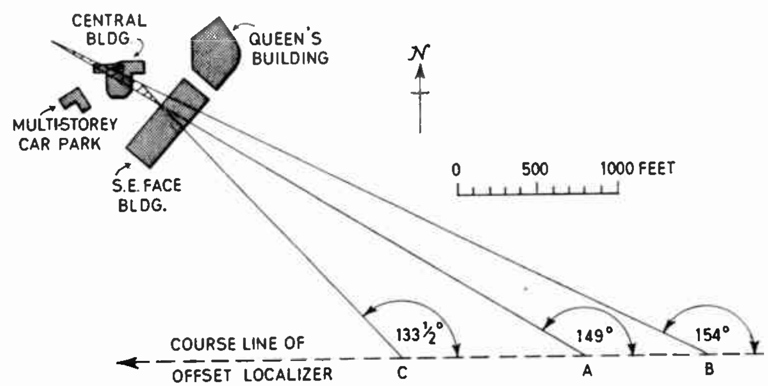

Fig. 11. Doppler intersections in central area.

7.3.3. Identification of the second Doppler source

As already explained, the long-period component of the composite pattern of Fig. 9 indicates the existence of a second source of disturbance requiring identification. At any point on the record corresponding to a known localizer distance, formula (1) can be used to relate the measured Doppler frequency to the bearing, θ, of the unknown source (Fig. 5), thereby assisting in its location. Although suitable in principle for such an analysis, Fig. 9 and companion records were not, in fact, used for this purpose, because these particular charts could not be accurately calibrated in terms of distance due to the temporary absence of Decca position fixes. Purely for the purpose of locating the second source, therefore, use was made of related earlier records exemplified by Fig. 10, which had the benefit of Decca fixes and hence of enhanced accuracy as regards aircraft speed and position.

Unlike Fig. 9, Fig. 10 is a record, taken over the same flight path, of conventional I.L.S. operation involving the left-right indicator current subject to respective time constants of 0·2 and 0·05 seconds. In this situation the higher-frequency ripple† already discussed is suppressed in the output filters, but the low-frequency ripple emanating from the unknown source passes through the filters to give the trace typified by Fig. 10 which, it will be noticed, is a fair replica of the low-frequency component on the composite trace of Fig. 9.

Based on Fig. 10 and companion records, the right-hand column of Table 2 shows the values of θ (Fig. 5) consistent with mean Doppler frequencies measured respectively at the edges and centre of the coherent portion of the pattern generated by the unknown source. As no substantial reflecting objects exist to the south of runway 5, the prescribed angles, θ (Table 2) were erected to the north of the course line at points defined by the relevant localizer distances. As seen from Fig. 11, the 'cocked hat' formed by the intersection of the resulting three lines indicates with fair precision that the second disturbance emanates from the Central Building.

† The rapid ripples just discernible on the lower trace of Fig. 10 result merely from the 60 Hz beat between the 150 and 90 Hz modulations.

From Fig. 7 it is seen that the Central Building is partly screened by the No. 3 Passenger Building, the Multi-storey Car Park and the S.E. Face Building. However, the southern face of the Control Tower is unobstructed above about 60 ft in both relevant directions. The upper part of the Control Tower emerges therefore as the only plausible reflector responsible for the lower-frequency course bends.

Table 2

Location of source of disturbance from measured Doppler frequencies

Date of test 7th April 1965; Frequency 110·5 MHz (λ = 8·90 ft); Aircraft height 50 ft; Aircraft speeds: 100 knots for runs 2 and 4, 104 knots for run 3.

Aircraft distance from localizer	Comments	Measured Doppler frequencies (Hz)†	Value of θ ‡
9800 ft	Start of coherent pattern	1·91	154°
9000 ft	Region of maximum disturbance	2·67	149°
7800 ft.	End of coherent pattern	5·85	133½°

†One-second-mean values, average of 3 runs typified by Fig. 10.
‡Inferred from Formula (1), see Fig. 5.

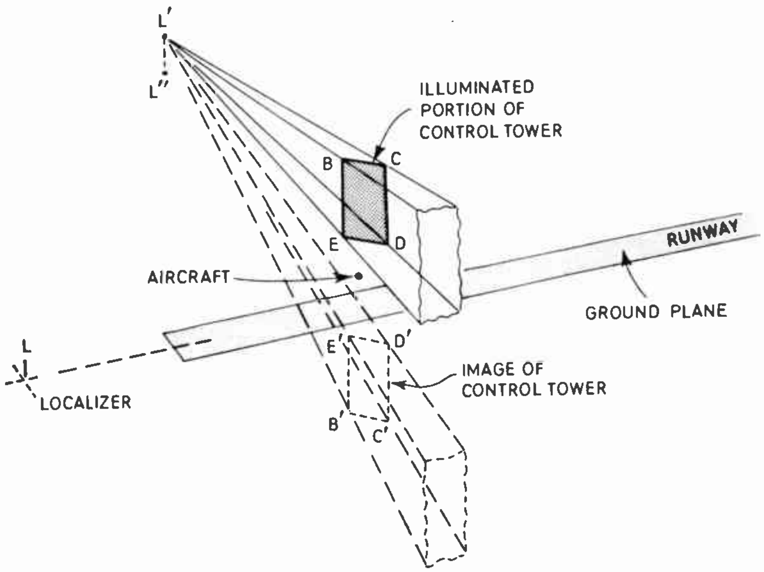

Fig. 12. Illumination restricted to upper part of control tower.

7.3.4. Theoretical disturbance factor attributable to the Control Tower

The Control Tower has a 33 ft wide facade which rises to 110 ft a.g.l. and faces slightly west of south so as to direct a fan of reflected radiation on a sector of course line roughly coincident with that affected by the B.E.A. Engineering Base (Fig. 7). In fact the geometrically reflected rays from the respective buildings happen to intersect almost on the course line.

Because of its narrow width coupled with the shallow reflection angle involved, the Control Tower spreads radiation far into the geometrical shadows. Moreover, because only the top 50 ft of the Tower is 'illuminated', the optical analogue takes the rather involved form of Fig. 12. It so happens that with the configuration considered, the contributions from the two Fresnel 'windows' BCDE and E'D'C'B' combine in partially additive phase at the airborne I.L.S. receiver to give the quite substantial disturbance factor, γ, of about 0·06, corresponding to a left-right indicator current of about ± 11 μA.

7.4. *Comparison between Theoretical and Measured Disturbance Factors*

Considering first the d.d.m. records typified by Fig. 10, we have a 2000 ft sector of course line showing coherent deviations produced by the Control Tower. Large regions outside this sector (excluded from Fig. 10) show relatively small and random fluctuations due to scattering from a multitude of small objects combined possibly with contributions from weak 'side-lobes' from larger randomly placed reflecting surfaces. As previously mentioned, the study of these important random influences cannot be pursued here, and attention must be confined to the coherent sections of the records attributable to identifiable discrete reflectors.

Table 3 summarizes the mean course deviations from three traces which were recorded in quick succession at a fixed aircraft height of 50 ft, and were directly calibrated in left-right indicator current and d.d.m. The disturbance shows, as expected on the score of the narrowness of the reflecting Control Tower, a wide spread with a poorly defined maximum at the geometrical reflection point. This trend will be best appreciated from a study of Fig. 10. Considering the inevitable departures between theoretical idealizations and practice, the theoretical course deviation at 9000 ft from the localizer (Table 3) shows promising agreement with the measured value.

7.4.1. The B.E.A. Engineering Base disturbance

Reverting to the detector output records, an impression of the B.E.A. Engineering Base disturbance can be obtained from the small-amplitude 40–50 Hz ripples on Fig. 9, bearing in mind that they are grossly de-emphasized relative to the 2–6 Hz Control Tower ripples due to the drooping frequency response of the recorder.

Assuming a linear receiver and an a.g.c. time-constant substantially longer than the slowest coherent ripple period, it can be shown that the disturbance factor (and hence the spurious d.d.m.) is proportional to the peak-to-trough amplitude of the relevant ripple. This relationship has permitted the calibration of the detector output records in terms of spurious d.d.m. by matching the reliable central portion of a d.d.m. record (Fig. 10) against the corresponding (long-period) ripples of a detector output record (Fig. 9). The d.d.m. scale thus derived has then been used to evaluate the B.E.A. Engineering Base disturbance after correcting the relevant short-period ripples according to the known frequency response of the recorder.

Table 4 lists the measured course-line deviations

Table 3

Measured and theoretical course deviations attributable to the Control Tower

Date of test, 7th April 1965; Aircraft height 50 ft; Aircraft speeds: 100 knots for runs 2 and 4, 104 knots for run 3.

Aircraft distance from localizer	Comments	Measured course-line deviation†			Course-line deviation theoretically attributable to the Control Tower		
		Left-right indicator current, (μA)	D.D.M.%	Disturbance factor, γ	Left-right indicator current, (μA)	D.D.M.%	Disturbance factor, γ
9800 ft	Start of coherent pattern	$\pm$ 8·5	$\pm$ 0·9	0·045		not computed	
9000 ft	Geometrical reflection point and region of maximum disturbance	$\pm$ 16	$\pm$ 1·65	0·08	$\pm$ 11·3	$\pm$ 1·17	0·06
7800 ft	End of coherent pattern	$\pm$ 7	$\pm$ 0·7	0·035		not computed	

† Two-second mean values averaged over 3 runs.

due to the B.E.A. Engineering Base, together with the corresponding theoretical deviations. It is evident that although the position and general shape of the theoretical pattern (Fig. 8) have been followed, the measured deviations exceed the theoretical values by a factor of about 3. Although not yet experimentally confirmed, the following reason for the discrepancy is suggested.

The low heights of localizer and aircraft involve a very small phase difference between the direct and ground-reflected paths, thus producing a condition very vulnerable to further relative phase shifts associated with diffraction through the appropriate rectangular aperture (Fig. 6). It can be shown, in fact, that for the conditions investigated, diffraction by this aperture imposes a supplementary relative phase shift of about 1·7° on that associated with unimpeded 'flat earth' reflection, and in such a direction in this instance as greatly to enhance phase opposition between the constituent ray paths. It may be significant moreover that *neglect* of this supplementary phase shift brings the theoretical figure into good agreement with measurement.

It may indeed be ultimately established that this small but influential phase shift tends in practice to be partly or wholly suppressed due to the fact that the facades of buildings seldom have the perfectly rectangular outlines implicit in the Fresnel theory. In this connection, irregularities in the upper edge of a reflecting facade may be particularly significant. Another critical factor may prove to be the building's surface irregularities.

It is hoped eventually to test these conjectures by further flight tests at a series of carefully controlled aircraft heights. In the final event a semi-statistical approach to the problem may prove necessary.

8. Appendix 3

The Co-Channel Performance of I.L.S.

It is often maintained (less often, perhaps, on this side of the Atlantic) that, on the score of operational experience accrued to this day, the magnitude of co-channel interference is too small, and its frequency of occurrence too rare, to be a cause of concern in the use of I.L.S. as a self-sufficient blind-landing facility. The problem may similarly be dismissed also by arguing that very simple (and compatible) modifications to the existing system would altogether eliminate co-channel interference.

It is doubtful, however, whether reliability parameters implying mean-time-between-failures of the order of five years may confidently be inferred from operational data gained over a comparatively short time. Also, if cure there is to be, the disease should be diagnosed first.

For this reason we shall take cognizance of the fact that anomalous propagation events do occur at I.L.S. frequencies, and shall attempt a quantitative assessment of the main parameter characterizing I.L.S. co-channel compatibility, namely the time percentage of significant noise from this source of interference. It is appreciated, however, that the final certification of any given I.L.S. installation will have to rely on the information gathered *in loco* by a suitably designed long-term performance monitor.

8.1. *Scope and Limitation of Assessment*

The number of parameters influencing I.L.S. co-channel performance, whether of the wanted and unwanted transmission, the ground and airborne equipment, or the ambient conditions, is very large. Of each parameter we may consider either a mean value, or its worst possible value, or a full probability

Table 4

Measured and theoretical course deviations attributable to the B.E.A. Engineering Base

Date of tests 16th May 1965; Aircraft height 50 ft; Aircraft speeds, see Table 1.

Aircraft distance from localizer	Comments	Measured peak-to-peak amplitude,† corrected for recorder frequency response	Corresponding course-line deviation, obtained by relating Figs. 9 and 10			Theoretical course-line deviation (see Fig. 8)		
			Left-right indicator current, μA	D.D.M.%	Disturbance factor, γ	Left-right indicator current, μA	D.D.M.%	Disturbance factor, γ
10200 ft	Start of coherent pattern	22·5 mm	± 30	± 3·1	0·15	± 3	± 0·3	0·015
9600 ft	Region of maximum disturbance	65 mm	± 86	± 8·9	0·44	± 21·5	± 2·2	0·11
9000 ft	Geometrical reflection point	40 mm	± 53	± 5·5	0·27	± 17·5	± 1·8	0·09
8000 ft	End of coherent pattern	8 mm	± 11	± 1·1	0·06	± 9·5	± 1·0	0·05

† Two-second-mean values averaged over 4 runs.

distribution. An analysis in terms of mean values alone or worst values alone would be clearly misleading. An assessment in the most general terms using for each parameter an exhaustive statistical characterization by means of appropriate statistical distributions seems hardly practicable, and its results would remain of dubious interpretation in any given situation. With a view to gaining useful insight into the nature of the problem and an appreciation of the magnitude of the effects to be expected, the approach followed has been to confine the full statistical characterization to only some of the parameters and to use for others either typical or worst-case conditions as detailed below.

(a) The wanted transmission is assumed to be radiated by an aerial 7 ft high and to be received by a landing aircraft 200 ft high at a distance of 13 000 ft from the wanted I.L.S. This is the approximate height of a landing aircraft at the middle-marker position and is representative of conditions during the 20-seconds period, 30 to 10 seconds before touchdown. It is considered that the assessment of interference at this stage of landing will be of significance if the azimuth guidance is disconnected 5 seconds before touchdown,[1] but it is appreciated that different blind-landing strategies may well shift the emphasis towards lower heights. The distance of 13 000 ft is typical of an average installation.

(b) The co-channel interference is assumed to originate at a distance of 140 statute miles which is the minimum allocated separation in the existing I.L.S. network in Western Europe. The landing aircraft is assumed within the main beam of the interfering station. In other words, we are assuming the worst possible case both with respect to distance and azimuth.

(c) The propagation conditions assumed will be those of an average land path in the U.K. Estimates of levels and time percentages of anomalous propagation events are liable to very large errors, which it is hoped to reduce by confining our assessment to the conditions stated.

8.2. *The Fundamental Parameter*

The parameter to be specifically assessed here is the time percentage of significant noise. Significant noise arises in an I.L.S. receiver-autopilot system when three conditions are met: (1) the interference level exceeds a threshold, (2) the interference spectrum contains a line within some specified band of the wanted I.L.S. spectrum, and (3) the rate of change of the frequency difference d is less than 0·1 Hz per second.[1]

There are several discrete bands of the wanted spectrum that need protection from interference (noise), and the probability P_s of significant noise, emerges as the sum of the products of three probabilities

$$P_s = \sum_{i=1}^{m} P_{si} = \sum_{i=1}^{m} P_{1i} P_{2i} P_{3i}$$

where each factor of the ith product qualifies the probability of the above-listed conditions with regard to the ith noise-susceptible band of the wanted spectrum. For a conventional I.L.S. receiver the required protection ratios from I.L.S. interference are shown in Table 5.

Table 5

Protection ratios required by conventional I.L.S. receiver against I.L.S. interference

Interference type	Frequency difference d, Hz	Protection required, dB
Spectral-line interference	$89·8 < \lvert d\rvert < 90·2$ $149·8 < \lvert d\rvert < 150·2$	46
Spectral-line interference	$\lvert d\rvert < 0·2$ $59·8 < \lvert d\rvert < 60·2$ $239·8 < \lvert d\rvert < 240·2$	38·1
Spectral-line interference	$179·8 < \lvert d\rvert < 180·2$ $299·8 < \lvert d\rvert < 300·2$	24·4
Audio capture	$70 < \lvert d\rvert < 110$ $130 < \lvert d\rvert < 170$	20·9
Audio capture	$\lvert d\rvert < 250$	14

An assessment of the probabilities P_3, P_2 and P_1 is given in this order in the following subsections:

8.2.1. The probability P_3

The frequency stability of a trans-horizon tropospheric signal is very high for most communication purposes. Evidence of this are the very large communication bandwidths achieved in tropo-scatter systems, where bandwidth limitations arise from other mechanisms (e.g. selective fading) than frequency instability of the received signal. However, data from Reference 2 suggest that such instability is indeed small but not zero. These authors have measured (over a 161 mile path at 400 MHz) 0·6 Hz for the standard deviation of the frequency fluctuations of samples integrated over $\frac{1}{8}$ of a second for a 10-minute data-sampling period. They do not quote figures for the instantaneous rate of frequency change, but their plots of received frequency against time suggest an average rate of variation as large as 1 Hz/s in spite of the smoothing introduced in processing. Under enhanced propagation conditions, frequency variability is reduced by approximately a factor of four, and a further reduction by four times would have to be introduced to convert this datum to the frequency of interest (100 MHz). Thus, the average rate of frequency change of an anomalous signal at 100 MHz over a 160 mile path could be estimated to be 0·06 Hz/s. If a figure for the standard deviation of this rate of change were known we could assess the probability for the rate of frequency variation of co-channel interference to be less than 0·1 Hz/s. Although it may well be found that such a probability is less than 1, and although the techniques used for the measurements quoted imply some underestimation of the parameter (in that measurements are carried out on a two-way path thus cancelling opposite Doppler components[3]), the evidence to date does not entitle us to claim benefit from any frequency variation—desirable in our context—of the interference. The probability P_3 will be assumed, in this paper, to be equal to 1.

8.2.2. The probability P_2

The assessment of the probability of the unwanted carrier occurring within a frequency interval of width d about some line of the wanted spectrum is in many ways elusive. It arises from the inter-play of many factors, some of which are in the nature of probability densities while others bear the connotation of statements of ignorance.

Its assessment may be attempted as follows. I.L.S. transmissions are prescribed to be stable within $\pm 5{\cdot}5$ kHz. Some commercial equipment achieves this without a crystal oven by means of crystals having manufacturing tolerances of 20×10^{-6}, i.e. $\pm 2{\cdot}2$ kHz carrier frequency spread. Drift due to ageing is about 5×10^{-6} (= 0·5 kHz). A further ± 3 kHz variation will occur for temperature variations from -20 to $+70$°C. We cannot claim the full benefit from this last source of variability, since the temperatures at two stations 140 miles apart may differ little. The two crystals, however, will have as a rule different frequency vs. temperature curves. It will be assumed on balance that at least one-tenth of their frequency spread can be relied upon to reduce the chance of the wanted/unwanted frequency difference being less than a specified value.

Summing up, the probability P_2 for the frequency of a co-channel interfering signal to occur within a band of width d Hz of the spectrum of the wanted I.L.S. signal, may be assumed to be less than $d/(6 \times 10^3)$.

For the bands of the wanted spectrum listed in Table 5 where protection is required, we obtain:

$$P_{2,1} = 0{\cdot}27 \times 10^{-3}, \qquad P_{2,2} = 0{\cdot}33 \times 10^{-3},$$
$$P_{2,3} = 0{\cdot}27 \times 10^{-3}, \qquad P_{2,4} = 26{\cdot}7 \times 10^{-3},$$
$$P_{2,5} = 83{\cdot}3 \times 10^{-3}.$$

It may be mentioned in this connection that an often-proposed means of reducing the incidence of co-channel interference would be to do away with frequency stabilization and to engineer the carrier to sweep at a suitably slow rate over a fraction, say one-half, of the allocated channel bandwidth (100 kHz). Then the corresponding value of P_2 would be $(d/50) \times 10^{-3}$, i.e. P_2 would be reduced by a factor of 10 with respect to the above assessment, the estimate being no longer a statement of ignorance but a genuine probability density. Higher rewards in terms of reduced values of P_3 (see previous section) would accrue if the rate of sweep were made different for different stations. This may well prove sufficient to relieve the propagation expert from the difficult task of predicting anomalous propagation in the inadequately explored 10^{-7} region. This is not yet the case, however, and we must proceed to assess our last probability factor.

8.2.3. The probability P_1

This is the probability for the wanted/unwanted voltage ratio to be less than certain specified threshold values. What is needed is the predicted probability distribution for this ratio, and, hence, individual probability distributions for numerator and denominator.

(a) *Wanted signal.* On the assumption of a flat propagation path for the wanted signal, and for the distance (13 000 ft) and height (200 ft) considered, standard line-of-sight propagation formulae give a wanted signal level of 16·4 dB below free space, i.e. 92·8–16·4 = 76·4 dB rel. 1µV/m = 69·2 dB rel. 1µV in 73 ohms, for 1 kW effective radiated power. This is a median value about which the signal may be expected to spread on three distinct scores: (1) variability in time owing to changes in propagation conditions and (2) in ground equipment parameters (power output and percentage modulation instability), and (3) site-geometry variations from one I.L.S. installation to another. On the first score, quick calculations indicate that even extreme variations in the atmospheric refractive-index lapse would cause variations of only ± 3 ft, i.e. $\pm 1{\cdot}5\%$ of the receiver effective height and a corresponding variation in signal strength. Other fading mechanisms, conceivable in theory, are unlikely to be significant at this frequency except at the extreme tail of the distribution. Ground equipment changes do not come strictly within the scope of our analysis of the electromagnetic environment of I.L.S. They will occur, however, and we may do well to record that the expected change does not exceed 2 dB.

Altogether, it would appear that a log-normal distribution with, say, 3 dB standard deviation should provide ample allowance for the variability in time (factors (1) and (2)) of the wanted signal over an ideally flat propagation path. Since, as will be seen below, the standard deviation of the interference predicted is much higher (13–15 dB), the latter influence will be dominant in deciding the behaviour of the wanted/unwanted signal-ratio distribution. Thus, as far as factors (1) and (2) are concerned, the wanted/unwanted distribution may be taken to be, to all practical accuracy, equal to the interference distribution suitably scaled. Matters stand differently for factor (3). The variability of wanted signal due to site-geometry variations from one I.L.S. installation to another does not lend itself to a statistical characterization. Changes can arise in this context from the use of humped runways (e.g. Bedford, Manchester, Saling), different length of runways, and thirdly (an aspect which is still under study) the use of different surface material (concrete, etc.). We shall renounce any attempt to deal with this factor in a statistical framework (which would be hardly meaningful anyway), and shall follow the worst-case approach by using the proposed specifications given for beam calibration of a Category III localizer in Reference 1. The minimum signal strength specified during final approach is 200 μV. In the absence of more detailed specification, this minimum is assumed to correspond to an aircraft 50 ft high, 10 000 ft from the localizer, and the corresponding value at the height and distance of interest (200 ft, 13 000 ft) will be 475 μV = 53·5 dB rel. 1 μV = 60·8 dB rel. 1 μV/m.

(b) *Unwanted signal.* There is no room here for even the briefest account of the physical mechanisms responsible for the arising of enhanced propagation conditions. They range from abnormally large refractive-index vertical lapses and ground-based ducts to local refractive-index inhomogeneity clouds, thin layers of limited horizontal extent (*feuillets*), elevated ducts and well-defined thin layers capable of supporting specular reflections. Each of these mechanisms lends itself to suitable idealizations and the calculation of their effect is well within the scope of the art, as is evidenced by the wealth of literature in this field over the last two decades. An empirical element will have to be introduced in any case to describe the frequency of occurrence of the corresponding meteorological features, so that the practical prediction has direct recourse to empirical field strength statistics. These may be obtained from C.C.I.R. documents (e.g. Recommendation 370, Geneva, 1963). For our purposes a preferred source of such statistics is the very extensive series of propagation measurements carried out by the B.B.C. during recent years, in particular the results of field strength measurements over land paths in the British Isles. These should provide a firmer estimate in that they are restricted to the particular climatic region of interest; they provide, moreover, estimates for the higher percentile range needed. In Reference 4 the mass of B.B.C. propagation measurements is summarized in the form of propagation curves according to the following analytical expression

$$20 \log E \text{ dB rel. } 1\,\mu\text{V/m} = 20 \log k - 20m \log D$$

where E is the signal level, in the units stated, exceeded for a given time percentage over a distance D (in km) for 1 kW effective radiated power, aerial heights 300 m and 10 m, the quantities 20 log k and m being given as a function of the time percentage by the following Table:

	%	20 log k	m
	99	233·8	5·312
	90	221·7	4·915
	50	194·7	4·165
	10	154·6	3·047
	1	103·4	1·608
	0·1	96·3	1·252
by extrapolation	0·01	101·0	1·180
	0·001	104·0	1·110
	free space	104·8	1

To apply this formula to the case of interest with 200 ft and 7 ft aerial heights, the distance D must be increased by 39 km to allow for the increase in obstructed-path length connected with lowering one height from 300 m to 200 ft, while 12·6 dB must be subtracted from 20 log k to allow for a linear height-gain factor from 10 m to 7 ft.

(c) *Wanted/unwanted ratio.* The translation of the above distribution to one expressing the probability of failure of corresponding levels of protection is obtained by subtracting the estimate thus obtained for the interference level from 60·8, i.e. from the previous estimate of the wanted signal:

$$S/N = 60{\cdot}8 - [20m \log (D+39) - 12{\cdot}6] \quad \text{dB}$$

This has been calculated for D = 225·3 km (= 140 statute miles) and is shown in Fig. 13, curve (a) giving the protection from co-channel interference as a function of the time percentage for which it is expected to fail, for a co-channel separation of 140 miles and for an average overland interference path under the conditions stated. We must bear in mind, however, that if there are a number of localizers installed at various airports in this country, each sharing a channel with another localizer at, say, 140 miles distance, the conditions obtaining for any pair of wanted and unwanted I.L.S. will differ from the predicted average conditions of curve (a). This spread may be described by a log-normal distribution with 8 dB S.D. (see Reference 4 as well as Fig. 4 of C.C.I.R. Recommendation 370) and may be combined with the temporal statistics of curve (a) to give a time-and-location distribution shown by curve (b). The use of curve (b) implies a shift of emphasis from the assessment of a single average pair of co-channel I.L.S. stations to the assessment of a large number of such (not necessarily average) pairs.

8.3. *The Probability of Significant Noise, and Conclusions*

The final arithmetic is straightforward. We read off, from Fig. 13, the values of $P_{1,i}$ corresponding to the above-listed protection levels required. Together with the values previously obtained for $P_{2,i}$, and for $P_{3,i} = 1$, they give

$$P_s = \sum P_{s,i} = \sum P_{1,i} P_{2,i} P_{3,i} = 0{\cdot}9 \times 10^{-5}$$

A similar calculation for $D = 200$ statute miles gives $P_s = 0{\cdot}4 \times 10^{-5}$.

These values may be compared with the target specified in Reference 1, Table 5, namely $1{\cdot}5 \times 10^{-5}$, which is relevant to an airborne-monitored I.L.S., showing that, while a separation of 140 miles comes just within the target, a 200-mile separation gives a predicted time percentage of significant noise appreciably smaller than the target and should prove adequate.

There is no room here for a critical discussion of the assumptions made in arriving at the conclusion just stated. It may be mentioned, however, that calculations following somewhat different approaches suggest that the final figure for the probability of significant noise does not depend critically on the method used. A safety margin of 1 : 3 in the predicted probability should provide ample allowance for this source of uncertainty. For $D = 200$ statute miles, P_s would still be smaller than $1{\cdot}5 \times 10^{-5}$.

A more serious limitation derives from the very definition of the $1{\cdot}5 \times 10^{-5}$ target. This target should be interpreted[1] as a maximum rate of diversions (or alternative courses of action), following a warning by

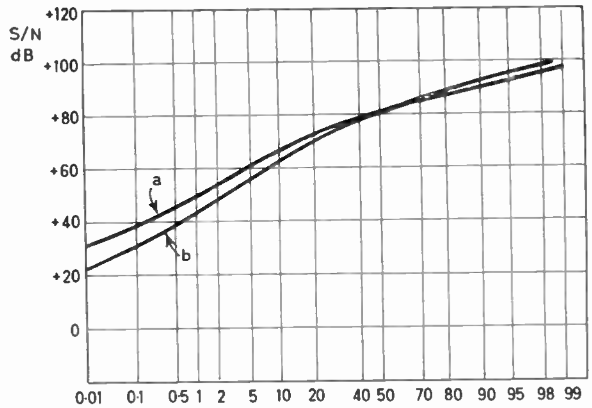

Fig. 13. Time (or time-and-location) percentage for which indicated signal/noise ratios are not achieved.

(a) Predicted distribution of signal/noise ratio for an average 140-mile overland interference path.

(b) Predicted 'time-and-location' distribution of signal/noise ratio for a number of 140-mile overland interference paths.

a suitable airborne monitor or predictor that a dangerous level of interference is being exceeded within specified bands of the wanted spectrum. Thus the equipment will have not only to monitor (or predict) the interference level (and the associated probability, P_1, can be as large as 10^{-2}) but also its frequency within very narrow limits. The latter task may prove too difficult to engineer, and the rate of false alarms could become intolerable.

For an unmonitored I.L.S. the target rate of significant noise from co-channel interference[1] is 6×10^{-9}. It appears that, even if we were prepared to waive some of the pessimistic assumptions made, the target for an unmonitored I.L.S. cannot be met, and alternative solutions other than co-channel operation should be investigated.

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Proceedings of the Symposium on Monitoring of I.L.S. Ground Equipment for Automatic Landing

The following papers, and the discussion on all six papers read at the Symposium, will be published in subsequent issues of *The Radio and Electronic Engineer*:

'An Experimental I.L.S. Echo-Monitoring System', J. G. Flounders.

'New Precision Techniques for I.L.S. Parameter Measurement', G. K. Lunn.

'I.L S. Transmitter Monitors for Blind Landing', F. G. Ferneau.

'I.L.S. Far-Field and Radio Environment Monitoring', A. M. Raeburn.

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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.

DESIGN METHODS FOR MULTI-SECTION DELAY LINES

A Czech paper draws a critical comparison of design methods for multi-section delay lines, namely the frequency-response method and the transient response method. The frequency-response method is shown to be less accurate, thus leading to incorrect notions about the advantage of *m*-derived networks as compared with constant-*k* networks. The transient response method, on the other hand, gives results closely corresponding to practice and has lately been corroborated by calculations of multisection lines by means of computers.

'Methods of calculating delay lines with lumped parameters', J. Klimeš, *Slaboproudý Obzor*, **27**, No. 10, pp. 605–8, October 1966.

PHASE CONTROL OF OSCILLATOR USING STANDARD TRANSMISSION

The writer of a recent French paper examines the practicability of transferring to a local oscillator the stability of a standard frequency transmission which is emitted intermittently. According to this arrangement the modulation can be removed and a stable continuous carrier wave can be produced locally to allow the regulation of a local primary standard. In this example, the local oscillator is a crystal oscillator of 1 MHz which can be locked to the transmission from Rugby on 16 kHz.

'Design and construction of an oscillator phase controlled by a standard transmission periodically interrupted', G. Thomas, *L'Onde Electrique*, **46**, No. 474, pp. 986–88, September 1966.

MICROWAVE FARADAY EFFECTS IN SEMICONDUCTORS

A quantitative method for studying the microwave Faraday effect in semiconductor specimen of low resistance has been described by the author of a German paper. Two equal parallelepiped-shaped H_{102}-resonators are employed, with coupling apertures cut in their covering surfaces along wall-current lines. The semiconductor wafer is arranged between the opposite cut-out covering surfaces of the two cavities to act as a resonator wall in the coupling aperture. One of the resonators is stationary and excited. Through the semiconductor wafer the microwaves pass to the second resonator which can be rotated. When a static magnetic field is present in the semiconductor, a Faraday rotation of the plane of polarization of the wave and its ellipticity occurs. The measurements of these quantities are carried out by the use of the rotatable resonator.

From the microwave Faraday effect the Hall mobility, and in liquid air even the relaxation time, of the majority carriers in the semiconductor can be determined at room temperature. These measurements were carried out at X-band and showed satisfactory results for semi-conductor wafers of thicknesses ranging from two to five times the penetration depth.

'Measurements of the microwave Faraday effect in low-resistance semiconductors', F. Seifert, *Archiv der Elektrischen Übertragung*, **20**, No. 3, pp. 169–79, March 1966.

VIDEO TELEPHONES

The authors of a German paper describe the long history of the efforts made on the subject of video-telephones and mention experimental equipment made in Germany and other countries. Technical problems relating to video-telephones are discussed including bandwidth requirements, line spacing, frame repetition rate, and the possibilities of reducing the video-bandwidth. A further investigation deals with the suitability of modern telephone exchange equipment and transmission networks for the application of video-telephones.

'The history of and technical problems in video-telephones', O. Horner and W. Passon, *Nachrichtentechnische Zeitschrift*, **19**, No. 7, pp. 409–416, July 1966.

RELIABILITY OF ELECTRONIC EQUIPMENT

L'Onde Electrique has published a series of papers concerning reliability of electronic equipment, as follows:

'The reliability of components according to the layout of equipment in use', C. Marcovici, pp. 913–16.

'Reliability of public service installations', F. Ordonneau, pp. 917–20.

'Reliability of components used in standardized electronic equipment at the C.E.A.', C. Guyot, pp. 921–23.

'Failure statistics for components used in equipment for closed circuit television', R. Dely, pp. 924–26.

'The reliability of components in P.T.T. equipment', C. Marcovici, pp. 927–30.

'Considerations concerning the reliability of relay transmitters', C. Leuchtmann, pp. 931–35.

'The reliability of circuits using solid-state devices', M. Régert Monod, pp. 936–44.

'Reliability of components under radiation stress: problems to be solved and some experimental results', A. Blin, A. d'Harcourt, and J. Le Ber, pp. 945–54.

L'Onde Electrique, **46**, No. 474, September 1966.

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An Investigation of Propagation Phase Changes at V.L.F.

By

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AND

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Summary: The phase of a received very-low-frequency signal, recorded as a function of time, has been utilized to investigate (*a*) the frequency off-set between two remotely located atomic frequency standards and (*b*) the variations of the diurnal phase shift—which is a characteristic of the phase comparison records.

A comparison of the precision obtained in the measurement of the frequency off-set by using 20 and 60 kHz transmissions over a path length of 1800 km, reveals that the measurement is more precise on 20 kHz transmissions. The diurnal phase shift, studied from the standpoint of the two-mode model, shows that the second mode interference increases with increasing frequency in the range 16–20 kHz as can be seen from the curves theoretically computed for diurnal phase shift as a function of path distance. Experimental results are presented for comparison. Variations of the general phase pattern and the monthly variation of the diurnal phase shift are also presented. The diurnal phase shift is found to be a minimum during the period of solstice and a maximum during equinox. The general consistency of the phase shift is shown to depend both on path length and its orientation. The value of the change in effective height of ionospheric reflection calculated from the yearly mean diurnal phase shift for a long propagation path is found to be in good agreement with the value used in theoretical calculations.

1. Introduction

It is well known that the phase of a v.l.f. carrier signal, observed at long distances, follows a regular variation throughout the day.[2,9,11,12] This variation is found to exhibit, in general, a standard trapezoidal form. Certain departures from this systematic variation have been reported for conditions of solar flare, magnetic activity, and disturbances such as nuclear explosions in the atmosphere.[3,5,8] The normal phase variation of the v.l.f. signal is caused mainly by the regular and consistent day-time to night-time shift in the level of the ionosphere which forms one wall of the waveguide propagating the signals. In addition to this regular diurnal phase change there is a continuous drift in the phase level of the received signal due to frequency difference between the two atomic frequency standards utilized in monitoring the v.l.f. signal. Any irregular changes superimposed on these systematic variations under normal conditions may be attributed to ionospheric irregularities. A part of the present investigation deals with the precision of the frequency off-set measurement between two remotely located atomic frequency standards by using transmissions on 20 and 60 kHz.

The other part of the investigation is concerned with the variations of the diurnal phase shift as a function of path distance, frequency and time. One important application of the phase shift measurement is in obtaining the change in height of reflection at the D-region level from day to night using the expression given by Wait[14] and the set of relative phase velocity curves of Wait and Spies.[17] The assumptions involved are that the ionosphere is sharply bounded and only the first-order waveguide mode of propagation is present. Hence, reducing phase shift data into corresponding height changes is valid only for propagation paths over which the effect of second and higher order modes is negligible.

In order to assess the validity of the changes in effective height of reflection calculated from the single mode model for different paths, one should know the minimum distance above which the contribution from the higher modes is negligible. This has already been determined, in part, from a study of the variation of the diurnal phase shift, or delay in transmission time, with path distance. Usually, the diurnal phase shift $\Delta\phi$, is expressed in terms of a diurnal time delay, ΔT, where $\omega\Delta T = \Delta\phi$ and ω is the angular

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frequency of the carrier wave. It is known that for distances over which the first-order mode is dominant, variation of phase shift with distance should be linear. Any departure from this variation indicates the possible interference from higher order modes. This aspect has been studied in some detail by Wait,[16] Burgess[6] and Blackband.[4] Their results suggest an oscillatory variation of ΔT for distances where the higher modes influence the resultant phase. However, their investigation is limited to a single frequency of propagation, namely 16 kHz.

In the present investigation the time delay has been studied experimentally at frequencies 16, 18 and 20 kHz; theoretical curves for ΔT versus the propagation distance d, were also computed for the three frequencies to determine how the variation changes with frequency. From this type of study, it is possible to determine, as a function of frequency, the path distance over which the higher modes are effective. The monthly variation of ΔT, over a period of a year, was also investigated.

2. Frequency Comparison Measurements by Transmission on 60 kHz and 20 kHz

Recently, two standard frequency stations, one operating on 20 kHz (WWVL), the other on 60 kHz (WWVB), were put in operation by the National Bureau of Standards at Fort Collins, Colorado. Both transmitters are located on the same site, but they are completely separate systems except for an atomic frequency standard, which is the frequency-controlling device common to both. Signals from both these stations were recorded simultaneously in Columbus, Ohio. The purpose of comparing the signals received at each frequency in Columbus against the local rubidium standard, Rb-3, was to compare the precision of measuring the off-set between the Boulder and Columbus standards on two separate frequencies.

The usual method was used to determine the frequency difference between the standards by measuring the average change in phase per second.[7, 12] In order to determine this value for any particular period of time, the best straight line, in the sense of least square-error, was fitted to the phase-versus-time records. Hourly values of the relative phase were taken from the records for analysis of the period of interest. Rather than using the standard deviation of the phase variations as the means to calculate precision of the off-set measurement, use was made of statistical methods. It can be shown that for a process which is essentially random and normally distributed, some measure of the precision of the slope of a best-fit line can be obtained through the use of Student's distribution.[1, 13] Two variables are formed for a set of data; one is distributed as chi-squared, the other is normally distributed. The ratio of these two variables turns out to be just the definition of the t-distribution of Student for $n-2$ degrees of freedom:

$$t_{n-2} = \frac{\Delta b\sqrt{(n-2)\sum(x-\bar{x})^2}}{\sqrt{\sum(y_i-\bar{y}_i)^2}},$$

where Δb = variation in the slope,
x = independent variable (time),
$\bar{x}$ = mean of x,
y_i = phase at any observation time,
$\bar{y}_i$ = phase from best-fit line, and
n = number of observations.

Thus, by entering a table of Student's distribution for the number of degrees of freedom (observations) involved and reading the corresponding value of t for some specified tolerance level, the confidence limits on the slope can be found.

For example, if 1% symmetrical limits are placed on the off-set determination, one can obtain from the table the value of t associated with any particular slope calculation. This, in turn, can be substituted into the expression above, which can then be solved for Δb, the tolerance of the slope about the value determined from the least-square-error procedure. The statistical meaning of these limits is that if one assumes, *a priori*, that the true slope exists within the interval, the assumption will be correct 99% of the time; it gives a 99% guarantee that the true slope lies within the limits defined above.

The afore-mentioned procedure was applied to records taken during a number of intervals on both 20 kHz and 60 kHz over the three-month period, October, November and December 1963. The results are summarized in Table 1, where it can be seen that the average precision of measurement at 20 kHz is better than at 60 kHz for long measurement times. The reason for this seems indicated in the variance

Table 1

Average precision for several measurement times

Measurement time (hours)	Frequency kHz	Average precision $\times 10^{-12}$	σ^2 $(\mu s)^2$
5	20	25·0	0·5
5	60	11·0	0·2
8	20	10·0	0·4
8	60	8·0	0·6
12	20	4·0	0·4
12	60	6·0	0·8
20	20	1·5	0·5
20	60	3·0	0·7
30	20	0·2	0·3
30	60	0·3	0·7

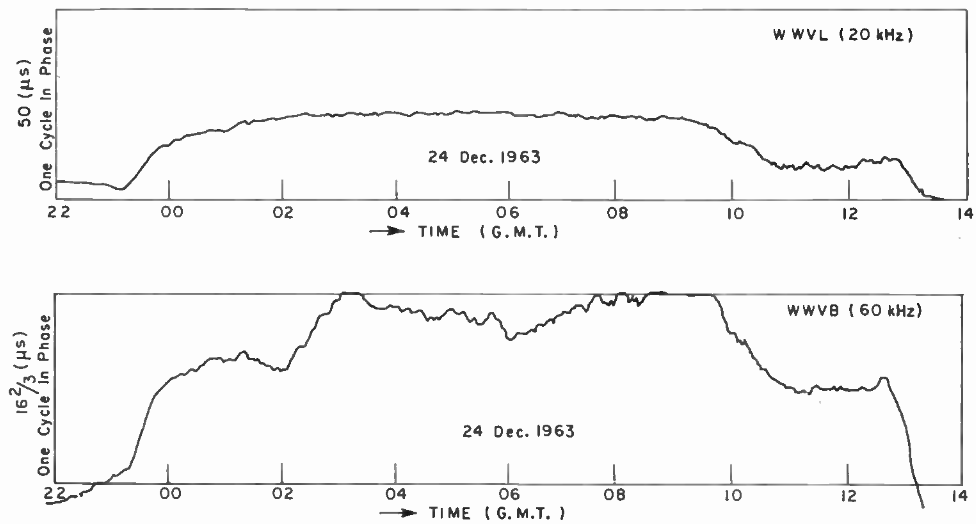

Fig. 1. Typical phase recordings of 20 and 60 kHz signals.

of phase fluctuations at each frequency. Consistently higher phase variance was obtained at 60 kHz, indicating that phase variations at this frequency were greater than at 20 kHz (see Fig. 1), thereby reducing the precision of frequency measurement.

It appears from these results that for distances commensurate with the Fort Collins–Columbus path more precise measurements of frequency difference between standards can be achieved at the lower frequency. This is not to say that a general statement can be made as to a preferred frequency since only one particular distance and path is represented by these measurements. It is interesting to note that for the shortest measurement time (five hours), 60 kHz gave higher average precision. Pierce[12] has found this to be generally true for a distance of some 5200 km when using phase variance to determine precision on 16 and 60 kHz transmissions. Watt *et al.*[19] also confirmed this, but their procedure of determining precision was different.

3. Variations in the Diurnal Phase Change of V.L.F. Transmissions

3.1. *Calculation of ΔT, the Diurnal Time Delay*

We shall first present the method of estimating ΔT, assuming a single waveguide mode of propagation, and then treat a more general case by adding the second mode.

Consider the expression for transmission time for a single mode,

$$T = \frac{d}{V} \qquad \ldots\ldots(1)$$

where T = transmission time,

V = phase velocity of the wave,

d = propagation distance.

For two different guide dimensions, two separate velocities, V_1 and V_2, are found. Thus, the transmission time delay for a given mode will be

$$\Delta T = (d/V_2 - d/V_1).$$

Neglecting the second mode,

$$\Delta T = \left(\frac{d}{V_2} - \frac{d}{V_1}\right) = d\left\{\frac{(V_1/c-1)-(V_2/c-1)}{V_1 V_2/c}\right\} \qquad \ldots\ldots(2)$$

Define

$$\delta_1 = \left(\frac{V_1}{c} - 1\right); \qquad \delta_2 = \left(\frac{V_2}{c} - 1\right)$$

so that

$$V_1 = (\delta_1 + 1)c \quad \text{and} \quad V_2 = (\delta_2 + 1)c$$

Hence

$$\Delta T = \frac{d}{c}\left\{\frac{\delta_1 - \delta_2}{(\delta_1+1)(\delta_2+1)}\right\} \qquad \ldots\ldots(3)$$

If δ_1 and δ_2 are much less than unity a good approximation to the above equation is

$$\Delta T = \frac{d}{c}(\delta_1 - \delta_2) \qquad \ldots\ldots(4)$$

From this expression ΔT can be estimated for a single waveguide mode of propagation, provided the values for δ_1 and δ_2 are known. Usually, δ_1 and δ_2 are taken from a set of curves, such as given by Wait and Spies,[17] of relative phase velocity versus frequency for different ionospheric heights. From a number of earlier investigations, a day-time height for v.l.f. of 70 km and a night-time height of 88 km seem to be the most reasonable for calculation.

If the second mode is also considered, the computation becomes a little more complicated. To deal with this case, the following method of analysis suggested by Knox[10] has been adopted in the present investigation. First one finds, for the first mode, the time delay relative to free-space propagation for the two ionospheric heights. Thus, for one mode and an ionospheric height of 70 km,

$$(\Delta T)_1 = \delta_1 \frac{d}{c} \qquad \ldots\ldots(5)$$

and for an ionospheric height of 88 km

$$(\Delta T)'_1 = \delta'_1 \frac{d}{c} \qquad \ldots\ldots(6)$$

where the subscript denotes the mode number and the prime indicates the 88 km height in the preceding and following equations.

The corresponding relative electric field magnitudes are found by the use of the attenuation rates which are given by Wait and Spies for different ionospheric heights. They can be represented as

$$E_1 = E_0 \operatorname{antilog} \frac{1}{20}\left[\frac{A_1 \times d}{1000}\right] = E_0 a_1 \qquad \ldots\ldots(7)$$

and

$$E'_1 = E_0 \operatorname{antilog} \frac{1}{20}\left[\frac{A'_1 \times d}{1000}\right] = E_0 a'_1 \qquad \ldots\ldots(8)$$

where E_0 is the magnitude of the unattenuated wave,

A_1 and A'_1 are the attenuation rates in decibels per 1000 km, and

d is the Great Circle propagation distance in km.

Converting the time delay, ΔT, in seconds into an equivalent phase shift in radians, one has

$$\Delta\phi = (2\pi f)\Delta T \qquad \ldots\ldots(9)$$

where f is the frequency of the v.l.f. signal.
Now, one can express the field for the first mode as a phasor for each ionospheric height as follows:

$$\varepsilon_1 = E_1 \angle \Delta\phi_1 = E_0 a_1 \angle \Delta\phi_1 \qquad \ldots\ldots(10)$$

and

$$\varepsilon'_1 = E'_1 \angle \Delta\phi'_1 = E_0 a'_1 \angle \Delta\phi'_1 \qquad \ldots\ldots(11)$$

The same procedure is followed in deriving corresponding field expressions for the second mode. Thus for the second mode the fields corresponding to the two ionospheric heights are

$$\varepsilon_2 = E_0 a_2 \angle \Delta\phi_2 \qquad \ldots\ldots(12)$$

and ....

$$\varepsilon'_2 = E_0 a'_2 \angle \Delta\phi'_2 \qquad \ldots\ldots(13)$$

The resulting phasors for the two modes are added to obtain the phase of the total field corresponding to each ionospheric height. This is done by combining eqns. (10) and (12) for 70 km yielding

$$\begin{aligned}\varepsilon_{70\,\mathrm{km}} &= \varepsilon_1 + \varepsilon_2 = E_0 a_1 \angle \Delta\phi_1 + E_0 a_2 \angle \Delta\phi_2 \\ &= E_0(a_1 \cos\Delta\phi_1 + ja_1 \sin\Delta\phi_1 + \\ &\qquad + a_2 \cos\Delta\phi_2 + ja_2 \sin\Delta\phi_2)\end{aligned}$$

Denoting the real and imaginary parts by x and y, respectively, we have

$$\varepsilon_{70\,\mathrm{km}} = E_0(x + jy) \qquad \ldots\ldots(15)$$

Similarly, by combining eqns. (11) and (13), the field for 88 km is expressed as

$$\begin{aligned}\varepsilon_{88\,\mathrm{km}} &= E_0(a'_1 \cos\Delta\phi'_1 + ja'_1 \sin\Delta\phi'_1 + \\ &\qquad + a'_2 \cos\Delta\phi'_2 + ja'_2 \sin\Delta\phi'_2) \\ &= E_0(x' + jy') \qquad \ldots\ldots(16)\end{aligned}$$

The corresponding phases of the field vectors, with respect to a free-space signal, are given as

$$\theta = \tan^{-1}(y/x) \qquad \ldots\ldots(17)$$

and

$$\theta' = \tan^{-1}(y'/x') \qquad \ldots\ldots(18)$$

From these the phase shift in the signal resulting from the height change, Δh, is given by

$$\Delta\phi = \theta - \theta' \qquad \ldots\ldots(19)$$

This phase shift, in radians, can then be converted back into an equivalent time delay, using the relation

$$\Delta T = \frac{\Delta\phi}{2\pi f} = \frac{\theta - \theta'}{2\pi f} \qquad \ldots\ldots(20)$$

3.2. *Results*

Using eqn. (20), ΔT values have been computed for path distances ranging from 1000 km to 10 000 km at intervals of 500 km for frequencies of 16, 18 and 20 kHz. In this evaluation, the day-time and night-time heights of reflection are assumed to be 70 km and 88 km, respectively, for all the three frequencies. The curves showing the relation between ΔT and propagation distance d are presented in Figs. 2(a), (b) and (c) for 16, 18 and 20 kHz transmissions, respectively. The dotted straight lines, showing $\Delta T/d$ as constant in the three figures, represent the variation of ΔT with distance under a single mode of propagation as computed by eqn. (4). With the second mode taken into consideration, the variation becomes oscillatory in nature about the constant $\Delta T/d$ relationship, as is apparent from the curves. The fact that the second mode effect decreases with distance is evident, assuming that the amplitude of oscillation is a measure of second-mode interference. It is obvious from the curves that this interference increases with frequency in the range 16–20 kHz. The distance over which the second-mode interference is significant is also found to increase rapidly with frequency; for example, it is about 5000 km at 16 kHz, 6000 km at 18 kHz, and more than 8000 km at 20 kHz. The fact that the second-mode effect decreases less rapidly with distance as frequency increases may be observed from Figs. 2(a), (b) and (c), revealing lesser attenuation rate of the variation with increasing frequency. This result is in accordance with the frequency dependence of the attenuation rates for the second mode reported by Watt and Croghan.[18]

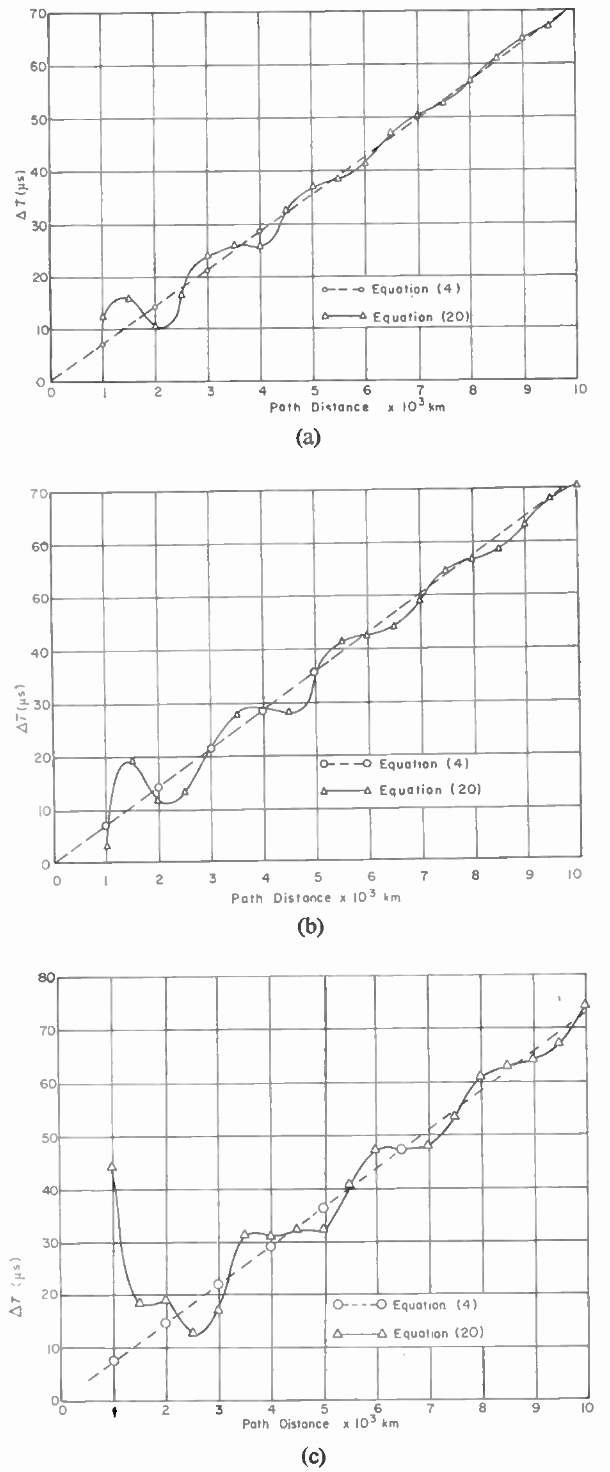

Fig. 2. Diurnal time delay ΔT as a function of path distance for *(a)* 16 kHz, *(b)* 18 kHz and *(c)* 20 kHz transmissions.

Other interesting features of these curves are the several points of intersection of the dotted straight line with the curve representing the total delay, including the second mode. For discrete distances, corresponding to these intersections, the phase shift is the same whether one considers one or two modes.

3.3. *Comparison with Experimental Results*

A comparison has been made between experimental data and the theoretically deduced values of Figs. 2(a), (b) and (c). The experimental data used in this connection consist of measurements for three paths terminating at Columbus, Ohio, and the data listed by Westfall.[20] In this investigation, only data related to 16, 18 and 20 kHz are considered. These values are presented in Table 2, along with predicted values scaled from Figs. 2(a), (b) and (c) for path distances corresponding to the experimental investigation. The theoretical and experimental data are also presented graphically in Fig. 3. Making allowances for the

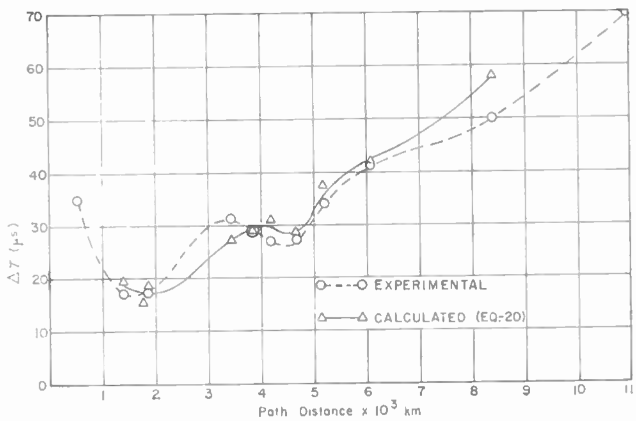

Fig. 3. Diurnal time delay ΔT as a function of path distance.

variation in composition of the paths (i.e. part land and the remainder sea) and for the accuracy with which these measurements could be made experimentally, the agreement between the experimental and theoretical values appears to be good. This agreement could be taken as a good support of the assumptions involved in the theoretical calculations. The difference in the ground conductivity for the paths listed in Table 2 (100% or 0% of land) does not seem greatly to affect the propagation time.

A consideration of 10·2 and 11·2 kHz data listed by Westfall, which were not presented here, suggests the great influence frequency has on the transmission time. Wait[15] pointed out that a more elaborate model is required to take account of the effect that 10·2 and 11·2 kHz transmissions have yielded somewhat greater height increments than those deduced for 16, 18 and 19·8 kHz.

4. V.L.F. Phase Variations with Time

4.1. *A Study of General Phase Pattern*

The phase of a received v.l.f. signal, with respect to a frequency standard, varies in accordance with the day-light-to-darkness changes of the phase velocity of the wave. The pattern of diurnal variation of this relative phase, after accounting for the oscillator vagaries, was shown by Pierce[12] to be trapezoidal for the path of 5200 km from Rugby, England, to Cambridge, Massachusetts. Later, from a series of measurements of the diurnal phase changes for the path Rugby–Malta (2210 km), Blackband[3] observed certain deviations from the standard form given by Pierce. Rather than a neat trapezium, he reported the variation as having a slight reverse drift before dawn, followed by an overswing of phase before the steady daytime value was reached. The phase overswing, he observed, was reported to be most apparent on a path which runs roughly parallel to the dawn or dusk line. Further, it has been said that even if the path follows the dawn or dusk line, the effect of overswing would not be expected to appear at distances so great that only a single mode is received at an appreciable amplitude. From this it appears that both the direction of propagation and the interference of several modes control the overswing of phase. In the light of these earlier observations of the phase variation, diurnal phase recordings available for four different paths—Boulder–Columbus, Balboa–Columbus, Balboa–Boulder and Balboa–Maryland—have been examined in the present investigation. For all these paths the phase variation is found to be remarkably similar, with only a slight reverse-drift of

Table 2

Diurnal time delays for 16 to 20 kHz radio waves over path distances of 535 to 11 000 km

Frequency kHz	Path and distance km	Path composition % land	Diurnal time delay ΔT (μs) measured	standard deviation	calculated
16	Rugby* 535	100	35	—	—
16	Rugby–Cambridge† 5200	0	34	—	37·5
16	Rugby–Columbus‡ 6083	—	41·25	4·9	42·0
16	Rugby–New Zealand§ 11 000	—	70	—	—
18·6	Seattle–San Diego‖ 1780	100	18	—	15·3
18	Panama–Columbus‡ 3455	—	31·27	2·36	27·3
18	Panama–Forestfast, New York¶ 3851	0	29	1·2	29·0
18	Panama–San Diego¶ 4670	90	27	4·7	28·5
18	Panama–Heiku¶ 8417	0	50	6·7	58·0
20	Boulder–San Diego¶ 1400	100	17	3·7	19·5
20	Boulder–Columbus‡ 1860	100	17·28	2·77	18·5
19·8	Haiku–San Diego¶ 4180	0	27	3·0	31·0

* Bain and co-workers.[21]
† Pierce.[11,12]
‡ Data of the present investigation.
§ Crombie, Alan and Newman.[9]
‖ Silkwood.[22]
¶ U.S. Navy Electronics Laboratory (Private Communications).

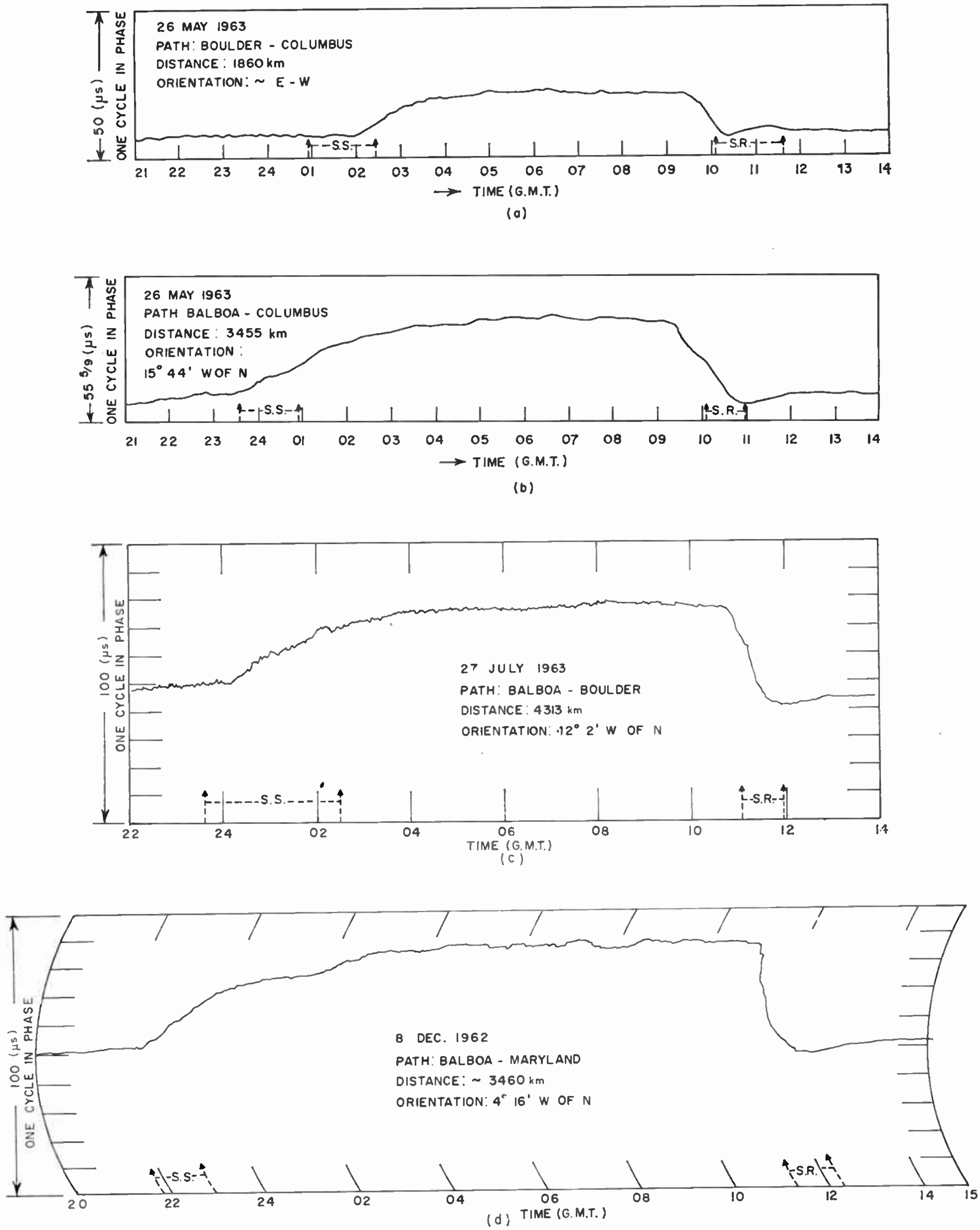

Fig. 4. Typical phase recordings for different paths. (S.R. and S.S. refer to sunrise and sunset times at ground.)

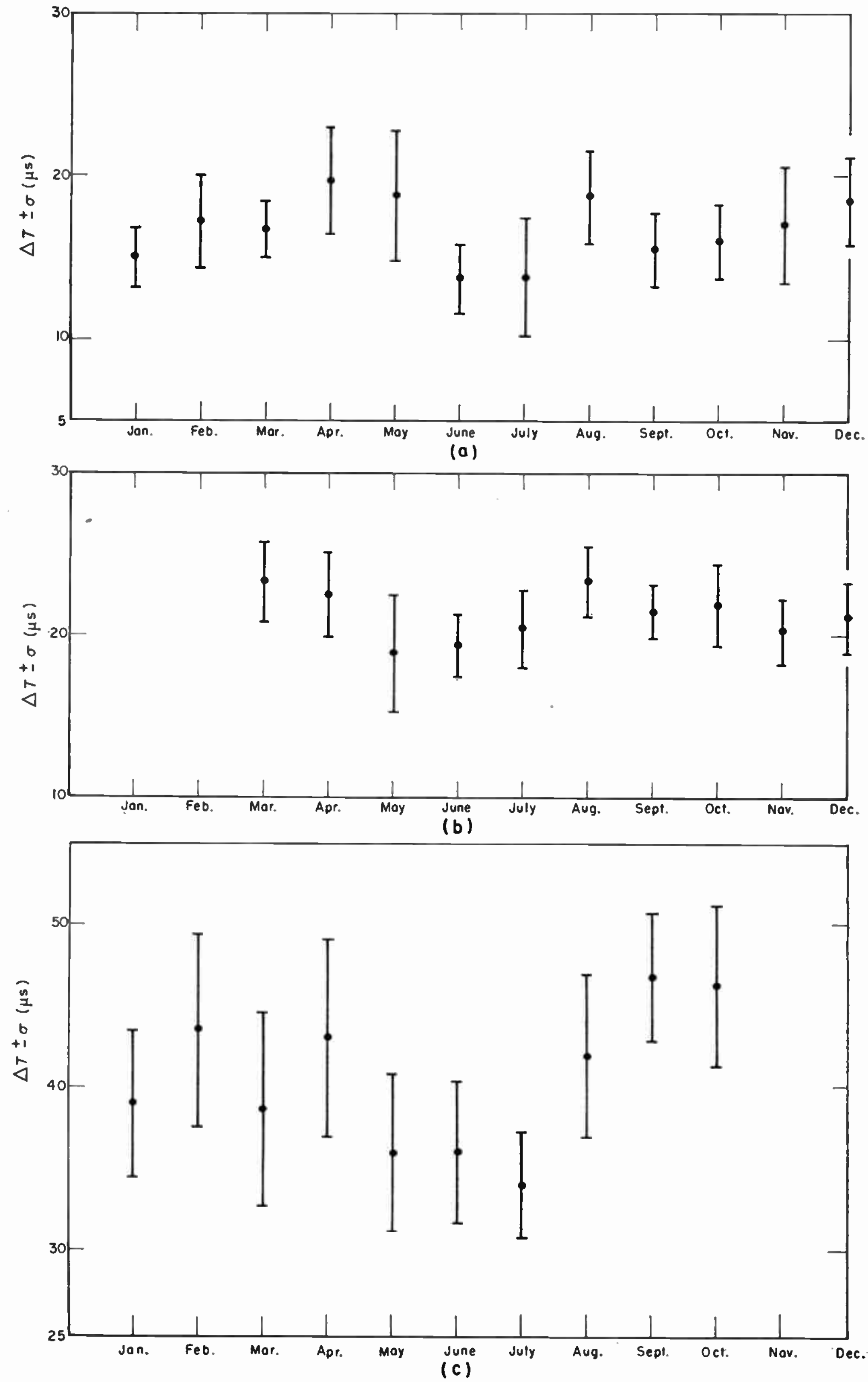

Fig. 5. Monthly variation of mean ΔT for the paths (*a*) Boulder–Columbus, (*b*) Balboa–Columbus and (*c*) Rugby–Columbus.

phase at the time of dawn when the whole path is just about illuminated. This feature is well-illustrated by means of typical runs presented in Fig. 4 for all four paths. The reverse drift in phase noticed on these paths is found to be far less than what could be seen from Blackband's curve for diurnal variation of phase. No overswing of phase could be seen from any of the four diurnal curves for the paths concerned. The direction of propagation does not seem to have any control over the observed reverse drift of phase since it is almost the same for paths running north–south or east–west. The reason for such manifestation seems to be a distinctly different pattern of mode interference at that particular time of day-break, as compared to the steady day- or night-time conditions. The fact that this feature is a result of some kind of higher mode interference is clearly borne out by its absence from the path Rugby–Columbus, over which only a single mode of propagation is appreciable. From the above observations it appears likely that a reverse drift effect in phase is common for almost all paths where the higher mode interference is significant. The absence of phase overswing from our records may indicate that this effect is restricted to paths commensurate with a distance of about 2000 km and direction north–south, as for example, Rugby–Malta.

4.2. *Monthly Variation of* ΔT

It has been pointed out that the accuracy of a navigational aid is dependent on the accuracy with which the transmission times of the v.l.f. signals can be predicted (Pierce[11]). The transmission time of the signals changes in a systematic manner from day-time to night-time because of the change in the phase velocity. The change in the phase velocity from day to night is quite predictable so long as the change in the dimensions of the Earth-ionosphere waveguide remains consistent. One might expect the waveguide characteristics to be subject to seasonal changes because of the changes in the lower ionosphere, such as the level of layer formation and ionization density. In view of this, there should be some significance to a study of the monthly variation of ΔT. Hopefully, a value of the transmission time might then be predicted at any time with reasonable accuracy. Furthermore, since it is possible to deduce day-time-to-night-time change in the reflection level of the ionosphere Δh from the value of ΔT for longer paths, it is possible to determine the seasonal dependence of Δh from the ΔT variation. The present investigation deals with such a study for the three propagation paths concerned, Rugby to Columbus on 16 kHz, Balboa to Columbus on 18 kHz, and Boulder to Columbus on 20 kHz. The data utilized here extend over the period September 1962 through December 1963. Monthly mean values and the standard deviations have been calculated, and the curves depicting the variation of ΔT are presented in Figs. 5 (a), (b) and (c) for the three paths. An inspection of the figures showing the monthly variation reveals certain common features for the three paths. For example, ΔT attains a minimum value during the month of June. The values are found to be generally low during the period of solstice and high for the equinoctial periods. Although there is no point-to-point correspondence in the variation of the three plots, there is still significant similarity between them. The variation observed in the present investigation is in good agreement with the results reported by Chilton *et al.*[8] and Pierce.[12] Pierce reported that the diurnal phase shift has a lower value in June than in September for a path of 5200 km from Rugby–Cambridge. Chilton *et al.* reported a seasonal variation of Δh from approximately 8 km in June to 16 km in October for the path Rugby–Boulder. This Δh variation may be deduced from ΔT variation since the path of their investigation was long enough to make only the first mode of propagation predominant. In the present investigation the deduction of Δh from ΔT is possible only for the path of Rugby–Columbus, since for the other paths higher modes are found to be effective. The Δh derived for this path from yearly mean ΔT is found to be 17 km, which is about the same as the value used for theoretical estimation of ΔT. The height increment, Δh, is found to be minimum, with a value of 14 km, in June, and a maximum, 19 km, in October. Although the individual values are somewhat higher than those of Chilton *et al.*, the seasonal variation is found to be in good agreement with their observation. Considering the variation and also the monthly standard deviations, it can be said that the Balboa–Columbus path has more consistent values for ΔT than the other two. The Rugby–Columbus path is the least consistent of the three in its diurnal phase shift, as is evident from the yearly standard deviations in Table 2. From these observations it appears that the consistency of ΔT depends on both the length and direction of the propagation path. The north–south path is seen to provide better consistency for ΔT than the east–west path. The fluctuation of ΔT increases with path length, as is expected, by reason of the more propagational vagaries for longer paths. The particularly large standard deviation observed in the present investigation for the path Rugby–Columbus may be partly due to the fact that the transmission path traverses portions of the Arctic region.

5. Conclusions

The results of these studies seem to support the following general conclusions:

Measurement of the frequency off-set between two atomic standards by v.l.f. phase comparison can be done more precisely at 20 kHz than at 60 kHz when the distance involved is of the order of 1800 km.

The effect of the second-order mode on any particular propagation path increases with increasing frequency in the 16–20 kHz range. In addition, the distance over which the second mode is significant also increases with frequency. This effect is very pronounced and occurs quite rapidly over a relatively small frequency range.

Variations in the diurnal time delay during a year indicate a minimum value for ΔT during the period of solstice and maximum during the equinox. The general consistency of the diurnal time delay appears to depend both on path length and orientation, the consistency being more for north–south than east–west paths. This result needs to be confirmed by considering a larger number of propagation paths. The value of the change in effective height of ionospheric wave reflection, calculated from the yearly mean diurnal time delay, for a long propagation path agrees well with the value used to estimate the diurnal time delay in other calculations.

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A Multi-Channel Synchronization Monitor for Triggered Spark-Gap Switches

By

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Summary: When a large number of triggered spark-gap switches are used in parallel, as in many thermonuclear fusion experiments, it is essential to detect when one fires late. A further requirement is that any spark-gap which breaks down spontaneously will be indicated, even if other gaps fire in consequence. Both functions are performed by the 40-channel equipment described, lamp indication being used for early and late firing, with a time bracket adjustable from 10 to 50 ns. The system used is insensitive to interference and poor wave-shape.

1. Introduction

Many experiments in controlled thermo-nuclear fusion research use a high-energy pulsed discharge to heat or to compress a plasma. There are a number of such machines at the Culham Laboratory, with energies from a few kilojoules to several megajoules. The electrical energy is usually stored in a bank of low-inductance capacitors at a voltage of a few tens of kilovolts, with discharge currents in the megampere range lasting a few microseconds. After each discharge or 'shot' there is a waiting period of the order of a minute whilst the capacitors are re-charged and the discharge tube possibly flushed and re-filled. The discharge switch is usually a triode or a tetrode air spark-gap, sometimes pressurized (see Fig. 1). The number of capacitors varies from 10 to more than 400, each capacitor having its own spark-gap switch. Owing to the high rate of rise of current, stray inductance must be kept low and this is minimized by having a large number of switches effectively in parallel.

2. Design Requirements

The machine for which the first monitor was designed and built is named *Tarantula*.[1,2] This has forty capacitors, each with a pressurized air spark-gap switch connected by 2 m of coaxial cable to the load assembly, where all cables are joined in parallel. The rise-time to peak current of this machine is about 0·3 μs, and to obtain this very fast rise-time the system was designed to operate at the exceptionally high voltage of 100 kV. Each capacitor is 0·5 μF, and the total energy is 100 kJ.

The monitor has been designed to overcome two problems. Firstly, due to the high capacitor voltage, there is a chance that stray contamination or other faults in a spark-gap could cause a spontaneous or 'spurious' break-down during charging or whilst waiting to fire. In this event, the high surge currents flowing in the unmatched load would probably cause excess voltages to be applied to some of the other gaps and these would also fire, or alternatively back-coupling through the trigger source might occur. It is therefore not sufficient just to know which gap or gaps have fired, but the first one must be identified. This may be only a few tens of nanoseconds before the others, yet the circuits capable of detecting it have to be held in readiness for many seconds.

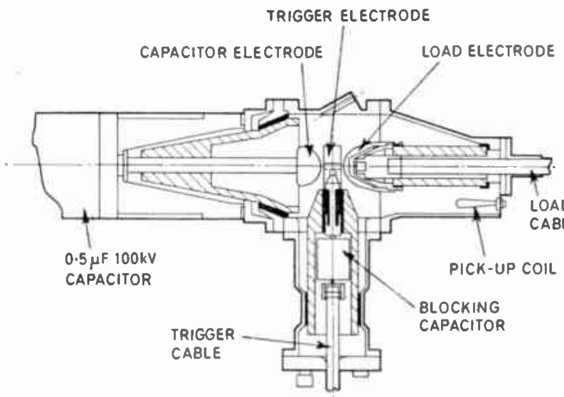

Fig. 1. Section of *Tarantula* spark-gap switch.

Secondly, when the machine is triggered normally, if a gap fails to fire immediately or within a time $2\tau_C + \tau_R$, (where τ_C is the transit time of the cable from the switch to the load and τ_R is the partial rise-time of voltage across the load) it will probably never fire. This is because the voltage applied by the other switches to the load will reach the output electrode of the spark-gap in question, and it will then have insufficient voltage across it to fire. The time $2\tau_C$ is 20 ns and τ_R approximately 10 ns in *Tarantula*. To aid preventive maintenance it is desirable to be able to detect late-firing gaps with a time bracket adjustable from 10 to 50 ns.

The equipment here described performs both functions using the same circuitry.

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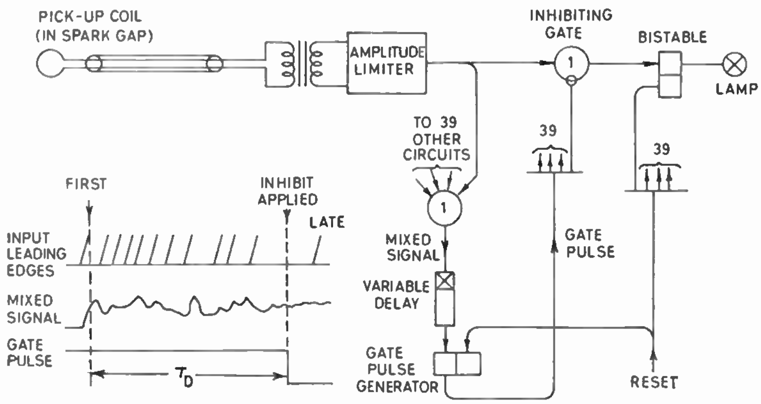

Fig. 2. Block diagram of monitor, with timing sequence.

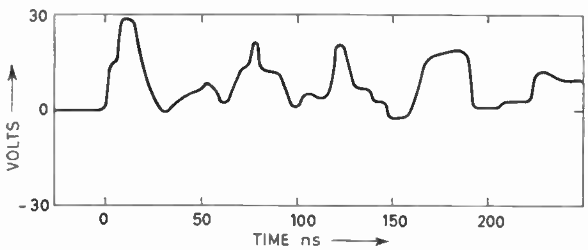

Fig. 3. Typical signal waveform from pick-up coil.

3. Logic Design

One of the most serious problems experienced with these high-power pulsed machines is electrical interference, and as a signal had to be obtained directly from each 100 kV spark-gap, a high noise level was expected. Furthermore, the signal available was only of the order of 10 V in 100 Ω (10^{-1} A) and a time change of 10 ns had to be detected; thus the circuit had to operate on less than 10^{-9} coulomb. For this reason all forms of coincidence circuit, which depend on adding signals, were ruled out. The logic adopted will be called the 'A-before-B' system. 'A' is the input signal, and 'B' is an internally generated step function. Each channel has a trigger circuit which is normally sensitive to input signals, preceded by a gate which blocks when step 'B' occurs. 'B' is generated by a fast avalanche transistor trigger circuit, which derives its input through a diode OR network from any or all of the input signals. This is delayed by an adjustable time τ_D, part of which is the inevitable circuit time delay. Each trigger circuit has a lamp which lights if A arrives before B, and which remains on until reset. Thus, in the case of normal triggering, any spark-gap which fires within time τ_D after the first gap will light its lamp, and faults (late firings) will not. In the case of a spurious break-down, step B will be produced after τ_D. This is less than the triggering time of the other spark-gaps (approximately 100 ns) and therefore only the first break-down will light a lamp.

The important feature of this system is that whatever overshoot, interference or other signal arrives after step B, false indication does not occur. A block diagram is shown in Fig. 2, and a typical signal waveform in Fig. 3.

4. Circuit Details

The circuit diagram of one channel is shown in Fig. 4. A signal of 20–50 V is induced in the small pick-up coil mounted within the spark-gap near the output cable (see Fig. 1) and a balanced cable and screened pulse transformer are used for interference rejection. A 28-channel version, designed for another experiment, is illustrated in Figs. 6 and 7. In this case, a more tightly coupled pick-up coil was available,

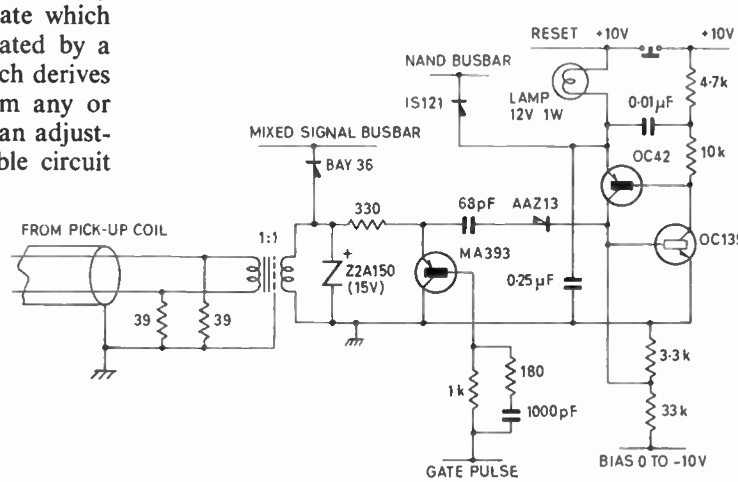

Fig. 4. Circuit of one channel.

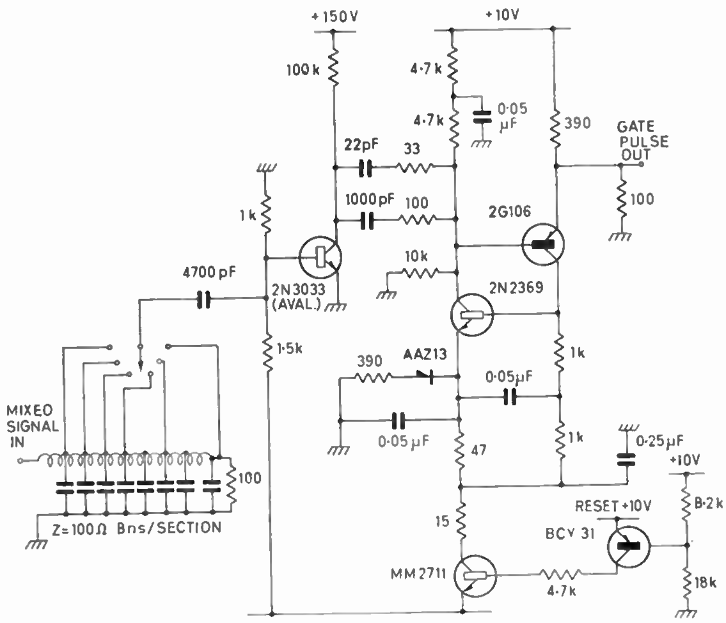

Fig. 5. Circuit of gate-pulse generator.

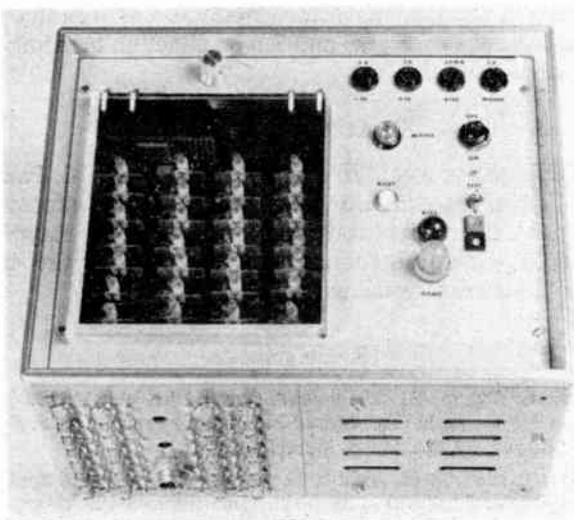

Fig. 6. 28-channel monitor.

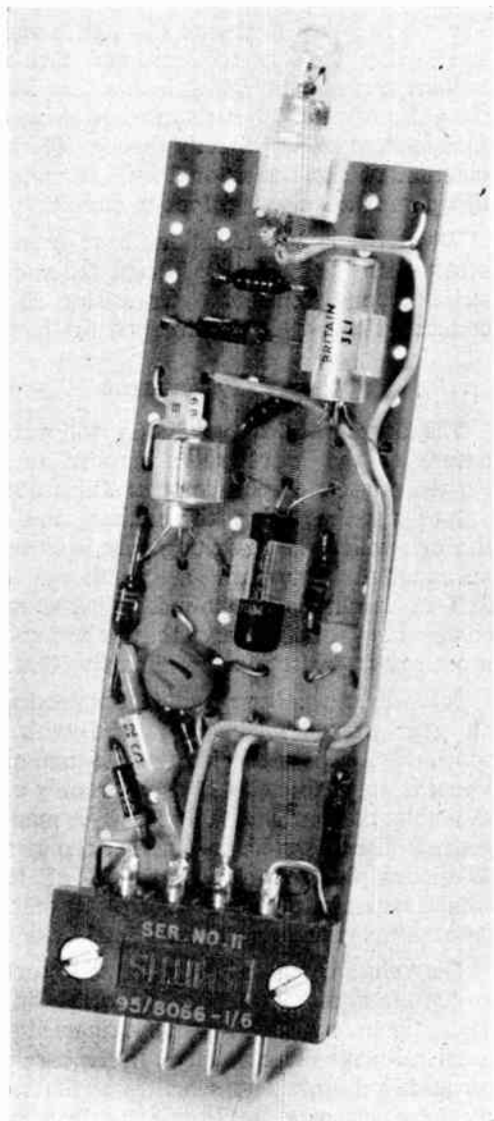

Fig. 7. One channel unit.

generating 500 V in a coaxial instead of a balanced cable, and resistive attenuators were used, not transformers.

The Zener diode has a very satisfactory limiting action in the nanosecond range if a low-capacitance type is chosen, and it also eliminates the negative overshoot. The diode BAY36 takes a portion of each input signal to trigger the gate-pulse generator. The gate transistor MA393 is normally non-conducting with +2 V on its base, and therefore an input signal can pass to the bistable switch which is initially 'off'. When the gate step-function B (−2 V) appears, the gate transistor closes to a low impedance and attenuates the signal. The bistable circuit consists of a p-n-p/n-p-n pair with unity coupling.[3] When triggered, the voltage drop across the pair falls to less than 1 V, and the lamp therefore lights and stays on until the RESET +10 V supply is interrupted. The diode 1S121, in conjunction with those in the other channels, forms a NAND gate, enabling remote indication that one or more lamps are off. Also, earthing this point tests all lamps. Remote indication that one or more lamps are on may be obtained by monitoring the current in the RESET +10 V line.

The gate-pulse generator circuit is shown in Fig. 5. The mixed signal is passed through a variable delay line having 8×8 ns sections, and then triggers an avalanche transistor[4] which in less than 3 ns produces the pulse of nearly 1 A required to drive forty gates. This current passes directly through the base-emitter junction of the transistor 2G106, thereby triggering this comparatively slow lock-in circuit.

When the lamp indications have been noted, the RESET +10 V line is briefly interrupted and the circuits again become sensitive. Transistor MM2711 disconnects the −10 V line to reset the lock-in circuit.

5. Performance

The apparent time resolution will vary with the nature of the input pulses applied in two ways. Firstly, a larger amplitude pulse will produce a faster output rise-time from the limiter, and it may go through whilst the inhibiting gate is closing, causing an apparent increase in gate width τ_D. A range of 3 : 1 in amplitude is allowed. This should cause a change in τ_D of about two-thirds the rise-time of the input pulses, which is approximately 10 ns.

Secondly, a smaller τ_D will be obtained when nearly all gaps fire correctly, compared with one firing spuriously. This is because the rise-time of the mixed signal is approximately 25 ns when only one signal is available, compared with 10 ns when many occur together. The mixed signal line has the capacitance of 39 diodes plus strays, perhaps 400 pF total, and a source impedance of the order of 50 ohms with one input pulse, giving approximately 20 ns time constant.

The values of τ_D which can be set by means of the switch are approximately 8, 16, 24, 32, 48 and 64 ns. These figures apply to normally triggered shots, when nearly all gaps fire correctly. In the case of spurious break-down, only one signal is available to trigger the gate-pulse generator, and these times are increased by approximately 15 ns. Variations in pulse amplitude affect the pulse rise-time, but a signal three times larger changes τ_D by, at worst, 6 ns. The variation from channel to channel is within ± 2 ns

In operation, the 40-channel monitor on *Tarantula* has been used for eighteen months, during which time there have been nearly 8000 shots of the machine. It enables the experimental staff to detect immediately any spark-gap which is failing, even intermittently. The equipment is regularly checked by inserting an extra length of cable in series with one or more inputs. As a cross-check, the waveform from each spark-gap has been examined on an oscilloscope and good agreement is found. This oscilloscope procedure takes two days to complete as many shots must be fired to obtain average values of time delay for each spark-gap, and would be the only method if the monitor were not available. To dismantle a faulty spark-gap takes a full working day, and using the monitor it is found that more than 90% indicated faults are genuine. Owing to difficulties in reaching 100 kV, the machine was first operated at 50 kV. The voltage is being increased gradually towards the design maximum, and the monitor is proving invaluable in tracing spurious break-downs.

6. Conclusion

The principle of 'A-before-B' has several advantages in situations where the wave shape of the signal is not well defined, since it uses the edges of the pulses and does not depend critically on their amplitude. This logic also enables one monitor to perform two functions. The circuits could probably be re-designed with the cheaper fast silicon transistors available now, but it may be difficult to obtain a low enough impedance in the gate transistor (MA393). As seen in the photographs of Figs. 6 and 7, a compact unit assembly is used.

7. Acknowledgments

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Telecommunications Research in the British Post Office

The work of the 1400 engineers, scientists and ancillary staff at the Post Office Engineering Research Station at Dollis Hill in North London is varied, but all of it is aimed at the extension, improvement and economy of the Post Office services, mainly in telecommunications, but also in postal engineering (such as the automatic letter sorting developments).

Every three years, the Station welcomes visitors from industry, universities, other government technical organizations and similar bodies. The exhibits, some of which are described below, illustrate the wide range of work from, for example, the study of the physical and chemical properties of substances which may later lead to new devices having profound influences on the means of communication, to the effects of human factors involved in the everyday use of the telephone itself.

Integrated P.C.M. Transmission and Switching

Pulse code modulation (p.c.m.) is a method of telephone transmission which was invented by A. H. Reeves, an Englishman, in 1937 but had to wait for the transistor to make it an attractive, practical proposition. In its application to the Post Office's telephone system, the speech waveform is sampled 8000 times a second, and each sample value is sent in telegraph code to the receiving equipment. The coded signal consists of 'on-off' pulses which can be recognized despite severe distortion and interference, regenerated to perfect shape at intervals along a route, and decoded by the receiver to yield the original wave substantially without error. The quality and voice level of calls are independent of distance.

Samples from several talkers can be interleaved in time so that two pairs of wires can carry, typically, 24 conversations. The fast but robust coded signals survive transmission over ordinary audio-frequency cables (intended for one conversation per pair), so p.c.m. provides for growth by a factor up to 12 times without new cables and the attendant road works. With modern semiconductor devices in the coders, decoders and regenerators, p.c.m. becomes attractive in cost at distances over 15 miles.

If, as is likely, p.c.m. is used extensively in the junction network, it will be attractive to keep signals in digital coded form where junctions are connected together in tandem, as for example in most cross-London calls, to save intermediate coding and decoding. This will halve the minimum economic junction length for the application of p.c.m. and conserve transmission quality. The tandem exchange should be smaller and cheaper in p.c.m. than in conventional form.

Integrated p.c.m. switching and transmission systems are being studied at the Research Station, and a model p.c.m. tandem exchange is being built to test the feasibility of incorporating them in the existing network.

Microwave Transmission

The Post Office microwave network is now growing at an extremely fast rate. Existing radio links use bands of frequencies around 2, 4 or 6 GHz, but these bands are being filled rapidly, so it will soon be necessary to expand into yet higher frequency bands. At these frequencies weather limitations—especially absorption of microwave energy by large raindrops—begin to become significant.

Continuous observations are being made on signals transmitted over an experimental radio path in an area of Bedfordshire for which detailed meteorological information is available. The experimental stations are unstaffed and the information obtained is sent automatically over a normal telephone line to Dollis Hill, where it is analysed by a computer to provide statistics which can be used for the design of future systems.

Present-day microwave links carry 960 or 1800 telephone channels per frequency allocation. In anticipation of demand for still further traffic capacity, experimental work has just started on the problems which will arise in transmitters and receivers carrying considerably higher number of simultaneous telephone conversations.

Psychometric Tests in Television

Economical design and maintenance of the links used to connect television studios to the transmitters depends upon setting appropriate tolerances for the many kinds of distortion and noise that may arise. The basic relations between the various impairments and their subjective effects are determined by a standardized method in which quality assessments are made by groups of lay observers.

Amounts of several types of impairment may be present simultaneously in practice, but in the laboratory it is impracticable to test all possible combinations. This difficulty has been overcome by finding a simple empirical law for predicting the total subjective effect of co-existing impairments, given the individual contributions. The method employs a summable subjective-impairment unit termed the IMP. In a colour-television demonstration, the effects corresponding to certain 'imp-values' are produced by applying various amounts of random noise to the luminance and chrominance channels.

Low-noise Parametric Amplifier for Earth Satellite Communication

In the global satellite-communication system the high-capacity earth stations will be required to receive several carriers located within the frequency band 3·7 to 4·2 GHz. It is therefore necessary to have a low-noise receiving system which will operate over this bandwidth.

The maser amplifier, which is in general use at earth stations, has a noise temperature of about 8°K but the bandwidth cannot readily be made to exceed 100–150 MHz. Low-noise parametric amplifiers are, therefore, being developed, the design target being a bandwidth of 500 MHz and a noise temperature of 15–20°K.

As a step in that direction, a helium-cooled parametric amplifier with a noise temperature of 20°K and a bandwidth of 50 MHz has been built, together with a 500 MHz bandwidth ambient-temperature amplifier. One problem which has to be overcome is avoiding 'shot' noise caused by diode current produced by the pump power. This, unlike thermal noise, cannot be reduced by cooling.

Submerged Repeaters

At the last Open Day a valve amplifier for a submerged repeater with a frequency range extending to 3 MHz was shown. Now its successor embodying transistors can be seen, but work is going on with a system which aims at a frequency range of 12 MHz.

Since a typical ocean cable route would require 600 repeaters, very small inaccuracies in the measurement of repeater gain and cable loss could result in catastrophic overall discrepancies. Therefore transmission measuring apparatus of extremely high precision is required. In certain cases absolute accuracy of the order to 0·01 dB and differential accuracy of 0·001 dB is necessary.

Prototype equipment is being designed to meet these requirements in the frequency range 0·01–15 MHz for losses and gains up to 60 dB; this range may be extended at a reduced accuracy to frequencies up to 100 MHz and to losses up to 100 dB.

To achieve the necessary accuracy, high grade components giving good return loss are employed, particular attention has been given to screening and the equipment is temperature controlled. Novel features of the design include the method of compensation for inherent attenuator errors and provision is made for the connection of auxiliary equipment for providing automatic print-out of the results.

Improved H.F. Radio Telephony

The Lincompex system in an earlier form was demonstrated in 1964.† Since then it has been given a successful field trial between the U.K. and India and has aroused considerable international interest. It has been further developed in collaboration with industry so that it is now suitable for wider trial on a number of routes.

The speech signal to be transmitted is compressed at syllabic rate, so that weak syllables go out at a higher level to overcome noise, and it is accompanied by a control signal indicating the degree of compression applied. This signal controls an expander at the receiving end to restore the speech to its original form. Considerable suppression of noise, particularly in pauses between words, and a constant transmission are achieved. The latter makes the conventional singing suppressors unnecessary and so permits an easier flow of conversation. Recordings and apparatus were demonstrated.

Transistors for use in Submarine Telephone Systems

A major objective of Post Office research and development on transistors is the provision of devices with the necessary performance and long-term reliability for use in the repeaters of submarine telephone systems. A system using a 12 MHz bandwidth requires transistors with a current gain cut-off frequency of at least 1000 MHz and, in the output stage, with a rated power dissipation of 1·5 W. Two devices, for input and output use respectively, have been developed at Dollis Hill and are now in pilot production.

† 'The Lincompex system for high frequency radio telephony', *The Radio and Electronic Engineer*, 29, No. 3, p. 172, March 1965.

Long-term reliability involves both the reliability of the physical structure and the stability of the electrical characteristics. The former has been much improved by the development of a new form of thermocompression bond. The latter has been aided by a study of the effect of processing on electrical stability using accelerated ageing conditions.

Cadmium Sulphide as a Semiconductor

Cadmium sulphide is a versatile semiconductor. It has long been used as a phosphor, and has recently been made to produce visible (green) laser light. It is strongly piezoelectric and can be made into transducers which resonate at microwave frequencies: the very small thicknesses required are produced by evaporation. The evaporation process can also be used to make thin-film integrated circuits containing CdS field-effect transistors offering high input impedance and high transconductance. Finally, the piezoelectric properties give rise to the electroacoustic effect, which has been used to make lossless delay lines and very simple amplifiers operating at 50–2000 MHz.

The exploitation of all these properties, which make it potentially a very useful substance in electronic devices, has been continually hampered by the difficulty of controlling the basic parameters of the material which stems from a lack of adequate scientific study. As a preliminary to any practical use of cadmium sulphide, therefore, a general study of its preparation and properties is now in hand at Dollis Hill. This includes the preparation of CdS powder in a high state of purity, growth into predominantly single-crystal 'boules', which are cut into oriented slices for measurement, studies of stoichiometry, evaporation, and annealing, and measurements of carrier mobility and acoustoelectric interaction. Preliminary work on the preparation and properties of thin films of cadmium sulphide produced by evaporation is beginning.

Electrical Conduction along Dislocations in Insulating Crystals

The electronic band structure of a crystal is expected to be perturbed in the neighbourhood of a dislocation. For a simple edge dislocation in a crystal of simple structure the perturbation will result either in a decrease of the forbidden energy band gap on the compressed side and an increase on the dilated side or vice versa. If the decrease is sufficiently great compared with kT out to a large enough distance from the dislocation core, electrons or holes (according to whether the conduction or valence band is more perturbed) will be constrained to move along the line, provided it is sufficiently free from ionized impurities or charged 'jogs'. Theoretical aspects of this type of conduction are being studied at the University of Strathclyde, and experimental studies of conduction along dislocations in KCl are in hand at the City University, both studies being under Post Office sponsorship with Ministry of Aviation support. Experimental studies of the nucleation, expansion, and electrical conductance, of edge dislocation segments are in progress at Dollis Hill. The work is expected to lead to a better understanding of conduction in insulating crystals, and may lead to the development of a new type of adaptive three-dimensional solid electronic circuit. Such a device, in the long term, could be applicable to computer-type apparatus which 'learns' by experience, e.g. automatic character readers.