THE

Telecommunication Journal OF AUSTRALIA

IN THIS ISSUE

VOL. 12, No. 3. Registered at the General Post Office, Melbourne, for transmission by post as a periodical. FEBRUARY, 1960

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In the last issue of this Journal we promised important news for subscribers. Here it is:

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Last year, many subscribers were unable to obtain Vol. 12, No. 1 (June, 1960) because of late forwarding of subscription renewals. Prompt return of the renewal slip will avoid a similar disappointment this year, particularly as many interesting articles will appear, for example, an excellent series giving a simple explanation of transistor circuits.

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The

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of Australia

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This Journal is available to members of the Telecommunication Society of Australia without charge. Non-members may obtain copies of all publications of the Society on the payment of 10/- p.a. Single copies of this Journal are available at 4/- each.

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MR. H. T. WRIGHT, A.M.I.E, Aust.

More than 200 officers and friends attended a function arranged by a Committee of the Victorian Engineering Division and held at Russell Exchange building on Tuesday, 15th March, 1960, for the purpose of bidding farewell to Mr. H. T. Wright on his retirement as Superintending Engineer, Metropolitan Branch, Victoria.

Mr. Wright has had an interesting and varied career in the Department. After spending his early life at Geelong, he was first appointed to the P.M.G.'s Department in 1917 as Draftsman in Hobart, and, in 1924 became Draftsman-in-Charge, Hobart. After a period as Engineer in the Lines Section in Hobart he was transferred in 1929 to Bendigo, Victoria, where he remained until 1935.

After transfer to Melbourne in 1935 Mr. Wright concentrated mainly on telephone equipment work serving successively in all Metropolitan Equipment Divisions. Appointment as Divisional Engineer in 1941 was followed in 1945 by promotion as Assistant Supervising Engineer, Equipment Service, Melbourne, and in 1949 he was appointed as Assistant Superintending Engineer, Internal Plant, Victoria. In this position he assumed the responsibility of all internal plant within the State, including Long Line and Telegraph equipment.

Following a Departmental reorganisation Mr. Wright in 1955 was promoted as Superintending Engineer, Metropolitan Branch, in which capacity he had the responsibility for all the external exchange and substation plant within the metropolitan area. In 1954 he was appointed by the Director-General to a select Committee formed to study and introduce the new system of exchange maintenance, which is known as Qualitative Maintenance throughout the Commonwealth. In this capacity he visited the major centres in the Commonwealth in order that the new system might be introduced smoothly taking into account any special local conditions. Also as the leading Victorian representative on the "Necessity to Balance Deliveries" Committee, which comprises representatives for the Headquarters and the four larger states, he played no small part in assisting in the economical distribution of available telephone equipment of good technical quality throughout the Commonwealth.

Apart from his technical and administrative achievements Mr. Wright has always taken a keen interest in staff welfare by keeping in close touch with officers and employees under his control and influence. His courteous, friendly manner and innate sense of fair play has earned for him a wide circle of friends which has been a material contribution to the excellent team spirit existing in the Engineering Division. He has always laid great stress on the desirability of safe working methods, and for the past two years was Chairman of the Safety Committee aimed at improving safety particularly in the Metropolitan and Country Branches. He was for most of his career a very active member of the Professional Officers' Association and in recognition of his efforts in this field the Central Council of that body in September 1954 appointed him a life member of the P.O.A.

Mr. Wright has always taken a keen interest in the affairs of the Telecommunication Society and contributed valuable articles to the Journal during his career. Our best wishes go to him in his retirement, which as a characteristic of his energetic outlook is to start in a new venture of developing a primary producing enterprise.

THE NATIONAL TELEPHONE PLAN—CALL CHARGING

*R. W. TURNBULL, A.S.T.C. (Elec. Eng.), G. E. HAMS, B.Sc., and W. J. B. POLLOCK, B.Com.**

INTRODUCTION

Since the June, 1959, issue of this Journal, which contained a description of the national numbering plan, further steps have been taken toward the implementation of the National Telephone Plan. A new tariff scheme designed to facilitate the automatic charging of all trunk line calls ultimately was announced by the Postmaster-General during the Budget session of Parliament. The new tariff scheme is being introduced in two steps, on 1st October, 1959, and 1st May, 1960. This article describes the principles of the call charging system. A subsequent article will describe the switching network plan which is the third element of the national telephone plan.

BASIS OF EXISTING TARIFF STRUCTURE

The measured service rate system is the basis of the telephone tariff structure in Australia. Each subscriber is charged a rental for his telephone service, which varies according to certain prescribed conditions, and is required to pay the charges for local and trunk calls made from the service. The annual base rental for a telephone service depends on the class of service whether business or residential and on the total number of subscribers accessible by a local call.

Local Call Charging: In the metropolitan areas of Sydney, Melbourne, Brisbane, Adelaide, Perth, Hobart, and Newcastle, local call networks are established and calls between all exchanges within a prescribed radius of the General Post Office are charged at the local call fee. In Newcastle the network centre is at Hamilton. The network radius is fifteen miles in Sydney and Melbourne and ten miles in the other metropolitan areas and Newcastle. In all other areas throughout the country, the local call range embraces all exchanges within a radius of five miles of the calling exchange. Throughout Australia, local calls are untimed and are charged at a uniform rate of one unit fee.

The rates for trunk line calls are related to the radial distance between calling and called exchanges. On manually operated trunk line calls, there is a fixed minimum chargeable period of three minutes. This is designed to cover costs of manual operation in having a telephonist set up a long distance call and record details, including the duration and charge for the call, followed by the clerical work of sorting the docket and finally, billing. For the subscriber, trunk charges based on three minute periods have the disadvantage that for calls of less than three minutes, or less than an exact multiple of three minutes, he pays for a longer period of time than the duration of his conversation. Moreover, lines and equipment are often engaged without absolute necessity because many subscribers are inclined to carry on for a full three minute unit, once such a unit has been commenced.

The recognition of individual exchanges for trunk charging and the selection of a rate from a large number of steps in the tariff scale do not impose any great difficulty under a manual system. They have very important implications, however, for an automatic charging system.

AUTOMATIC CALL CHARGING METHODS.

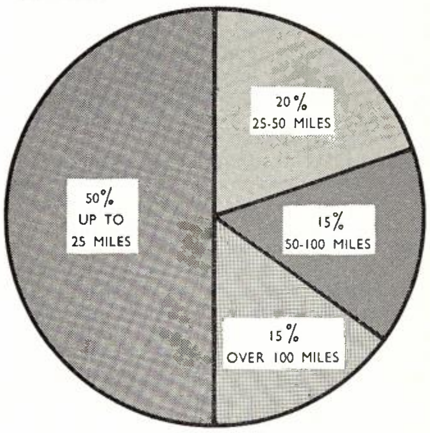
In most telephone systems where trunk calls are connected manually by telephonists, the operator handling the call records on a docket the particulars of the calling and called numbers and the duration of the call. From these, the charge is calculated and entered. With a subscriber trunk dialling service, it is necessary to provide special automatic equipment to carry out this charging function.

Toll Ticketing: One method of automatic charging is known as toll ticketing. Where this is used, the calling and called subscribers' numbers and the date, time and duration of the call are recorded either on a ticket or on some other medium such as perforated tape or magnetic tape. For reasons of economy, it is impossible to provide each subscriber's circuit with an individual apparatus to record all these details and the recording apparatus is provided as common equipment to which subscribers have access. This imposes the necessity for equipment to identify the individual calling subscriber's number. The called subscriber's number is ascertained from the digits dialled. The toll ticketing system provides for detailed billing of calls but is expensive.

Multi-Metering: The method of automatically charging trunk calls, which has been adopted by most European countries, is that of multi-metering. With this system, the charges for trunk calls are recorded on the subscriber's meter or its equivalent, as is the case for local calls. The meter registers at intervals according to the charge rate of the call. Separate billing of trunk calls is not practicable in this system and only one total charge for all automatically established calls, both local and trunk, is shown on the subscriber's account. This is referred to as bulk-billing.

Application in Australia: The plan for the Australian subscriber trunk dialling system is based on the multi-metering method of call charging. In Australia, where the measured rate system applies, individual meters are provided on all automatic services to record local call charges. These will also be utilised for recording trunk call charges. If a flat rate system not involving the recording of local call charges had been in operation, the factors to be considered would have been different. A toll ticketing system would make possible more detailed billing of calls, but only at increased cost.

A feature of trunk traffic in Australia is that although great distances are involved on some calls, 85% of the trunk calls are made over distances of less than 100 miles. The distribution of calls over the various mileage ranges is shown in Fig. 1. Experience in Australia on the short distance trunk routes which have been equipped with automatic multi-metering or where manual methods of applying charges, using techniques similar to multi-metering, have been employed indicates that this charging method is acceptable to subscribers.



PERCENTAGE TRUNK LINE CALLS IN RELATION TO MILEAGE
FIG. 1.

On multi-metered calls, a separate record of each call is not available, but this does not vary from the practice of other utilities which adopt bulk-billing for the supply of water, gas and electricity.

Operator services will always be retained for subscribers wishing to have their trunk calls individually recorded and, in those instances such as guest houses or hotels where subscribers find it essential to know the charge for each call, it will be possible to provide, at an appropriate rental, meters at the subscriber's premises, from which the cost of each call can be calculated.

There will be a requirement, in particular instances, for access to the trunk line network to be barred from a particular subscriber's telephone. Instances of this could be from some extensions on private automatic exchanges or even on individual subscriber's services. Where practicable, this facility will be made available to subscribers upon payment of a suitable fee.

*The authors of this article are the members of the Headquarters' A.N.S.O. Committee. Mr. R. W. Turnbull, the Chairman of the Committee, is a Superintending Engineer attached to Headquarters; Mr. G. E. Hams, the other Engineering representative, is an acting Sectional Engineer at Headquarters and Mr. W. J. B. Pollock, who represents the Telecommunications Division, is an Assistant Controller at Headquarters.

The adoption of the multi-metering method as the basis for call charging in the subscriber trunk dialling system imposes special requirements for the design of the call rate charging schedule. The main requirement is that the tariff structure should be simplified as much as possible to achieve the maximum benefits from the new system.

FEATURES OF MULTI-METERING SYSTEM.

With a multi-metering system, the call charging equipment must, at the time of the call, ensure that pulses are applied to operate the calling subscriber's meter according to the duration of the call, after having determined the appropriate charging rate.

Adoption of Periodic Metering for Trunk Calls: There are several ways in which pulses could be arranged to operate the subscriber's meter. For instance, a number of pulses could be applied at the start of a call to preregister the charge for a 3-minute period. In this case, it would be necessary to time each call individually, so that the time of application of each group of pulses would be related accurately to the time of commencement of the call. With large numbers of calls of random time incidence to be handled by equipment individual to each connection, the total investment in exchange equipment for the charging function alone would be considerable.

The method adopted for Australia, which is considered to be the most equitable, is periodic metering, in which the meter registers one unit at a time at regular intervals during the progress of a call. The intervals between meter registrations vary with the charging rates for the distances over which the calls are made. Periodic metering has the significant advantage that callers can relate the cost of a call more closely to the period of conversation. Charges for automatic calls are not then arranged in basic 3-minute periods and calls of short duration, even over a long distance, may cost only a very small fee.

This system is relatively simple to apply in practice because less precision is required in the timing of pulses in relation to the start of each call. A common source providing each of the required pulse rates can be used to serve a whole exchange or even a number of exchanges. The greatest difference in charges for calls of the same duration and to the same destination which could result would be one unit fee. All calls being charged at a particular rate would be connected to one common source of pulses. In this way, the amount of individual charging equipment to be associated with a particular call is kept to a minimum.

In the periodic metering system planned for Australia the calling subscriber's meter operates once when the called subscriber answers, the first of the regular timed pulses arriving at random is suppressed and there is a fixed interval between subsequent pulses. The frequency of the regular pulses is determined by the particular tariff rate applying to the call.

This form of the periodic metering system provides for one initial registration on all chargeable calls no matter how short the duration. However, the interval between the initial registration and the first regular pulse arriving at random can be less than the fixed interval appropriate to the call and the first regular pulse is suppressed to avoid the possibility of the subscriber being overcharged.

Number of Charging Rates: Costs of multi-metering equipment increase with the number of charging rates in the tariff scale because additional equipment is necessary to make the selection of the rate to be applied from a larger number of alternatives and the pulse generation equipment is more costly. The optimum requirement here is to have sufficient pulse rates generated to yield adequate flexibility for tariff changes, yet to keep the number as low as possible. The number of pulse rates which meets the requirements of the Australian system is 16, with pulsing rates as follows:—3, 4, 5, 6, 9, 10, 12, 15, 18, 20, 30, 36, 45, 60, 90 and 180 second intervals. It is necessary to ensure that the addition of a limited number of pulse rates can be arranged readily if required at a later stage.

From these rates, an 8 step tariff scale represents the most suitable balance at this stage between equitable charging and technical consideration. The 8 timed rates, with the addition of the untimed unit fee for local calls and free calls to certain Post Office services, such as for "Information", make a total of 10 individual charging categories from which the rate appropriate to any call may be selected by the equipment. Limitation to a maximum of 10 categories makes most effective provision for the use of automatic switching equipment employing 10 point rate selecting mechanisms or a decimal type of signalling in and between exchanges, as used in the present automatic system, for centralised control of automatic charging. Under this system, additional costs would be incurred and complexity introduced if more steps were required in the charging scale.

Determination of Charge Rate: The charging equipment determines the rate of charge to be applied to a call by examining the digits in the number dialled by the calling subscriber and selecting the appropriate rate from a range of categories in the trunk line tariff scale. It follows, then, that the cost and complexity of the equipment is closely related to the extent of the digital examination necessary and, as discussed, the number of different charging rates in the tariff scale. Because equipment design is influenced to a great extent by the digital examination necessary, there must be close co-ordination between the numbering plan and the call charging system.

Exchange to Exchange Charging: If exchange to exchange charging were employed as in the existing manual system, the charging equipment would have to examine sufficient digits to identify each of the exchanges in the Commonwealth, of which there are approximately 7,300. On calls within a 5 digit numbering plan area, as described in the earlier article dealing with the numbering plan, this would require examination of one (D), two (DE) or three (DEF) digits for exchanges of 10,000, 1,000, and 100 lines respectively; in 6-figure numbering plan areas discrimation on an additional digit would be needed in each case.

For national calls other than those to capital city networks, four (ABCD) five (ABCDE) or six (ABCDEF) digit examination would be necessary for the three sizes of exchange. Exceptions would occur whenever all the exchanges with numbers commencing with the same A, B or C digit were situated in the same mileage category from a particular originating exchange.

Exchange to exchange charging would also require that charging equipment should be individual to every exchange with subscriber trunk dialling facilities. Although to facilitate maintenance and secure some plant economies, it would be possible to centralise charging equipment at switching centres by the use of reverse pulsing, individual charge translations would still be required for each exchange.

Grouping Exchanges for Charging: For economic reasons then, a national subscriber trunk dialling system with multi-metering requires a departure from the principle of exchange to exchange charging and the adoption of a system, somewhat similar to that followed in Australian capital cities for many years, of grouping exchanges so that all exchanges in the group have a common charging basis. The charge for calls between any two exchanges is related to the distance between groups of exchanges to which they belong. Charging equipment need, therefore, only discriminate on sufficient digits to identify the called group of exchanges and not the individual exchange. Also, all the exchanges in a group can share common charging equipment which can be centralised at a convenient switching centre. Pulses can be sent back to operate meters at remote exchanges or an indication of the charge rate can be sent back and used at the remote exchange to select the appropriate rate of meter pulses to be applied to the call.

Although grouping a number of exchanges for charging purposes effects economies in equipment, it follows that different rates may apply to calls between exchanges which are the same distance apart. Such anomalies based on distance considerations alone are most noticeable on short distance calls, particularly in the case of exchanges with adjacent boundaries but which are located in different charging groups. They are not greatly significant on long distance calls. However, the distance between exchanges does not always reflect the cost of connecting the call. Even where the distance between exchanges is the basis for charging, there is still the possibility of a different rate applying between two subscribers' services which are the same distance apart, due to the irregular shapes of exchange boundaries.

If the grouping of exchanges could be so arranged that they covered areas of uniform shape and size, this would best satisfy charging requirements but the layout of trunk channels and switching equipment does not follow any regular geometric pattern. From the viewpoint of plant provision, to accept charging groups conforming to the layout of the trunk switching system would effect the greatest economies. The areas which the groups would cover following the adoption of this basis, however, would be very irregular in shape and varied in size. They would result in so many distance anomalies that they would be unsatisfactory for charging purposes. Some compromise between the two conditions is essential.

GROUP CHARGING PLAN FOR AUSTRALIA .

A group charging plan has been designed to meet the requirements of the Australian telephone system, having regard to the characteristics of the distribution of trunk traffic as shown in Fig. 1. It will be introduced on 1st May, 1960. The plan recognises the modern trends for community of interest to extend over a wide range as shown by the large percentage of trunk line calls which are made over short distances up to 25 miles. Under the plan, many of these calls will become local calls. Again, as 85% of trunk calls cover distances of less than 100 miles, fairly accurate assessment of the distances involved on these calls is required and the zone charging basis ensures that this is achieved. About 15% of calls cover distances in excess of 100 miles and extremely accurate calculation of distances on such long calls is not required. The district charging plan applying to the longer calls recognises this characteristic and utilises it to achieve significant equipment simplifications.

Principles of Group Charging.

The principles of the group charging system for the Australian network are illustrated in Fig. 2. The basis is as follows:—

Local Charging: Exchanges are grouped to form zones. Calls within a zone and to adjacent zones will be treated as local calls.

Trunk Charging — on Zone basis: Zones are grouped to form districts. Calls (other than local calls) within a district or to adjacent districts will be charged at trunk rates based on the distance between zones.

Trunk Charging — on District basis: Calls between districts which are not adjacent will be charged at trunk rates based on the distance between districts.

Determination of Charging Groups.

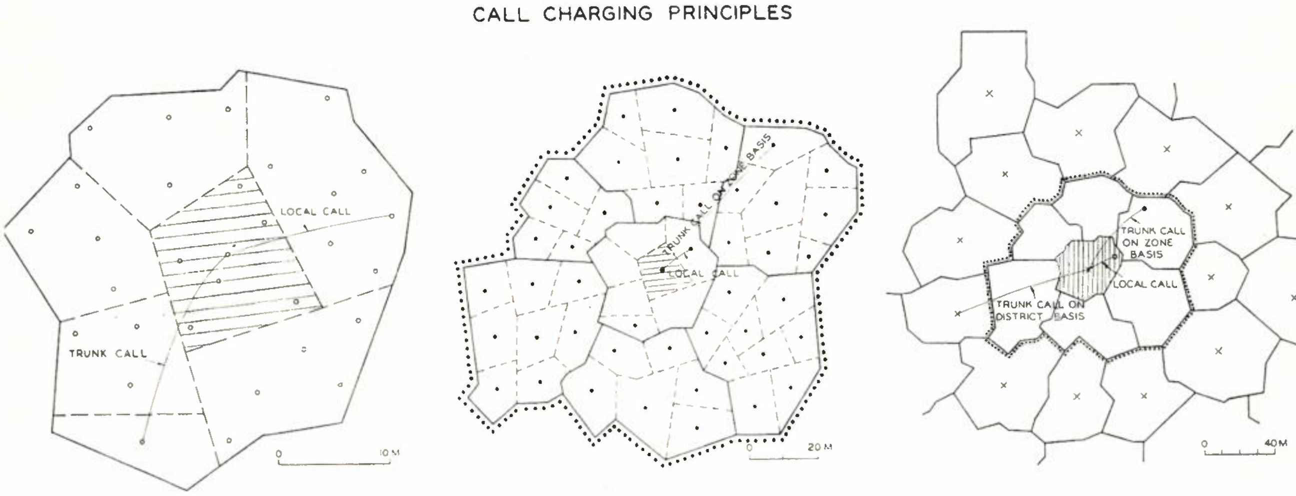
In the plan the emphasis is on groups of exchanges rather than on geographical areas in referring to the charging groups. Groups of exchanges serve areas in which the subscribers' services connected to the various exchanges are located, but the principle of group charging is such that no fixed boundary lines are intended to define the extent of charging groups. Any new exchanges established will be included in the most appropriate group for charging purposes depending on the economic and community of interest factors influencing the network development.

The trunk switching plan to be described in a later article has been evolved having regard to community of interest and the most economic disposition of plant. It therefore provides the most suitable basis for the initial approach to the determination of charging groups. The general principles already described relating to the grouping of exchanges for charging purposes have been applied in determining the zones and districts to be followed in the charging plan.

In practice, the minor switching areas in each secondary area were taken as the starting point. Where possible, the whole of a minor trunk area was included in the one zone. This was not possible in many cases and large minor areas were divided and some small ones combined to form acceptable zones. Under the trunk switching plan, nearly all traffic to and from a minor area routes through the minor trunk centre. This is, therefore, the most convenient location for centralised charging equipment which, at these centres, will be required to determine rates for calls — either untimed unit fee, free calls or trunk calls — at least those based on zone charging.

In the formation of charging districts, secondary trunk switching networks were taken as the starting point. However, in some cases because of the very large area covered by a secondary network or its irregular shape, it was found desirable to divide a secondary area for district charging purposes. In these cases, a minor switching area generally forms a charging district. Again, because of the small areas covered by some secondary areas, it was found desirable to

CALL CHARGING PRINCIPLES



LOCAL CHARGING

EXCHANGES ARE GROUPED TO FORM ZONES
CALLS WITHIN A ZONE AND TO ADJACENT ZONES WILL BE TREATED AS LOCAL CALLS.

THE DIAGRAM SHOWS THE LOCAL CALL RANGE FOR SUBSCRIBERS IN THE SHADED ZONE.

TRUNK CHARGING ON ZONE BASIS.

ZONES ARE GROUPED TO FORM DISTRICTS.
CALLS (OTHER THAN LOCAL CALLS) WITHIN A DISTRICT OR TO ADJACENT DISTRICTS WILL BE CHARGED AT TRUNK RATES BASED ON THE MILEAGE BETWEEN ZONE CENTRES.

THE DIAGRAM SHOWS THE RANGE OF TRUNK CALLS FROM THE SHADED ZONE WHICH WILL BE CHARGED ON THIS BASIS. THE UNIT FEE RANGE DESCRIBED IN FIG 1 IS ALSO SHOWN

TRUNK CHARGING ON DISTRICT BASIS.

CALLS BETWEEN DISTRICTS WHICH ARE NOT ADJACENT WILL BE CHARGED AT TRUNK RATES BASED ON THE MILEAGE BETWEEN DISTRICT CENTRES.

THE DIAGRAM SHOWS THE BOUNDARY ⁚⁚⁚⁚⁚ WHERE DISTRICT TO DISTRICT CHARGING COMMENCES FOR CALLS FROM THE SHADED DISTRICT.
WITHIN THIS LIMIT, ZONE AND LOCAL CHARGING AS SHOWN IN FIGS 1 & 2 WILL APPLY.

FIG. 2.

combine two such areas to form the one charging district. The charging equipment at switching centres of higher order than minor trunk centres, and even at some minor trunk centres as indicated, will be required to determine rates on calls charged on the district basis, in addition to those charged as untimed unit fee, free calls or on a zone basis.

Extended Local Service Area.

The importance of the extent of an individual zone is made much less critical by charging calls between adjacent zones at the same fee as for calls within a zone. By adopting the local call charge of one unit fee for an untimed call, a great deal of existing automatic equipment can be more effectively utilised without the need for special charging equipment.

If a higher fee were applied on calls to adjacent zones, a short call across the limits of a zone would cost more than a longer call within the zone. Short distance trunk calls form the great bulk of trunk traffic and distance anomalies on these could appear unreasonable to subscribers.

By applying a uniform local call charge on calls within a zone and to adjacent zones, a "buffer ring" of local call access is placed around each zone. Not only does this overcome the anomaly mentioned, but the actual shape and size of the zone in which his exchange is located loses some importance from the subscriber's viewpoint. Of greater importance is the extent of the limits of zones adjacent to the subscriber's own zone which comprise the extent of local call access.

The extent of local service areas is determined having regard to overall costs and revenue requirements. Under the plan, many calls up to 25 miles and some even up to 35 miles will become local calls.

Numbering Implications of the Charging Scheme.

The charging plan has been carefully co-ordinated with the numbering plan and the extent of the digital examination to determine charge rates is as follows:—

(a) For calls within a numbering plan area — not further than the E digit.
(b) For calls beyond the numbering plan area but within the charging district or to adjacent charging districts, where charging is on a zone basis — not more than the ABCDE digits.
(c) For calls beyond adjacent charging districts, where charging will be on a district basis — not more than the ABC digits.

SUMMARY

The new Australian call charging system to be introduced on 1st May, 1960, provides for:—

Extended Local Service Areas (ELSA).
Group Charging.
Fewer Trunk Rates.
Multi-metering on Subscriber Dialled Trunk Calls.

ELECTRONIC FAULT LOCATOR—TYPE F.L.O.S.

*W. E. ROSS**

INTRODUCTION

The Alan H. Reid Electronic Fault Locator Type F.L.O.S. shown in Fig. 1 is now used extensively by the Department. The technique involved makes use of the fact that constructional irregularities or major fault conditions on a line give rise to impedance irregularities which should not be experienced on a normal and properly constructed line. Impedance irregularities cause a portion of the energy being propagated along the line to be reflected back to the source. A major fault condition existing on the line would give rise to a large reflection. The time taken for an electro-magnetic wave pulse to reach an impedance irregularity on a line and to be reflected back to the source is known as the echo time, and the measurement of this echo time is the basis of operation of the locator. The instrument gives a direct indication of the echo time on the screen of a cathode ray tube.

GENERAL PRINCIPLES

The echo time can be converted to distance to a fault if the velocity of propagation of electro-magnetic waves along the telephone line is known. The velocity is generally in the range 20,000 to 186,000 miles per second. As an example, if the speed of propagation is 180,000 miles per second, which applies to open wire lines, the time taken for a wave to reach the fault would be $10^6/180{,}000$ or 5.5 microseconds for each mile of line between the wave source (or testing station) and the fault. As waves have to be propagated to the fault or irregularity and reflected back to the source, the total echo time measured for each mile would be $2 \times 5.55 = 11.1$ microseconds.

The fault locator uses a undirectional pulse. Insofar as the length of the pulse is concerned, it must be sufficiently small to allow irregularities close together on a line to be distinguished from each other. If possible the pulse should also be of a suitable duration to allow it to be completed before an echo is received from an irregularity very close to the testing point. To determine the pulse qualities required, the attenuation offered and speed of propagation have to be taken into account. For example, if the echo time to cover one mile with a speed of propagation of 180,000 miles/second is 11.1 seconds, then a one microsecond pulse would occupy a distance of .09 miles, and a fault at this distance from the testing point could be detected. The attenuation of the line at frequencies involved in the pulse spectrum affects the optimum pulse length due to the pulse being attenuated whilst traversing the line and while being reflected, the degree of attenuation depending on the frequencies involved in the spectrum. The frequency spectrum of one microsecond D.C. pulse contains frequencies of significant amplitudes up to several megacycles/second. Variation of the pulse length alters the frequency spectrum and attenuation of the pulse on the circuit under test. A particular pulse length is usually chosen in order that irregularities can be ascertained on the maximum range of an instrument without undue attenuation of the pulse, thus minimising screening of various irregularities. An additional pulse must not be transmitted from an instrument before the previous pulse has reached the furthest point on the maximum range of the instrument and has returned to the instrument. Taking the speed of propagation as 180,000 miles per second, the echo time on a 150 mile range would be $2 \times 10^6 \times 150/180{,}000$ equals 1,666.6 microseconds; therefore the maximum number of pulses which could be transmitted per second is $10^6/1{,}666$ or approximately 600. This means that the repetition frequency of the pulses to line would have to be less than 600 per second to allow for the maximum range of 150 miles. The echoes or reflections returned to the instrument are amplified and displayed on the screen of the cathode ray tube, thus giving an indication of the irregularities existing on the line under test. Calibration of the distance on the cathode ray screen is obtained by marker pips spaced equally in a vertical position across the horizontal trace displayed on the screen.

The marker pips are located at 55.5 microsecond intervals, corresponding to a distance of 5 miles on open wire lines. The locator transmits a unidirectional negative pulse to line of about 300 volts peak with a duration of about one microsecond. Pulses are transmitted to line at the commencement of each cycle of the 50 cycle mains supply, thus giving a pulse repetition frequency of 50 pulses per second.

CIRCUIT

A circuit of the Reid type F.L.O.S. locator is shown in Fig. 2.

Power Supply: The instrument is mains operated. A regulated H.T. voltage of 300 volts D.C. is supplied to all tubes via voltage regulators V13 and V14, type OD3/VR150. A high A.C. voltage is rectified by V16, and the resulting D.C. voltage, which is about 1500 volts, is used for the cathode ray tube.

Square-Wave Generator: Valves V1, V2 and V3 are used to produce a square shaped pulse of about 100 volts amplitude and about 1200 microseconds duration which is repeated every 200 microseconds. The amplitude is adjusted by

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*Mr. Ross is Supervising Technician at Jerilderie, New South Wales.

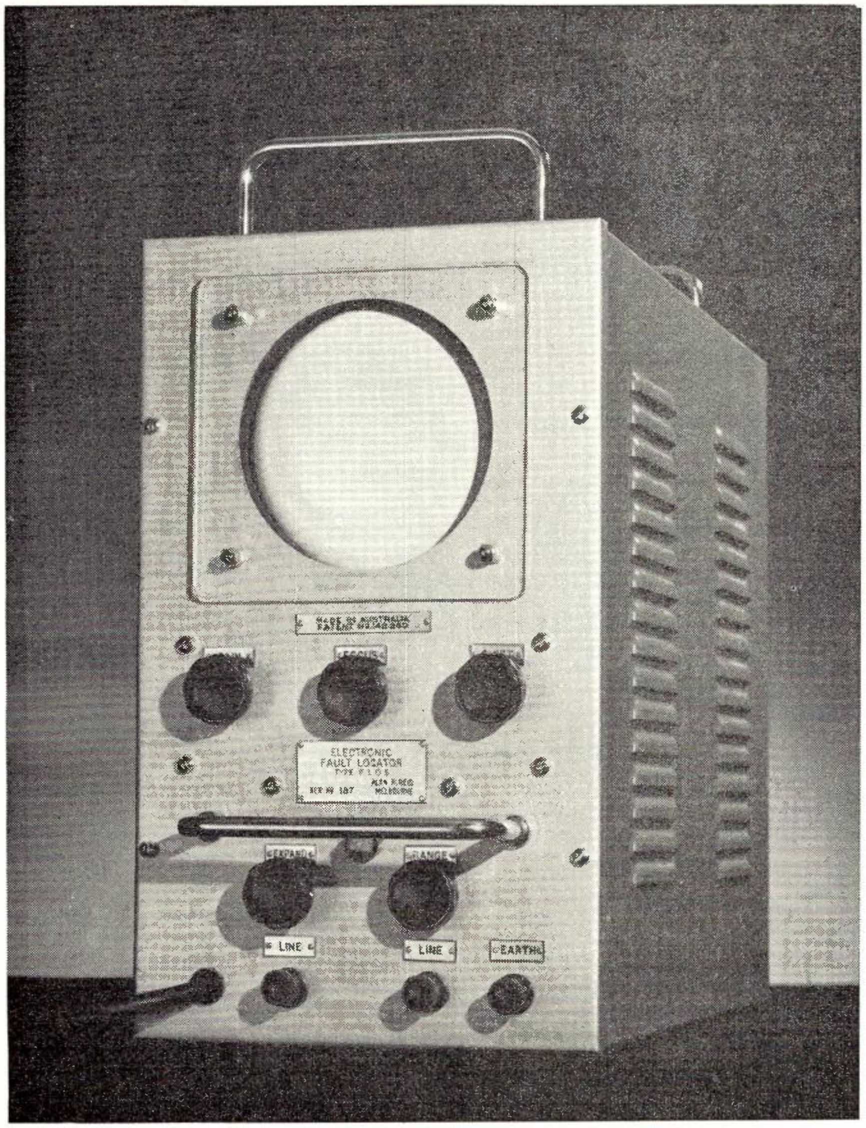


Fig. 1 — Photograph of Fault Locator.

means of R10 but is not critical. The square wave is started by applying A.C. to the grid of V1, and the output of V1 consists of a 50 c/s square wave produced by limiting the top and bottom peaks of the input sine-wave. The 50 c/s square pulse is shortened by the R2 — C1 combination to the required duration of 1200 microseconds, and amplified by V2 and V3. The duration is set in the factory by adjustment of the value of C1. The duration determines the maximum range of the instrument since the output pulse controls the time base circuit described later, and the brightening circuit of the cathode ray tube. Thus using a pulse duration of approximately 1200 microseconds and assuming a rate of propagation of 180,000 miles per second, which gives echo time of 11.1 microseconds per mile, the range would be in the vicinity of 1200/11.1 or approximately 108 miles. Positive and negative pulses are produced at the output of V3 across R8 and R7 which are of equal resistance; the wave forms are shown in Fig. 3. The pulses are used to perform four functions. Two are performed by the negative, and two by the positive half of the wave form. It is very essential for accurate location of faults that these functions are all started at precisely the same time.

Time Base: The time base is fed to the horizontal plates of the cathode ray tube and is obtained from valves V4, V5 and V6. V4 is the time base starter, V5 the time base amplifier and V6 the phase inverter. The output of V5 is fed to one horizontal plate and the output of V6 is fed to the other horizontal plate. It can be seen that V4 normally offers a low impedance to the time base components R12 and C4. When the negative wave is in evidence on the plate of V3 and applied to the grid of V4, the impedance of V4 reaches a high value and condenser C4 is charged via resistance R12. The resulting potential rise is applied to the grid of V5 where it is amplified and fed to one horizontal plate, the phase being inverted by V6 and fed to the second horizontal plate. The saw tooth wave obtained from V5 and V6 is shown in Fig. 4. The V5 stage contains the range and expand controls, the functions of which are explained later.

Marker: The calibration pips (or marker pips or distance markers) are provided as vertical indications on the horizontal trace. The pips are controlled by an 18kc/s oscillator and are displayed at intervals of 55.5 microseconds across the horizontal trace. Valves controlling the operation are V8, V9, V10 and V12. Valve V8 performs a function similar to valve V4 in the time base circuit in that it acts as a starter for the oscillator when the 1200 microsecond negative pulse is received from the plate of V3. Valve V8 becomes high impedance and allows the Hartley oscillator valve V9, with its tapped inductance, to oscillate. These oscillators are squared by V10, and peaked by C20 and R46. Diode V12 suppresses each alternate pip, these pips being of the wrong polarity and not required. The pips are applied to one vertical plate of the cathode ray tube. The 18 kc/s marker pip oscillator is in operation for 1200 microseconds. One cycle occupies a period of 10^6/18,000 or 55.5 microseconds, therefore, the rectified pips appear every 55.5 microseconds corresponding to a length of five miles of open wire line, and as the oscillator is in operation for 1200 microseconds the pips are in evidence on the complete horizontal time base.

Line Pulse: Valve V7 generates the pulse sent to line. It is a thyraton valve which operates when a positive wave is applied from the cathode of V3. This has the effect of discharging condenser C11 to the external load connected via the line transformer. The size of C11 has been chosen to give a unidirectional pulse of width at the half amplitude points of the order of one microsecond duration when working into a 600 ohm load. The negative bias for the thyraton is obtained via R9.

Echo Amplifier: Echoes returning from the line under observation are applied via the 600-600 ohm transformer and the

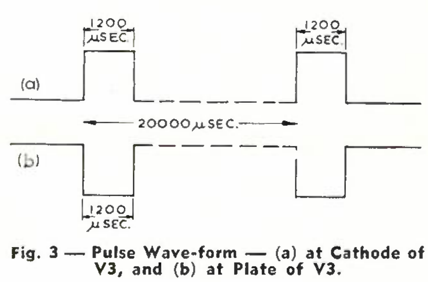


Fig. 3 — Pulse Wave-form — (a) at Cathode of V3, and (b) at Plate of V3.

adjustable 5Kohm gain potentiometer to valve V11, where it is amplified and applied to the second vertical plate of the cathode ray tube.

Cathode Ray Tube: Since the pulse repetition frequency is 50 pulse/sec., the time between two pulses is 20,000 microseconds. The active time of each period

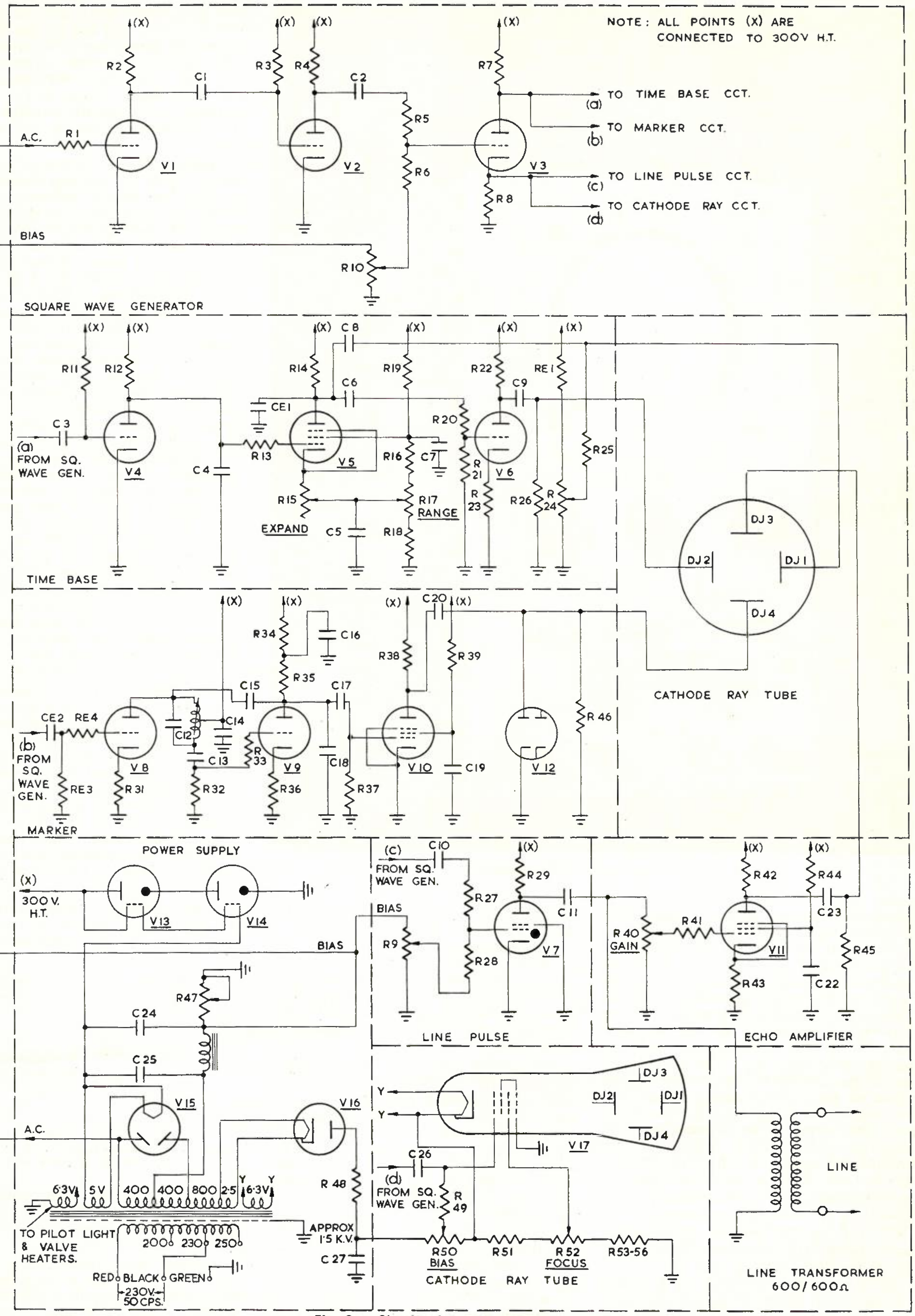


Fig. 2 — Circuit of Fault Locator.

is 1200 microseconds, leaving the instrument inoperative for 18,800 microseconds. The trace on the cathode ray screen is illuminated by a positive brightening pulse via C26 during the active portion of the cycle only.

OPERATION

Practical application of the locator has resulted in accurate location of faults on open wire lines. It is particularly valuable for the location of intermittent and H.R. faults. However, owing to the nature of the pulse, satisfactory tests cannot be obtained through line filters or loaded cables, but modifications to overcome these conditions have been made to a number of instru-

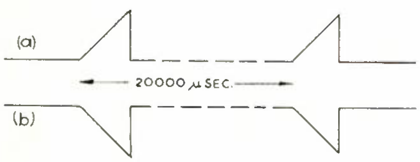


Fig. 4 — Saw-tooth Wave — (a) at Plate of V6, and (b) at Plate of V5.

ments by the Postmaster-General's Department.

The recommended operating procedure is as follows:

The control knobs should be adjusted to the following positions:

(a) Gain knob turned fully anti-clockwise (minimum);
(b) Expand knob turned fully anti-clockwise (minimum);
(c) Range knob fully clockwise (maximum).

Focus and shift knobs are usually in the desired position from previous operation to give the required intensity and position of the trace on the screen. After connecting the commercial supply and the trace has appeared on the screen, the shift knob is adjusted until the trace commences approximately one inch from the left-hand end of the screen. Next, the focus control knob is adjusted until the trace appearing is of a sharp nature to give the best trace of the line being observed. The lengths of physical lines terminating at control stations are always known, and the trace appearing on the screen is adjusted by means of the range adjustment R17 to cover the particular distance involved. With the line connected, the gain control is turned clockwise, increasing the amplifier gain. The fault condition existing will be indicated by a V-shaped peak on the horizontal trace, being introduced at the fault point on the trace by the amplification of the return echo from the fault impedance irregularity. The V-shaped peak appearing on the trace is of a sharp nature for an open circuit (upwards) and short circuit (downwards) whilst other impedance irregularities may not be as sharply defined. These other irregularities, such as from transpositions, change of route conditions, line spacings, small section loading, etc., do not cause such severe echoes as direct open or short circuit faults. Distant measurements are taken to the top of the echo peak occurring on the screen. When the fault is observed, the gain control is further increased until the echo appearing on the trace is approximately the same height as the vertical pips appearing on the trace. By operation of the expand control knob, R15 in the time base circuit, the fault condition is made to take up a position close to the right-hand end of the screen. Thus the maximum expansion of the distance to the fault can be viewed across the screen. By altering the expand and range knobs, a particular section can be selected for optimum viewing and the best possible determination of the exact distance to a particular fault. Alternatively, a graduated length of perspex may be placed over the screen to enable the distance to a fault to be determined with greater accuracy. It should be noted that severe fault conditions, such as short circuits, produce large reflections, and it is possible that the reflected echo may not be fully absorbed in the terminating resistor of the instrument. This has the effect that a portion of the echo returning from the fault is reflected back into the line and produces an echo at a distance past the fault equal to that from the testing station to the fault. This echo is smaller than the echo produced by the fault condition, and it may occur several times beyond the fault. In many cases these secondary echoes are readily discernible from the proper fault echo. A majority of lines will display small impedance irregularities owing to the route and circuit conditions existing. Depending on the type of the irregularities, these echoes appear either as downward or upward deflections on the horizontal trace. Thus, if possible, graphs of each trunk line connected to a station should be recorded to indicate particular impedance irregularities peculiar to each line. By this means, any fault condition occurring on a trunk, however large or small, can be speedily checked with the recorded trace, giving a visual indication of small variations. Various line diagrams and conditions are shown in Figs. 5-8. The instrument proves very valuable in enabling faulty circuits to be returned to normal traffic in the shortest possible time.

Fig. 5 — Trace showing an Open-circuit Fault 37.5 Miles from Testing Station on a 12-channel Open Wire Bearer. The Test was made on Line Side of 32 Kc/s Line Filter.

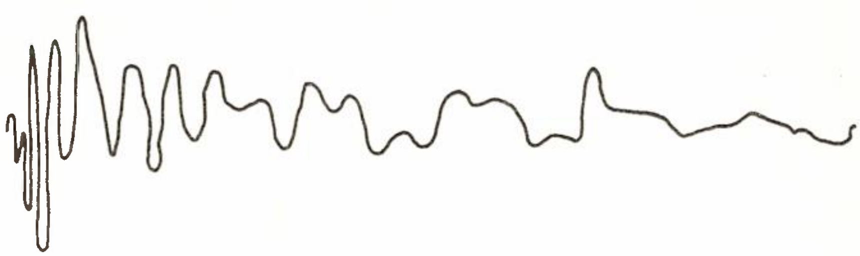


Fig. 6 — Trace of Fault in Fig. 5 Tested Through 32 Kc/s Line Filter. (This Shows Need to Bypass Filter.)

Fig. 7 — Trace Showing Short-circuit Fault, 37.5 Miles from Testing Station, on Same Circuit as Fig. 5.

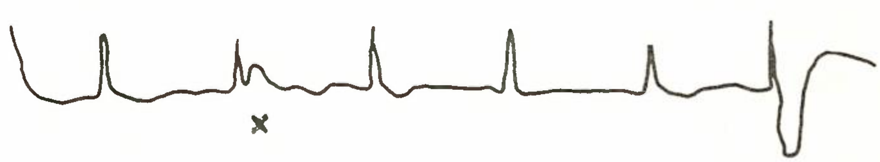


Fig. 8 — Trace for a 31 Mile Physical Circuit. At Point X, the Construction Changes from 6 inch Spaced 200 lb. Copper to 9 inch Spaced 100 lb. Iron.

TRANSISTOR VOLTAGE ALARM

*J. E. SANDER, B.E. (Hons.)**

This circuit is designed as a self contained unit having only two input terminals. The terminals are connected across the voltage source to be monitored. If the voltage is higher than the value set by the initial adjustment then the relay contained in the unit will operate. If the voltage is lower than the preset value the relay will remain unoperated. There are two changeover contacts fitted to the relay and these are available for connection to an alarm, an emergency power circuit, or whatever may be required.

The design was developed primarily to replace the PF relay (positive battery fail alarm) on the AER rack in automatic exchanges, but it is expected to find many other applications requiring an adjustable voltage sensitive relay in the range 30 volts to 60 volts. With the circuit as shown the margin between "operate" and "release" voltages is dependent on the setting of the control RV1 and upon the gain of the particular transistors used. In general the reoperate voltage will be higher than the release voltage by less than one volt. This degree of sensitivity is adequate for the original application and is probably adequate for some applications to which contact voltmeters are presently put.

The circuit is very simple as can be seen from Fig. 1, and in fact it should be possible using suitable miniature components to mount all of it (excluding the relay) inside a standard 2 microfarad condenser can.

The design of the circuit can be considered in two sections.

(i) An adjustable voltage divider to sample the terminal voltage applied to the unit. This consists of R1, RV1 and R2.

(ii) A temperature compensated switching circuit to operate the relay when the voltage sample from the potential divider is greater than the voltage developed across the zener reference diode MR1. This makes up the remainder of the circuit.

The use of two complementary transistors (one NPN and one PNP) in the manner shown gives a circuit which does not appreciably change its input switching voltage as the ambient temperature varies. The base-emitter voltages of the transistors are effectively in series, from the potential divider through to the zener reference diode. As they are shown connected any temperature induced drift in the base-emitter voltage of one transistor is cancelled out by a complementary shift in the base-emitter voltage of the other provided both transistor junctions change through the same temperature range. By strapping together the outer metallic cases of each transistor the junctions can be induced to move through similar temperature ranges and in the prototype a shift in terminal switching voltage of only 0.9 volts was observed for a change in ambient temperature from 18°F. to 120°F. This amount of change can be attributed partly to the temperature coefficient of the zener reference diode and partly to the imperfect cancellation of the transistor base-emitter voltages. Over the temperature range 47°F. to 110°F. the total voltage variation was 0.3 volts on the prototype unit.

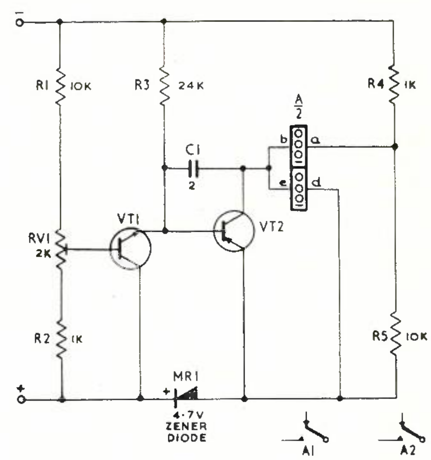


Fig. 1 — Transistor Voltage Alarm Circuit.

*Mr. Sander is a Group Engineer in the Telephone Equipment Laboratory at Headquarters.

RETIREMENT OF Mr. A. R. GLENDINNING

After 50 years' service with the Department, Mr. A. R. Glendinning, Sectional Engineer, Workshops, Headquarters Staff, retired on 15th January, 1960. "Angus", as he is well-known, started work as a Telegraph Messenger in Western Australia, serving later as a Telegraphist and Postal Clerk at such distant places as Meekatharra and Townsville. He was appointed engineer in 1937 and did sterling work in restoring service in the Townsville area after devastation by typhoon in 1940. During the war years Mr. Glendinning transferred to the General Works Section, Central Office, and worked on the coordination of the Workshops effort in the manufacture of communication equipment.

A farewell gathering was held at the Russell Exchange on Friday, 15th January, when Mr. Sawkins presented a chiming mantel clock and a gold wristlet watch to Mr. Glendinning from his fellow employees. A Scottish flavour was added to the gathering when the gifts were piped in by members of his staff.

We all join in wishing "Angus" a long and happy retirement.

THE DESIGN OF TRANSISTOR CIRCUITS

H. S. WRAGGE, B.E.E., M.Eng.Sc., A.M.I.E.E., A.M.I.E.Aust.

INTRODUCTION

In order to fully realise the possible advantages which may be gained from the use of transistors, it is necessary that sound design procedures should be adopted. It is also desirable that full information on the behaviour of the transistor should be available to the circuit designer. Basic design considerations for low frequency sine wave circuits and sufficient general information concerning transistor behaviour to enable the design of a particular circuit to be carried out to achieve desired characteristics (that is, low noise, wider bandwidth, low distortion, etc.) are discussed in this article. It must be borne in mind, however, that usually the fullest development of some desirable characteristics can only be obtained at the expense of some other characteristic. Detailed information concerning the mechanism of transistor action may be found in Refs. 1-3 and information concerning transistor circuitry is given in Refs. 2-6.

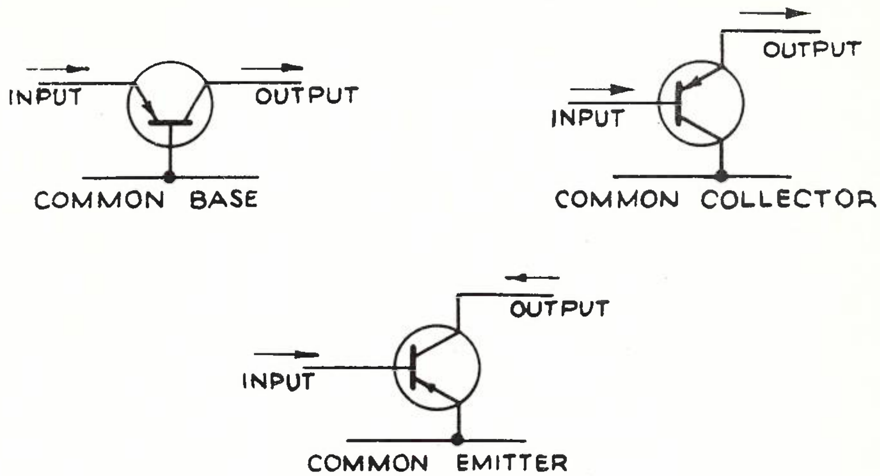


Fig. 2.—The Three Basic Operating Configurations.

CHARACTERISTICS AND PARAMETERS

The general behaviour of the transistor can be best described by families of characteristics in much the same way as valves. Whereas only one family of characteristics can completely describe the behaviour of the valve, at least two families are necessary to fully describe the behaviour of the transistor. This is due to the facts that the transistor has a low input impedance, causing definite input currents to flow, and also has a "feedback" factor, that is, variations of the output will affect the input.

The selection of the families of characteristics used to represent the transistor's behaviour is unimportant; however, some sets are more convenient than others. Those most used are the collector voltage and current with the input current as the specified parameter, and the base-emitter volt-amp. characteristic. The collector characteristics facilitate the use of load line techniques similar to those used for the design and analysis of valve circuits.

In addition to the static characteristics, the dynamic, or small signal, parameters of the transistor are necessary for the design and analysis of the transistor circuit. These may be derived from the static characteristics after the operating point has been chosen, or may be obtained from published data. As these parameters vary with operating point, corrections must normally be applied to published parameters if a different operating point is used to that for which the data was quoted. Methods of making these corrections will be described.

Static Characteristics

As has been stated, the behaviour of the transistor can be completely specified by any two independent families of characteristics. Those which are of most use are a family of collector characteristics with base current as the independent parameter, and the base-emitter characteristic, which is substantially constant for collector voltages in excess of 0.2 volt. These are illustrated in Fig. 1. These curves are sufficient to enable all the operating conditions to be obtained for a given operating point.

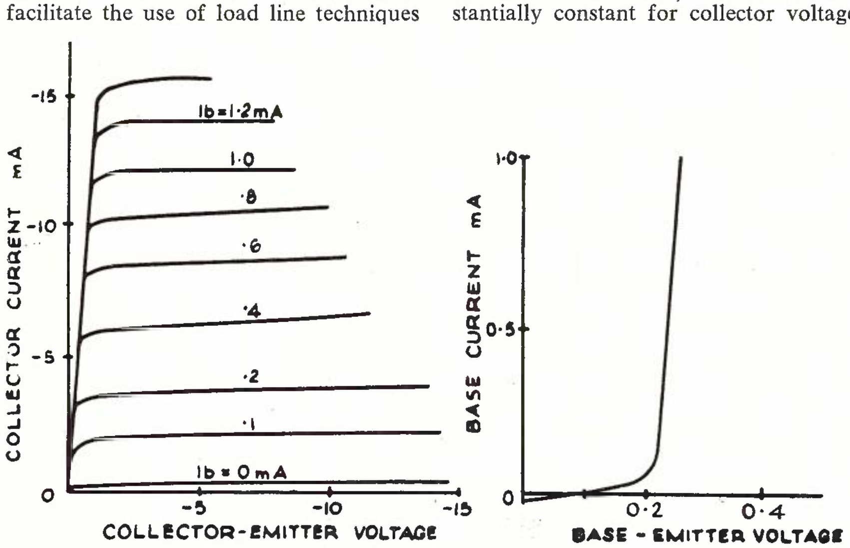


Fig. 1.—Common Emitter Static Characteristics.

The curves shown in Fig. 1 refer to the common emitter connection, in which the emitter is the common connection between the input and output, as shown in Fig. 2; this is the manner in which the transistor is usually operated. The curves shown can be redrawn as shown in Fig. 3 to permit load line techniques to be used when the transistor is operated in the common base configuration (see Fig. 2) and are frequently presented by manufacturers in this form, as this is the classical transistor configuration.

From an examination of the characteristics shown in Figs. 1 and 3, it may be seen that the collector voltage has very little effect on the collector current in the common emitter connection, and even less in the common base connection; also, the linearity of the collector current versus base or emitter current is good, even down to very small collector voltages.

The collector current for zero emitter current in the common base connections (conventionally designated I_{co}) is the sum of the saturation current of the reverse-biased collector junction and the leakage current across the surface of the transistor. The saturation current is substantially independent of the collector voltage for voltages in excess of about 0.2 volt, but the leakage current, which is extremely small for well manufactured transistors, is ohmic in nature, that is, it varies directly as the collector voltage. For a given collector voltage, I_{co} is the minimum collector current that can be obtained in a common base connected transistor; in the common emitter connected transistor, however, the collector current for zero base current (conventionally called I'_{co}) is β times I_{co}. (β is the common emitter current gain). This current, however, can be reduced from I'_{co} to a minimum of I_{co} by reducing the potential of the base. This is of little practical importance except where the transistor is being operated at temperatures where thermal stability is marginal.

The value of I_{co} is independent of collector voltage for voltages above 0.2 volts, but is strongly dependent on temperature. A rise of 12°C. in the junction of the transistor will double the value of I_{co} in a germanium transistor; a 16°C. rise is required to double I_{co} in a silicon transistor (Ref. 7). Under very adverse conditions, however, when I_{co} tends to become large (that is, at high collector dissipations), a rise of temperature will cause I_{co} to increase greatly, thus increasing the collector dissipation and also the collector temperature. If stabilisation precautions are not taken, this increase in I_{co} may build up to a value where the circuit becomes thermally unstable, and the collector current commences to increase at a very high rate; this is known as thermal runaway, and is liable to damage the transistor through overheating.

The emitter-base volt. amp. characteristic is also slightly dependent on temperature; as the temperature rises, the voltage necessary for a given current drops by about 2 millivolts per degree C. rise in temperature for both germanium and silicon transistors (Ref. 7) as shown in Fig. 4. This has the effect of lowering the beam (that is, the D.C.) input resistance of the transistor as the temperature increases. If the bias currents are derived from sources having a resistance comparable with or lower than the beam input resistance of the transistor, the bias currents will rise as the temperature rises tending to increase the dissipation in the transistor and thus increase the possibility of runaway. This effect can be reduced considerably by the use of appropriately designed bias networks.

Dynamic Parameters

When designing a transistor circuit, some knowledge of its small signal parameters is essential in order to ensure that satisfactory performance will be obtained. The small signal parameters may be obtained directly from the static characteristics, or published values may be used.

As the transistor is a non-reciprocal bilateral element, four independent parameters are necessary to completely describe the relations between its input and output voltages and currents. Of the many sets of general network parameters which may be used, the most convenient are the so-called "h" or "hybrid" parameters, which are defined by the following equations:—

$$E_1 = h_{11}I_1 + h_{12}E_2$$
$$I_2 = h_{21}I_1 + h_{22}E_2$$

where E_1 and I_1 are the input voltage and current; E_2 and I_2 are the output voltage and current; h_{11} and h_{21} are the input impedance and forward current gain respectively with the output short-circuited; h_{22} and h_{12} are the output admittance and reverse voltage gain respectively with the input open-circuited.

The h parameters are general network parameters, consequently they can be used to represent the transistor when it is connected in any of the three operating configurations, common base, common emitter or common collector; the respective values will be different in each case, however, and will be designated as follows in this article:

h_{11} etc. refers to common base
h'_{11} etc. refers to common emitter
h''_{11} etc. refers to common collector.

Two common abbreviations are α for $-h_{21}$ and β for h'_{21}, that is, α is the h_{21} and β for h'_{21}, that is, α is the common base short-circuited current gain and β is the common emitter short-circuited current gain.

Typical values of these parameters for the various operating configurations are as follows for a typical low power germanium transistor:

Formulae for the conversion from one set of parameters to another are given in Appendix II. However, it may be noticed that the value of the input impedance of the common emitter and common collector transistor is approximately β (that is, h'_{21}) times that of the common base transistor, whereas the open circuited output conductance is approximately β times greater.

The "h" parameters may be derived from the "4 quadrant" characteristics shown in Fig. 5. The characteristics shown in Fig. 3 appear in the upper right and lower left quadrants, with the other characteristics derived from them in the manner shown in Fig. 5. The gradients of the characteristics at a given operating point then represent the small signal "h" parameters for this operating

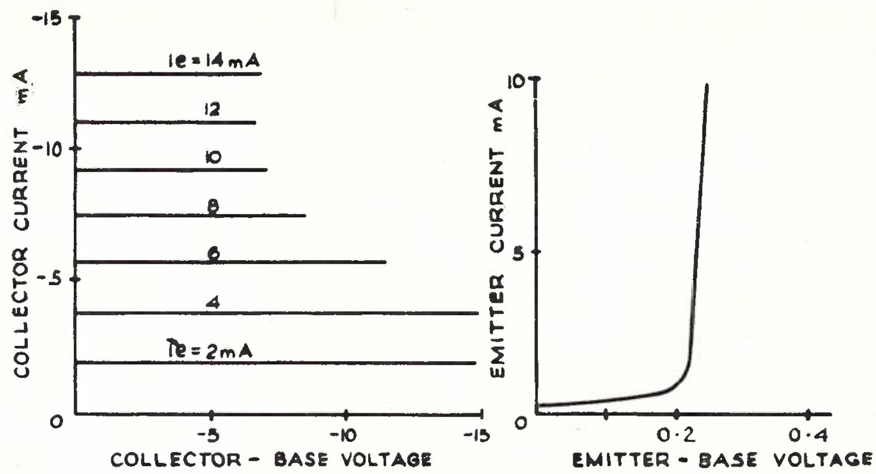


Fig. 3.—Common Base Static Characteristics.

	Common Base	Common Emitter	Common Collector
h_{11}	37	1850	1850 ohms
h_{12}	1.2×10^{-3}	0.6×10^{-3}	1.00
h_{21}	-0.98	49	-50
h_{22}	2.0	100	100 μmho

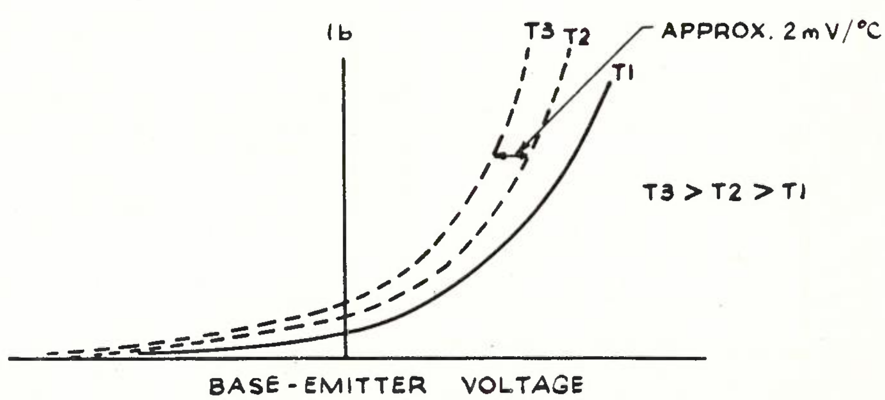


Fig. 4.—Variation of Base-Emitter Voltage with Temperature.

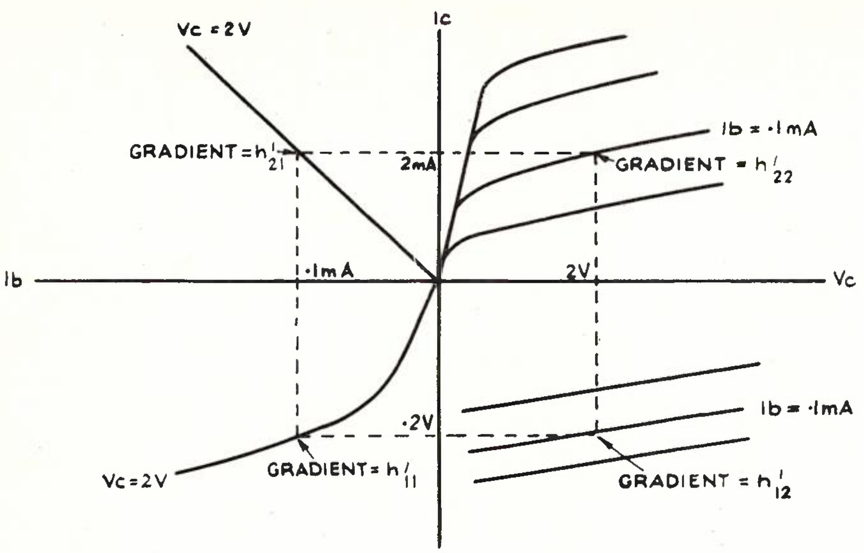


Fig. 5.—Derivation of A.C. Parameters from Static Characteristics.

point only. Care must be taken that the appropriate sign is associated with each gradient, or parameter; the simplest method of ensuring this is to adopt the signs given in the above table for junction transistors.

The small signal parameters of the transistors vary typically with operating point in the manner shown in Figs. 6 and 7. It may be seen that some parameters vary strongly with a change of operating point, whereas others vary only slightly. To a first approximation, the variations of the parameters for all configurations are as follows:—

$$h_{11} \propto 1/I_e$$
$$h_{12} \propto 1/V_c^{\frac{1}{2}}$$
$$h_{22} \propto I_e/V_c^{\frac{1}{2}}$$

As may be seen from Figs. 6 and 7, h'_{21} and h'_{12} vary in a rather unpredictable way with operating point, but h'_{11} and h'_{22} obey the above relations quite closely over most of the useful operating range.

Parameter variation curves similar to those shown in Figs. 6 and 7 exist for all junction transistors; usually the differences between those for different types is only very slight (Ref. 8).

Maximum Ratings

There are several maximum ratings which must not be exceeded if a transistor is to perform satisfactorily according to the manufacturer's data. These ratings, which differ for various transistor types, are always included among the manufacturer's data.

The most important of these are the junction temperature, the collector voltage and the emitter current.

1. Junction Temperature: The maximum junction temperature specified for a transistor is typically about 75° C. for germanium and 150°C. for silicon although other values are commonly encountered, depending on the impurity materials used. At higher junction temperatures, solid state diffusion of the impurity atoms will be greatly accelerated, resulting in a premature dropping off of the current gain. If the maximum junction temperature is sufficiently exceeded to cause melting of the parent material, the transistor will be immediately destroyed.

The junction temperature is dependent on three factors — the power being dissipated in the transistor, the thermal resistance between the junction of the transistor and the atmosphere, and the ambient temperature. The relation between these may be expressed in the form:—

$$T_j = T_a + P.R_T$$

where T_j °C. and T_a °C. are the junction and ambient temperatures respectively, P milliwatts is the power being dissipated in the transistor (the sum of the collector and emitter dissipation) and R_T is the resistance of the transistor, usually expressed as °C./mW. The thermal resistance and maximum junction temperature are usually given in manufacturer's data, thus allowing the maximum power dissipation for a given ambient temperature to be calculated from the relation

$$P_{max} = (T_{j\ max} - T_{a\ max})/R_T$$

The value of R_T may be decreased by the use of heat sinks, thus permitting a higher power dissipation to be attained for a given maximum ambient temperature. Data concerning the rating for various heat sinks is often given in manufacturer's data, especially for power transistors.

2. Maximum Collector Voltage: The collector voltage is limited by two factors; first, the voltage at which a breakdown occurs, and second, by "thermal runaway" conditions. The collector breakdown can occur in one of two ways, one of which is destructive. As the collector voltage is increased, the effective base width decreases, and for large voltages may reach zero, when "punch through" occurs and the collector and emitter usually are effectively short-circuited. The heavy current which then flows usually fuses the transistor, thus permanently destroying it. Other mechanisms leading to collector breakdown are the Zener and avalanche effect. As the collector voltage is increased substantially, all the voltage is developed across the junction, producing a very high potential gradient, which may strip electrons for conduction from the lattice in a manner similar to field emission in vacuum tubes. This is the Zener effect. In the avalanche process sufficient voltage gradient exists across the bulk of the crystal to enable conduction electrons to liberate additional electrons by impact; these electrons are then accelerated sufficiently to cause further electrons to be liberated and so on.

There is very little practical difference between avalanche and Zener breakdowns and it is frequently very difficult to differentiate between them. This type of breakdown is harmless to the transistor provided that the collector dissipation is kept within safe limits. The punch-through breakdown, on the other hand, is fatal to the transistor and must be avoided at all costs. Unfortunately, there is no definite way of knowing which effect will occur first as the voltage is raised. In general, it appears so far that the punch-through voltage is the lower in surface barrier transistors and the avalanche and/or Zener voltage is the lower in medium power p-n-p alloyed junction transistors (at about 50-80 volts).

Thermal runaway is dependent on the temperature dependent I_{co}, as mentioned earlier. If precautions to be described are not taken "runaway" takes place in which the collector current suddenly increases to a very high value, resulting in over-

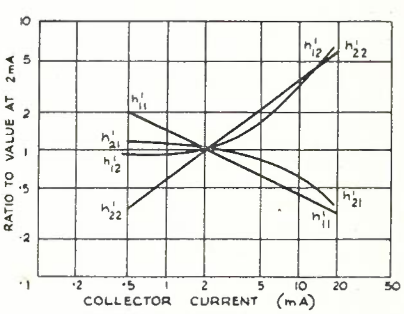


Fig. 6.—Variation of Parameters with Emitter Current.

heating of the transistor. Very high collector voltages tend to increase the possibility of runaway due to the increased collector dissipation.

3. Maximum Emitter Current: The maximum emitter current is not limited as precisely as the collector voltage or junction temperature, but is more of an arbitrarily chosen value which may be exceeded if due precautions are taken. As the emitter current is raised, the current gain rises to a maximum and then falls progressively as the current is raised, as shown in Fig. 8. At high currents, the lower current gain introduces distortion, and the maximum emitter current is chosen to keep this distortion to a reasonable value. Provided that the net dissipation (that is, the sum of the collector and emitter dissipation) is such that the maximum junction temperature is not exceeded, the emitter current may have any value. Due allowance must be made, however, for the lower current gain at high currents and all its consequent effects.

Variation with Temperature

The effect of varying temperature on the collector saturation current I_{co} and the input resistance have been discussed; the small signal parameters are also affected by temperature as shown typically in Fig. 9 (Ref. 9). There is a general tendency for the values to rise as the temperature rises; probably the most important variation is that of the current gain h'_{21} (that is, β) and its consequent effects.

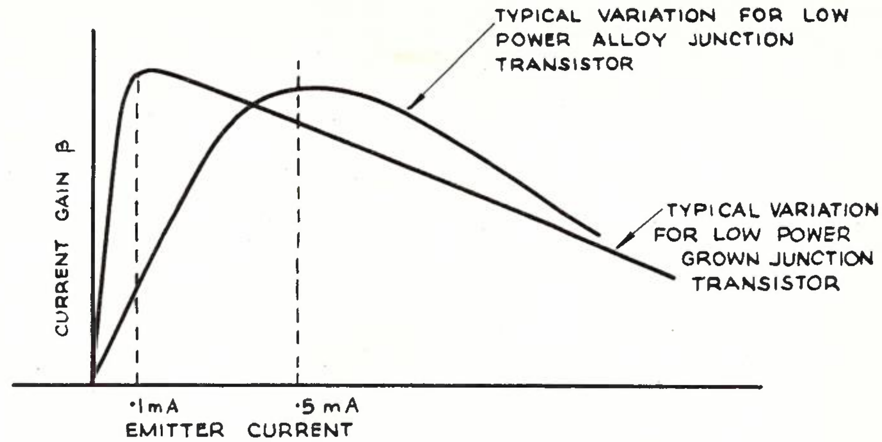


Fig. 8.—Variation of Current Gain with Emitter Current.

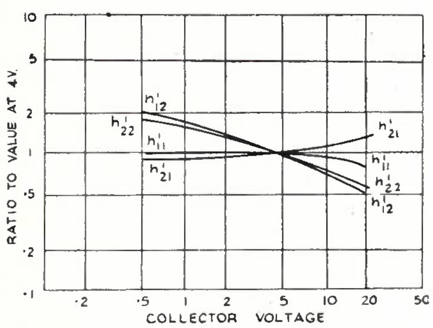


Fig. 7.—Variation of Parameters with Collector Voltage.

CHOICE OF CONFIGURATION

The transistor may be connected in any one of three operating configurations — common base, common collector, or as is most usual, common emitter. These were illustrated in Fig. 2. The choice of the actual configuration to be used varies with the application, as the properties of the stage differ considerably for each configuration. When dealing with class A stages (that is, those in which the collector current is always directly related to the base current), the most important factors to be considered are as follows:—

(a) stage gain
(b) input and output impedances
(c) upper frequency cut-off
(d) phase changes between input and output
(e) coupling circuits.

Stage Gain: The stage gain, for stages which have reasonably matched source and load impedance, is always greatest for the common emitter configuration for frequencies below the cut-off frequency. Approximate stage gains of the various configurations are summarised in the following table:

Methods of computing stage gain will be considered in detail at a later stage.

Stage Input and Output Impedances: The stage input and output impedances depend greatly on various factors, of which the most important are:—

(a) the actual values of the transistor parameters
(b) the load and source impedances
(c) any feedback loops.

The first and last of these will be discussed in a later section; load and source impedances are also discussed further later and the effects for a typical transistor stage are shown in Figs. 15 and 17. From these, it may be seen that the common base connection is typified by a large output impedance and a small input impedance whereas the reverse applies to the common collector stage. In the common emitter stage, the input and output impedances are much closer in value and are substantially less dependent on load and source impedances than those of the common collector and common base stages. The operating point also controls the input and output impedance of a transistor stage via the transistor parameters.

Upper Frequency Limit: The upper frequency cut-off of a transistor stage depends primarily on the value of the current gain of the transistor, which itself is governed by the operating configuration. In the common base stage, the controlling quantity is the current gain α, which commences to drop off at the α cut-off frequency f_α (usually of the order of 500 kc/s for low power audio transistors). The common emitter and common collector stages, however, are dependent on a much lower cut-off

	Common Emitter	Common Base	Common Collector
Voltage gain	$\dfrac{\beta R_L}{R_{in}} \simeq \dfrac{R_L{}^*}{R_E}$	$\dfrac{R_L}{R_{in}}$	1
Current gain	$\dfrac{\alpha}{1-\alpha} = \beta$	α	$\dfrac{1}{1-\alpha} \simeq \beta$

*If unbypassed external emitter resistance R_E is large compared with R_{in}/β.

frequency (sometimes termed the β cut-off frequency) which has the value $(1 - \alpha)\ f_\alpha$. (The cut-off frequency is the frequency at which the current gain has dropped by 3 db.) As a generality, the values of f_α for a given batch of transistors will be quite closely grouped; in high gain transistors, the values of α will be close to unity, consequently very small variations in α cause large variations of $(1 - \alpha)$; the values of the common emitter and common collector stage cut-off frequencies can thus vary over a wide range, being lowest for the highest gain transistor and vice versa. Typical variations of current gain for the various configurations are shown in Fig. 10.

Phase Changes Between Input and Output: There is no phase change between the input and output voltage of a common base or of a common collector transistor stage, but the phase is reversed in a common emitter. Directions of signal current flow are shown in Fig. 2.

Coupling Circuits: As the transistor has a definite input and output impedance, care must be taken to provide a reasonable degree of impedance matching between stages if maximum power gain is to be realised. This is a simple matter if appropriate transformer coupling is used exclusively between stages, regardless of their configuration. However, as transformers are relatively bulky, heavy and costly, it is desirable to use resistance capacitance coupling where possible. From Figs. 15 and 17 it may be seen that the input impedance of the common base stage and the output of the common collector stage can both be of the order of 25 ohm; if capacitance coupling is to be used, very large values of capacitance must be used if reasonable low frequency performance is required. The common emitter stage, on the other hand, has input and output impedances which are much more comparable in magnitude than those of the other configuration, and which lead to more reasonable coupling circuit element values if RC coupling is used.

General Considerations: The choice of which configuration should be used for a transistor stage will depend largely on the prevailing conditions and the judgment of the designer. As a general rule, the common emitter stage is the most useful configuration in low frequency amplification stages, mainly because of—

(i) its greater power gain
(ii) it is amenable to the use of RC coupling
(iii) it avoids the occurrence of widely fluctuating impedance levels.

These reasons, however, should not be allowed to dictate the universal use of the common emitter stage; the designer of a transistor circuit must also consider the other aspects discussed in detail in subsequent sections of this article.

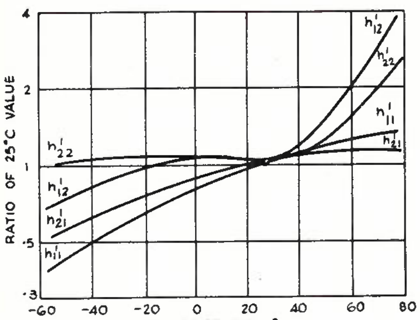


Fig. 9.—Typical Variation of Parameters with Temperature.

CONTROL OF THE OPERATING POINT

When a transistor circuit is being designed, the design is usually carried out with the aid of published average characteristics, and a suitable operating point is selected. When the circuit is wired, the tolerances on various component values and the deviation from average values of the transistor characteristics could result in the transistor having quite a different operating point from that intended, also temperature changes tend to produce operating point changes. It is desirable that this divergence of the operating point from its designed value should be held within reasonable limits for the following reasons:—

(i) to ensure that the desired voltage and current swings can be maintained;
(ii) to ensure that the maximum collector dissipation cannot be exceeded;
(iii) to maintain the A.C. parameters near their design values;
(iv) to ensure that circuits can be mass produced satisfactorily.

More caution is needed in stablising operating points in transistor circuits than in valve circuits for two main reasons: first, the transistor is often operated with voltage and current swings near their ideal maxima, and any divergence of operating point due either to component tolerances or temperature would seriously impair the operation of the associated circuitry. Secondly, transistors are usually operated far more closely to their maximum ratings, especially power dissipation, than are valves.

There are three main causes of deviation from the designed operating point; these are temperature effects, characteristic and parameter deviation from average, and ageing. The effects of parameter variations between transistors are obvious; ageing contributes a slight drift due mainly to a gradual decline in β and a gradual increase in I_{co}. Temperature variations change the values of I_{co} and the input beam resistance, both of which tend to shift the operating point in the same direction by increasing the collector current as the temperature rises. Fortunately all of these effects can be simultaneously reduced by suitable design of the bias network.

DESIGN OF BIAS NETWORKS

As the transistor has finite input and output resistances, two separate bias sources are required. These may be derived from either one or two batteries. Two battery bias is not often used, mainly because it necessitates the use of either a tapped battery or two batteries. An advantage of single battery bias is that both bias supplies are switched together, thus eliminating the possibility of one bias supply being applied in the absence of the other, which may be undesirable under some circumstances. Several two battery bias schemes are discussed for illustrative purposes.

Two Battery Bias Methods: Two methods of providing two battery bias are shown in Fig. 11. In the first, Fig. 11 (a), transformer coupling is used in the input and output of a common base stage. The emitter bias current is controlled by the resistor in series with the battery and is bypassed to A.C. by the capacitor. In addition to permitting the emitter current to be set to its required value, the emitter resistor also tends to swamp the input beam resistance of the transistor, thus reducing thermal effects. The input beam resistance will be low, typically of the order of 10-30 ohms in low power transistors, and less in higher power transistors. The collector voltage is equal to the battery voltage E_2; the collector current, which is substantially independent of collector voltage, is almost equal to the emitter current. Actually, the collector current i_c is given by—

$$i_c = \alpha i_e + I_{co}$$

where i_e is the emitter current and α is the common base current gain. The temperature dependent component, I_{co}, of the collector current in low power transistors is not likely to cause appreciable change in the collector current if the collector current is in excess of one milliamp.

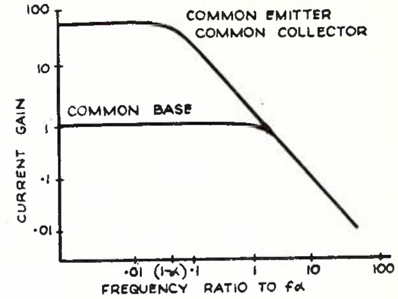


Fig. 10.—Variation of Current Gain with Frequency.

In the RC coupled emitter stage shown in Fig. 11 (b), the emitter current i_e and the collector current are both controlled by the battery E_1 and resistors R_1 and R_2. The collector voltage V_c is controlled by the battery E_2 and resistor R_L. The relations are:—

$$i_e = E_1/\{R_2(1 - \alpha) + R_1\}$$

$$V_c \simeq E_1 + E_2 - i_e\,(\alpha R_L + R_1)$$

The bias return resistor R_2 shunts the input of the transistor, and therefore should be kept larger than the input impedance in order to prevent excessive signal power loss. As will be shown in the next section, the transistor will be stabilised against β variations and thermal effects if $R_2\,(1 - \alpha) << R_1$. A suitable ratio is of the order of 1/10.

Single Battery Bias Methods: Three commonly used methods of providing base and collector bias are shown in Fig. 12. These bias methods apply mainly to class A common emitter stages, but can also be applied to the common collector stages by reducing the value of R_L to zero in each case.

In the "fixed bias" arrangements, Fig. 12 (a), base bias is applied via the base resistor R_1, thus establishing the magnitude of the collector current (β times i_b). After this has been established, the collector voltage may be established by appropriate choice of R_L. The emitter resistor R_2 has a stabilising influence in much the same way as a cathode bias resistor. If the collector current attempts to assume an excessive value, the voltage across R_2 will increase, reducing the voltage across R_1, thus tending to reduce the base current and collector current. The emitter resistor may be suitably bypassed to A.C. if the degeneration caused is objectionable.

The so called "self bias" method shown in Fig. 12 (b) is in effect the

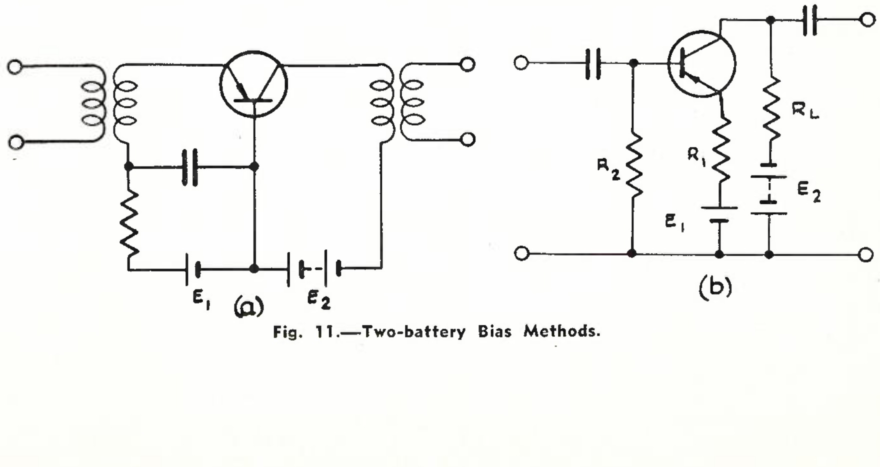


Fig. 11.—Two-battery Bias Methods.

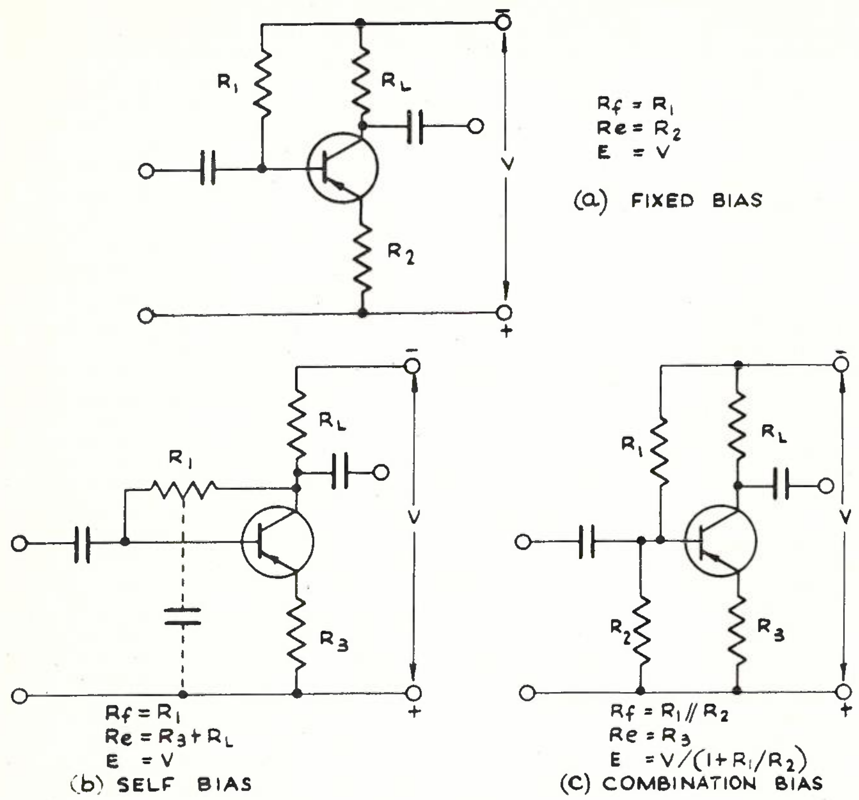


Fig. 12.—Single-battery Bias Methods.

same as the fixed bias method shown in Fig. 12 (a), as the position of R_L and the power supply may be interchanged so far as D.C. conditions are concerned, thus making R_L effectively in series with R_3. The circuit has very different A.C. properties due to the feedback provided by R_1 which produces a slight loss in gain and lower input and output impedances. This feedback may be obviated by decoupling R_1 as shown with the dotted capacitance in Fig. 12 (b); the capacitance should be connected to the mid-point of R_1 to avoid loading the input or output too severely. This also has the advantage of requiring the minimum capacitance for a given low frequency cut-off.

The "combination bias" method, shown in Fig. 12 (c) utilises a potential divider to maintain the base voltage at its desired value; as the emitter-base voltage e remains at about 0.2 volt over most of the useful operating range, the emitter voltage is held by the potential divider. The emitter resistor R_3 therefore controls the emitter and collector current for a given potential divider; as in the fixed bias circuit, the collector voltage is controlled by the value of R_L. In this circuit, the parallel resistance of R_1 and R_2 shunt the input of the transistor, and may cause significant power losses if their values are too low.

Comparative Performance of the Various Single Battery Bias Methods: It may be shown that for the three circuits described above the collector current is given by:—

$$i_c = \frac{E + (R_f + R_e)I_{co} - \beta e}{(\beta + 1)\,R_e + R_f}$$

The values of E, R_f and R_e for the various circuits are given in the following table:—

	Bias Fixed	Self Bias	Combination Bias
R_f	R_1	R_1	$R_1//R_2$
R_e	R_2	$R_L + R_3$	R_3
E	V	V	$V/(1 + R_1/R_2)$

The first term in the expression for the collector current represents a current caused by the supply voltage and ideal transistor action. The second term represents a component of collector current caused by the collector to base saturation current I_{co}. The third term represents a current due to the voltage e between the emitter and base. As was mentioned earlier, the magnitude of this voltage varies slightly with temperature, falling approximately 2 millivolts per degree C. rise in temperature, as is shown in Fig. 4; this voltage fluctuation will cause fluctuations in the collector current. (This voltage has a negative value, so will represent an increase in collector current).

If the bias circuit is designed so that the condition

$$(\beta + 1)\,Re >> R_f$$

holds, then the expression for the collector current becomes

$$i_c \simeq \frac{E}{R_E} + \left(1 + \frac{R_f}{R_e}\right) I_{co} - \frac{e}{R_e}.$$

Under these conditions, the major part of the collector current, E/R_e, is determined primarily by circuit constants and not by transistor parameters, in particular β. This is desirable if reproducible operating points are required from various transistors. The I_{co} component is always greater than I_{co}, but usually smaller than I'_{co} which is βI_{co}. Rising temperature reduces the value of e about 2 mV/°C., thus producing an increase of approximately 2 μA/millimho of emitter conductance per degree C. rise in temperature, thus the input resistance variation effect can be reduced to negligible proportions by sufficiently increasing R_e.

Further examination of the bias circuits shows that the inequality given above can only be met in the combination bias circuit. The inequality implies that all the supply voltage is developed across the emitter resistor R_2 in the fixed bias circuit, Fig. 12 (a), leaving no voltage drop across the transistor and load, which is clearly impractical. This type of circuit cannot be designed to provide independence of transistor parameters or as great a minimisation of thermal effects as is desirable. In the self bias circuit, Fig. 12 (b), the total supply voltage would be developed across the emitter and load resistors, but none would be developed across the transistor; this gain is impractical. However, better stabilisation can be obtained than in the fixed bias case. The combination bias circuit, however, can be fully stabilised and is the preferable bias circuit from the stabilisation viewpoint.

Design of Bias Network: Design procedures are suggested for each of these three bias circuits in Appendix III. It is not intended that these methods should always be followed rigorously, but they are given rather as an indication of good design principles. The load resistance, collector current and voltage swings, and supply voltage are assumed to be already known from other requirements.

PROPERTIES OF TRANSISTOR STAGES

Various properties of low power audio transistor stages, such as gain, impedances, noise, distortion and frequency response are examined with the object of showing their dependence on factors under the control of the designer.

Formulae are quoted in terms of the two most commonly encountered sets of parameters, the h, or hybrid, parameters, and the t parameters. The appropriate equivalent circuit for each set is shown in Fig. 13 (Ref. 10). Conversion formulae relating these sets of parameters are given in Appendix II. Typical values of these parameters are given in Appendix I and examples are also given for the stage properties to be described later.

Stage Gain

The term "stage gain", when applied to transistor circuits, usually implies a power gain. Having a finite (and comparatively low) input impedance, the transistor possesses a definite current gain in addition to a voltage gain and the power gain is the product of the current and voltage gains. These three gains are all very dependent on the load impedance of the stage concerned, as shown typically in Fig. 14 for the common emitter, common base and common collector stages at low frequencies. The common base stage has a current gain which is very close to unity, and the common collector stage has a voltage gain of approximately unity, consequently the power gain of the common base stage is substantially equal to its voltage gain, and that of the common collector stage is substantially equal to its current gain. The common emitter stage possesses a voltage gain substantially equal to that of the common base stage and a current gain which is almost equal to that of the common collector stage, hence its power gain is, in general, greater than that of the other two stages. The common emitter configuration is the only one which introduces a phase reversal.

Exact and approximate formulae are given in Appendices IV & V for the stage gain in terms of both t and h parameters. Typical values are given in Appendix I.

Low Frequency Stage Input Impedance

The input impedance of a transistor stage at low frequencies (that is, at frequencies less than β-cut-off) depends on—

(a) the parameters of the transistor,
(b) the load resistance,
(c) the bias network of the stage concerned.

The effect of the bias network may be twofold; first, it represents a passive impedance in parallel with the input impedance of the transistor, and secondly, it may modify the input impedance of the transistor by any feedback that it produces. For example, in the bias networks shown in Fig. 12, the bias resistor R_1 in the fixed bias method is in parallel with the input impedance of the transistor, which is modified by the emitter resistor R_2. In the self bias case, if R_1 is not bypassed, both R_1 and R_2 will contribute feedback modifications to the input impedance of the transistor.

The typical input impedance variation with varying load resistance is shown in Fig. 15; exact and approximate formulae are given in Appendices VI and VII in terms of both t and h parameters. Typical values are given in Appendix I.

Allowances may be made for series feedback by increasing the values of r_e or r_b as necessary in the t parameter formulae, or by using modified "h" parameters given in Appendix VIII to replace the transistor and feedback element. Fig. 16 refers.

As an example, consider the input resistance of the stage shown in Fig. 12 (a); the input resistance (for $R_L << r'_c$) is given by:—

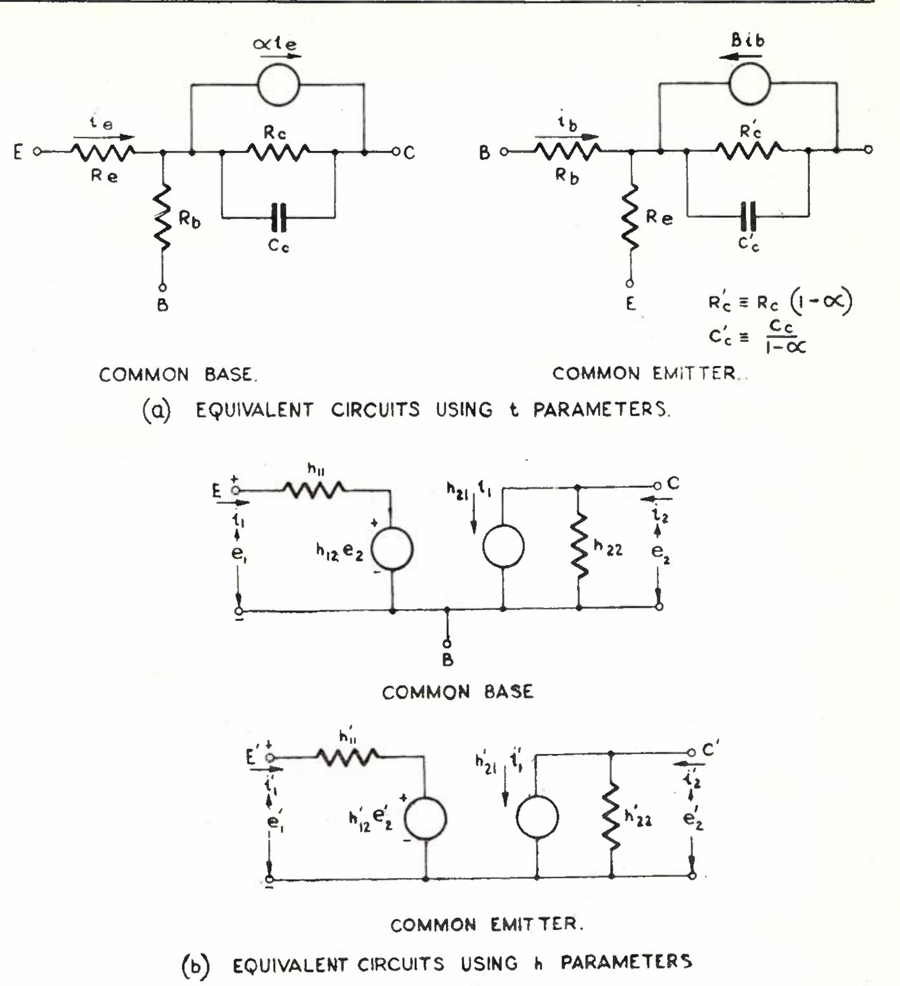


Fig. 13.—Equivalent Circuits.

$$r_{in} = \left(R_b + \frac{r_e + R_2}{1 - \alpha} \right) // R_1$$

(using t parameters)

$$r_{in} \simeq \{h'_{11} + R_2(1 + h_{21})\} // R_1$$

(using h parameters)

Low Frequency Stage Output Impedance

The output impedance of a transistor stage at low frequencies depends on:—

(a) the parameters of the transistor,
(b) the source impedance,
(c) the output networks,
(d) the bias network.

The output network usually applies an impedance in parallel with the output impedance of the transistor itself, for which due allowance must be made. The bias network affects the output impedance in the same way as the input impedance, and the same methods may be used in assessing the variation due to feedback; that is, modifying the t parameters in the series feedback case or using modified h parameters for either case.

Typical variation of the output resistance for varying source resistance is shown in Fig. 17. The relevant expressions are tabulated in terms of t and h parameters in Appendix IX. Typical values are given in Appendix I.

Impedance Behaviour at Medium Frequencies

As the operating frequency is raised, two effects become significant in the behaviour of the transistor. The capacitance of the collector junction becomes significant and modifies the output impedance and, to a lesser extent, the input impedance and stage gain. Also, an effect somewhat similar to transit time in vacuum tubes becomes apparent; the main result of this is an apparent drop in current gain, which becomes progressively greater as the frequency increases (see Fig. 10). This change in gain also affects the input and output impedance in the common emitter and common collector stages.

An equivalent circuit which is sufficiently accurate for frequencies up to the α-cut-off frequency (that is, the frequency at which the value of α has dropped by 3 db) is shown in Fig. 18. This is basically the same as the low frequency equivalent circuit shown in Fig. 13; the only difference is the addition of the frequency sensitive current generator αi_e.

A different equivalent circuit which is sometimes used is shown in Fig. 19. In this equivalent circuit a frequency independent current generator is used, and a capitance C_e representing the emitter junction barrier and diffusion capacitance is added. The current generator is controlled by the current flowing in r_e, not the total current i_e flowing into the emitter. Both equivalent circuits yield equivalent results; however, the former is slightly easier to handle analytically.

The collector junction capacitance C_c varies slightly with applied collector e.m.f. according to the following approximate relationship (Ref. 11).

$$C_c \propto V_c^{-\frac{1}{2}}$$

for pnp alloyed junction transistors.

$$C_c \propto V_c^{-\frac{1}{3}}$$

for grown junction transistors.

Typical values of collector capacitance for low frequency low power transistors are of the order of 35 $\mu\mu$F.

The frequency dependence of the current gain may be sufficiently accurately represented by the expression:

$$\alpha = \alpha_0/(1 + jf/f_\alpha)$$

for frequencies up to the α-cut-off frequency $f\alpha$ (α_0 is the low frequency value. This is shown in Fig. 10. Typical values of f_α are of the order of 500kc/s for low frequency low power transistors.

This dropping-off of the common base current gain with rising frequency causes the current gain in the common emitter or common collector stages to commence dropping at much lower frequencies, as shown in Fig. 10. It may be shown that the frequency at which both the common emitter and common collector currrent gain has dropped by 3 db is $(1 - \alpha_0)\, f_\alpha$.

Operation of the transistor at "high" frequencies (that, is above α-cut-off) is not recommended for amplifiers and will not be discussed in this article.

1. Power Gain at Medium Frequencies: The power gain at medium frequencies is modified in three different ways:

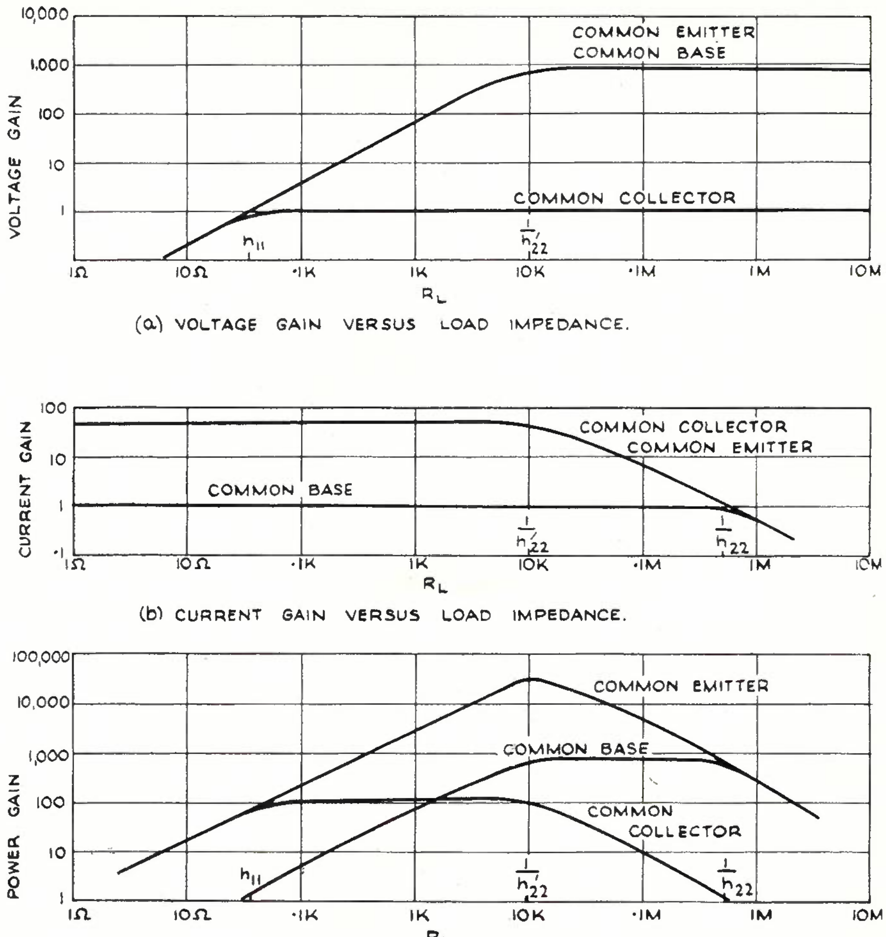
(a) the input impedance is modified,
(b) the current gain is modified,
(c) the output impedance is modified.

The magnitudes of these effects depend partly on the transistor parameters and strongly on the load and source impedances.

The effect of input impedance variations on power gain will largely depend on the ratio of the source impedance to the input impedance. If the source impedance is large in comparison with the input impedance, the input current will be substantially independent of input impedance variations. However, if the source impedance is relatively small, input impedance variations will cause corresponding variations in the input current, thereby varying the apparent stage gain.

The effects of changes of current gain are obvious; they will cause corresponding power gain changes.

The effect of output impedance changes depends on the ratio of load impedance to output impedance. If the output impedance is large in comparison with the load impedance, variation in output impedance will not have significant effects; however, if the output impedance is relatively small, variations will cause corresponding variations in power gain, that is, a drop in output impedance will cause a drop in power gain due to the shunting effects of the collector capacitance.



(a) VOLTAGE GAIN VERSUS LOAD IMPEDANCE.

(b) CURRENT GAIN VERSUS LOAD IMPEDANCE.

(c) POWER GAIN VERSUS LOAD IMPEDANCE.

Fig. 14.—Variation of Stage Gains with Load Impedance.

2. Input Impedance at Medium Frequencies: Approximate expressions for input and output impedances have already been given in Appendices V and VII in terms of t parameters. If a reasonably precise estimate of the input impedance is required, these expressions may be used, provided the validity condition is not violated, if the following substitutions are made:—

(i) $\dfrac{r_c}{1 + j\omega C_c}$ is substituted for r_c

where ω is the pulsatance and C_c is the collector capacitance.

(ii) $\dfrac{\alpha_0}{1 + jf/f_\alpha}$ is substituted for α

where is α_0 is the low frequency value of α and f_α is the α-cut-off frequency.

The general behaviour of the input impedance at medium frequencies is shown in Fig. 20, together with the significant break points and limiting values. It is of interest to note that the input impedance of a common base stage rises with frequency (and is inductive) whereas the input impedance of the common emitter and common collector stages falls (and is capacitive) with increasing frequency. It may also be noted that the input impedance of these two stages commences to fall at $(1 - \alpha_0)\, f_\alpha$, the same frequency at which the current gain for these two stages commences to fall.

3. Output Impedance at Medium Frequencies: Approximate expressions for the output impedance have already been given in Appendix VII in terms of t parameters. If it is necessary to determine the output impedance these expressions may be used with the substitutions given in 2 above for α and r_c. The general behaviour of the output impedance for the three configurations is shown in Fig. 21, together with significant break points and limiting values. The only approximations made are that the source impedance is small in the common emitter case, and much less than r_c in the common collector case.

It is of interest to note that the output impedance commences to drop at the same frequency for both the common emitter and common base stages (and is capacitive) but commences to rise at a higher frequency (and is inductive) in the common collector stage.

4. Operation in the Medium Frequency Range: Owing to the reactive effects and varying magnitude of the current gain and input and output impedances which are encountered at frequencies in excess of $(1 - \alpha_0)f_\alpha$, it is suggested

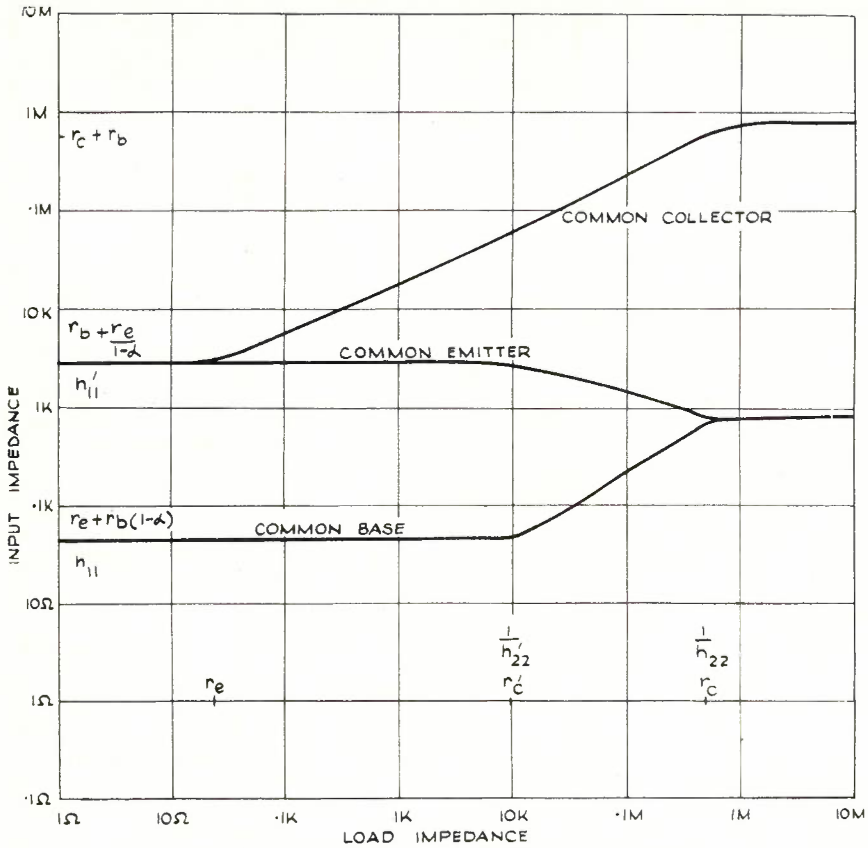


Fig. 15.—Variation of Input Impedance with Load Impedance.

that transistors should not be operated for general use at these frequencies, unless possibly at one constant frequency. The correct solution where operation at a higher frequency is necessary to use a high frequency transistor.

However, if it is essential to operate a transistor at frequencies in excess of $(1 - \alpha_0)$ f_α, the following principles should prove helpful in maintaining a constant power gain in a common emitter stage.

(a) Use a small source impedance. As the frequency rises above $(1 - \alpha_0)f_\alpha$, the input impedance will fall; a voltage generator would cause an input current to flow which would rise at 6db/octave above $(1 - \alpha_0)f_\alpha$. This will counteract the 6db/octave fall in β above this frequency, thus tending to maintain a constant output current.

(b) Use a small load impedance. This will prevent the generator current flowing through the load impedance being shunted by the collector capacitance. The critical time constant in this case is that formed by the load resistance and β times the collector capacitance.

(c) If a rise in stage gain is required at some particular frequency f_1, this can be obtained by using a bypassed emitter resistor. The time constant should equal $1/f_1$. The frequency f_2 at which the gain ceases to rise is equal to f_1, multiplied by the ratio of the gain with the bypassed emitter resistor to that with the unbypassed emitter resistor (see Fig. 22).

Transistor Noise

The noise factor of a transistor varies with source resistance, collector current and collector voltage. Typical values of noise factor for commonly available transistors operated under near optimum conditions are of the order of 6 db, but low noise transistors having noise factors as low as 3 db are available. In practice, the noise factor is essentially the same for all three operating configurations (Ref. 12, 13).

Transistor noise consists of two components, one being essentially white noise and the other being "1/f" noise or "semiconductor" noise in which the noise power per octave is constant, that is, the noise power per cycle rises according to a 1/f law at low frequencies. With the higher grade transistors now being manufactured, the 1/f noise only becomes apparent at frequencies of the order of 100c/s. The 1/f noise is believed to be due to surface effects, one of which is collector to base leakage; in general, higher leakage is associated higher 1/f noise. At higher frequencies the noise factor again increases, due to the reduced gain of the transistor. The noise factor is generally minimum for a source resistance of about 500 to 1000 ohms. Typical variation of the noise factor is shown in Fig. 23.

For a transistor with low collector leakage, the noise factor is reasonably independent of collector voltage for potentials under about 10 volts. If leakage is present, the noise factor increases strongly with an increase in collector voltage, also the 1/f noise is greatly increased. A practical rule-of-thumb procedure is to use about 3 volts on the collector of input stages.

The noise factor is minimum for an emitter current of the order of 0.5mA, and rises as the emitter current rises. In actual fact, it is minimum at the value of emitter current at which the current gain is maximum. Typical variation is shown in Fig. 24.

The noise factor of a transistor is usually specified for a narrow bandwidth, usually near 1000 c/s for a given source resistance, of the order of 500 ohms. This figure is useful for frequencies between those at which the 1/f noise becomes apparent, and the current gain commences to drop, but is of not much use outside this range.

Distortion

Distortion in class A amplifiers can arise from two main sources, as follows:—

(a) the non-linear input impedance;
(b) non-linearity of the transfer characteristic.

The distortion in class A transistor stages which are not driven into clipping is composed almost entirely of second harmonic, odd and higher order even harmonics being almost negligible.

The non-linear input impedance generates substantially all the stage distortion; the transfer characteristic contributes a small fraction of the overall distortion which may either aid or oppose the input impedance distortion. In common base stages, the transfer characteristic (that is, α) distortion is insignificant; in common emitter and common collector stages, the transfer characteristic (that is, β or $(\beta + 1)$ respectively) distortion will oppose the

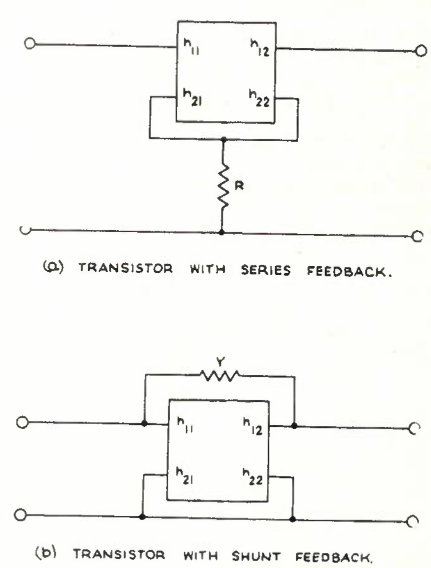


Fig. 16.—Networks with Feedback Elements.

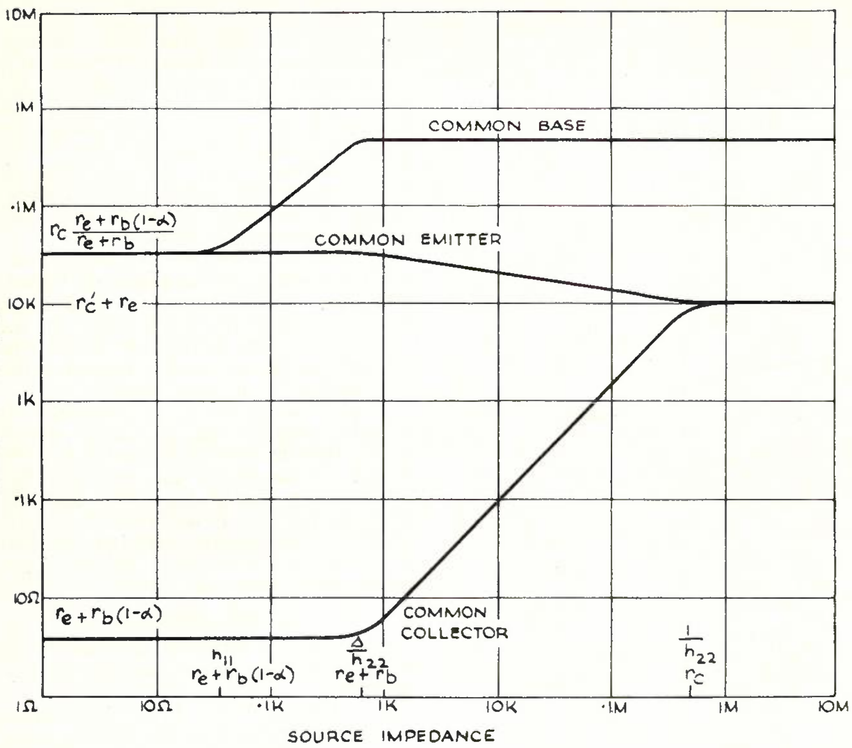


Fig. 17.—Variation of Output Impedance with Source Impedance.

input circuit distortion if the emitter current is greater than that for which β is maximum (see Fig. 8) and will aid if the emitter current is less than that for which β is maximum. For low power alloy junction transistors this current is of the order of ½ milliamp and is less for grown junction types.

At high emitter currents the transfer characteristic distortion increases and may even exceed the input circuit distortion. This, however, would only take place when emitter currents of the order of 100 mA are flowing or if the source resistance is high, thus tending to limit the input impedance distortion.

In low power stages where the input impedance is the dominating factor, the distortion depends primarily on the generator impedance and the ratio of the A.C. input current to the quiescent input current. The distortion is directly proportional to the ratio of input A.C. current to quiescent input current and drops as the generator impedance is raised.

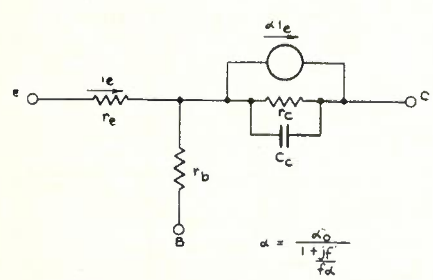


Fig. 18.—Equivalent Circuit for use in the Medium Frequency Range.

In a common emitter state with by-passed emitter resistor, the percentage input impedance distortion D may be empirically expressed as:—

$$D = \frac{32}{1 + (R_g + r_b)I_e/25\beta} \cdot \frac{i_s}{i_B} \cdot \%$$

where R_g is the generator resistance in ohms;

r_b is the base resistance (obtained from manufacturers' data) in ohms;

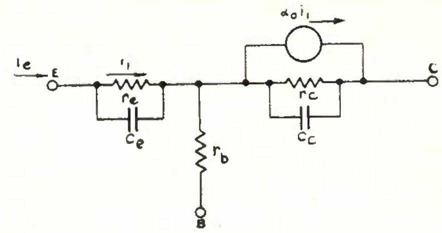


Fig. 19.—An Alternative Equivalent Circuit for use in the Medium Frequency Range.

I_e is the emitter current in milliamps;

i_s is the r.m.s. input current;

i_B is the quiescent base current expressed in the same units as i_s;

β is the current gain.

The distortion in a typical common emitter transistor stage usually is of the order of from 4% to 8% depending on operating conditions.

Where using the above expression, it must be borne in mind that the value of R_g is the output impedance of the driving source in parallel with the resistance to ground of the bias network. For example, in the case shown in Fig. 12 (c), R_1 and R_2 are both effectively in parallel with the internal impedance of the driving source when considering the source impedance driving the transistor.

CONCLUSION

Various aspects of transistor circuitry have been discussed with the object of indicating the degree of dependence of various circuit properties on circuit configurations, transistor parameters, operating conditions, etc., so that the designer of a transistor circuit can appropriately influence the behaviour of his circuitry. In general, all the desired properties of a circuit cannot be realised, and some desirable attributes must be sacrificed to

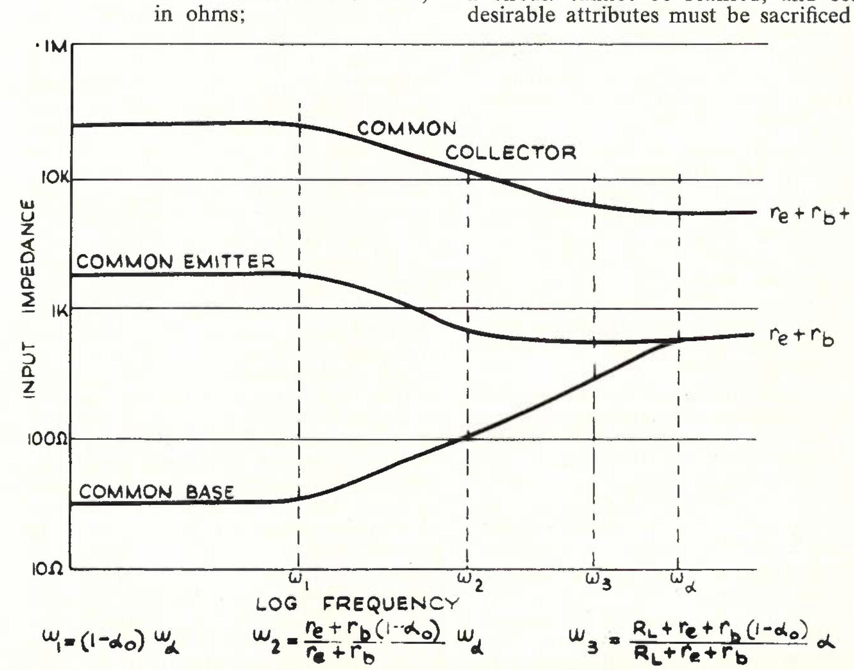


Fig. 20.—Variation of Input Impedance with Frequency.

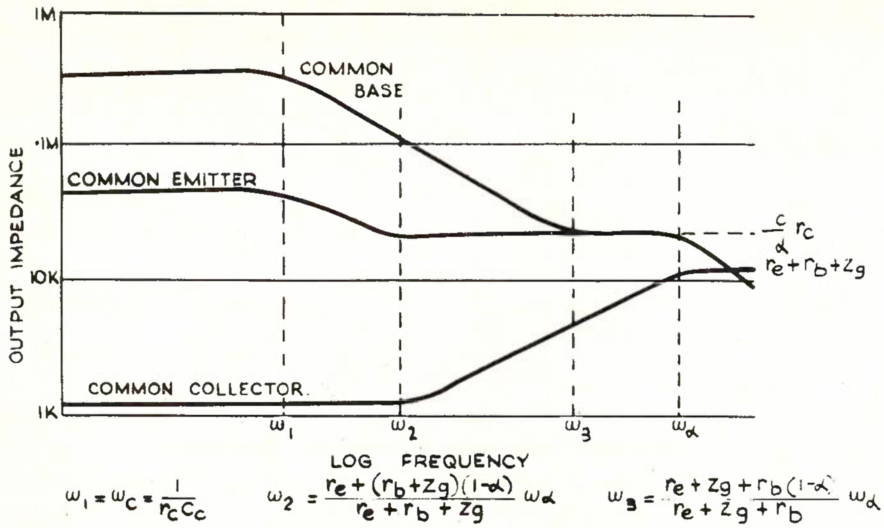


Fig. 21.—Variation of Output Impedance with Frequency.

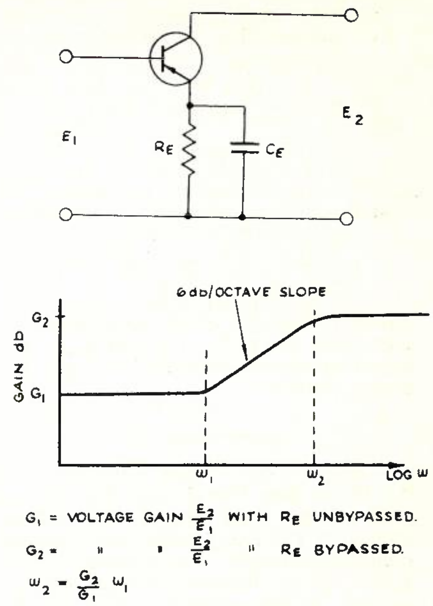


Fig. 22.—Effects of Unbypassed Emitter Resistance.

enhance the properties which are of most importance. The degree to which this compromise must be met rests in the final analysis with the designer, and it is hoped that the information contained in this paper will be sufficient to enable the more simple class A stages to be designed to meet specified needs.

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APPENDIX I

Typical Values for a Low Power Low Frequency Amplifier Stage

1. Transistor Parameters—

	h Parameters			t Parameters	
	Common base	Common emitter	Common Collector		
$h_{11}(\Omega)$	37	1850	1850	r_e	25 Ω
h_{12}	1.2×10^{-3}	2.5×10^{-3}	1.00	r_b	600
h_{21}	−0.98	49	−50	r_c	500 kΩ
h_{22}(mho)	2×10^{-6}	100×10^{-6}	100×10^{-6}	α	0.98
Δ	1.25×10^{-3}	62×10^{-3}	49.8		

$(\Delta = h_{11} h_{22} - h_{12} h_{21})$

2. Stage Properties—

Property	Common Base	Common Emitter	Common Collector
Generator Impedance	50 Ω	600 Ω	5000 Ω
Load Impedance	50 KΩ	5000 Ω	5000 Ω
Input Impedance	91 Ω	1440 Ω	167 KΩ
Output Impedance	64.5 KΩ	20.1 KΩ	49 Ω
Voltage Gain	490	−113	1.00
Current Gain	0.89	−32.7	33.3
Power Gain	436 (26 db)	3700 (36 db)	33.3 (15 db)

APPENDIX II

Conversion Formulae for Transistor Parameters

Common base h parameters in terms of t parameters—

$h_{11} = r_e + r_b (1 - \alpha)$
$h_{12} = r_b/r_c$
$h_{21} = -\alpha$
$h_{22} = 1/r_c$

2. t parameters in terms of h parameters—

$\alpha = -h_{21}$

$$r_e = h_{11} - \frac{h_{12}(1 + h_{21})}{h_{22}}$$

$r_b = h_{12}/h_{22}$
$r_c = 1/h_{22}$

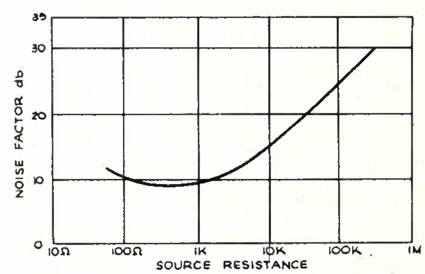

Fig. 23.—Variation of Noise Factor with Source Resistance.

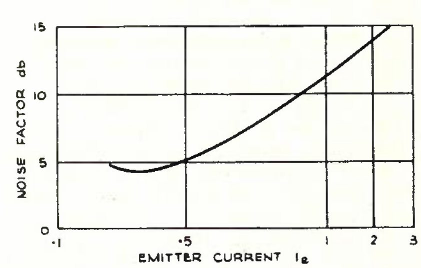

Fig. 24.—Variation of Noise Factor with Emitter Current.

3. Common emitter parameters in terms of common base h parameters—

$h'_{11} = h_{11}/(1 + h_{21})$

$h'_{12} = \dfrac{h_{11}\,h_{22}}{1 + h_{21}} - h_{12}$

$h'_{21} = -h_{21}/(1 + h_{21})$

$h'_{22} = h_{22}/(1 + h_{21})$

4. Common collector parameters in terms of common base h parameters—

$h''_{11} = h_{11}/(1 + h_{21})$

$h''_{12} \simeq 1$

$h''_{21} = 1/(1 + h_{21})$

$h''_{22} = h_{22}/(1 + h_{21})$

Note: All of the above equations (except those for r_e, r_b, h'_{22} and h''_{22}) are approximations, accurate enough for design purposes.

APPENDIX III

Bias Circuit Design Procedure

A. Fixed Bias Circuit (see Fig. 12 (a))

1. Having due regard to voltage swings in the load, select the quiescent collector voltage V_c and collector current I_c so as to develop as high a D.C. voltage V_1 as possible across R_2, remembering
   (a) distortion increases as the ratio of peak current swing to quiescent value increases;
   (b) input and output impedances fall as collector current increases.

It is desirable in the interests of thermal stability to develop at least one volt across R_2.

2. Calculate R_2 from:

$R_2 \simeq V_1/I_c$

3. Calculate R_1 from:

$R_1 = (V - V_1 - 0.2)/I_b$

where the quiescent base current I_b is derived from characteristics. If these are unobtainable use I_c/β as I_b.

4. If necessary bypass R_2 with a suitable capacitor.

B. Self Bias Circuit (see Fig. 12 (b))

1. Select operating point in same way as for fixed bias case, developing the stabilising voltage V_1 across R_3.
2. Calculate R_3 from:

$R_3 \simeq V_1/I_c$

3. Calculate R_1 from:

$R_1 \simeq \{V - i_c(R_3 + R_L) - 0.2\}/I_b$

where the quiescent current I_b is derived from static characteristics. If these are unobtainable use I_c/β as I_b.

4. If necessary, bypass R_1 and R_3 with suitable capacitors.

C. Combination Bias Circuit (see Fig. 12 (c))

1. Select operating point in same way as for fixed bias case, developing the stabilising voltage V_1 across R_3.
2. Calculate R_3 from:

$R_3 \simeq V_1/I_c$

3. Find quiescent base current I_b from static characteristics or use I_c/β.
4. Calculate R_2 from:

$R_2 = (V_1 + 0.2)/kI_b$

k should be as large as practicable; a satisfactory value for good stabilisation is about 10; however, as k increases, R_2 will decrease and absorb power being fed into the stage.

5. Calculate R_1 from:

$R_1 = (V - V_1 - 0.2)/(k + 1)I_b$

6. If necessary bypass R_3 with a suitable capacitance.

APPENDIX IV

Stage Gains in Terms of h Parameters

Gain	Exact Expression	Approximate Expression	Validity Condition for Approximate Expression
Voltage	$\dfrac{-h_{21}Z_L}{h_{11} + \Delta Z_L}$	$\dfrac{-h_{21}Z_L}{h_{11}}$	$h_{11} << Z_L << 1/h'_{22}$
Current	$\dfrac{-h_{21}/Z_L}{h_{22} + 1/Z_L}$	$-h_{21}$	$h_{11} << Z_L << 1/h'_{22}$
Power	$\dfrac{h_{21}^2}{(h_{11} + \Delta Z_L)(h_{22} + 1/Z_L)}$	$\dfrac{h_{21}^2 Z_L}{h_{11}}$	$h_{11} << Z_L << 1/h'_{22}$

$\Delta = h_{11}h_{22} - h_{12}h_{21}$

APPENDIX V

Approximate Stage Gains in Terms of t Parameters

Gain	Common Emitter	Common Base	Common Collector
Voltage	$\dfrac{-\alpha Z_L}{r_e + r_b(1-\alpha)}$	$\dfrac{\alpha Z_L}{r_e + r_b(1-\alpha)}$	1
Current	$\dfrac{\alpha}{1-\alpha}$	α	$\dfrac{1}{1-\alpha}$
Power	$\dfrac{\alpha^2 Z_L}{(1-\alpha)\{r_e + r_b(1-\alpha)\}}$	$\dfrac{\alpha^2 Z_L}{r_e + r_b(1-\alpha)}$	$\dfrac{1}{1-\alpha}$

Validity Condition: $r_e << Z_L << r_c(1-\alpha)$

APPENDIX VI

Transistor Input Impedance in Terms of h Parameters

Configuration	Exact	Approximate	Validity Conditions for approximate formulas
Common Emitter	$\dfrac{\Delta' + h'_{11}/Z_L}{h'_{22} + 1/Z_L}$	h'_{11}	$1/Z_L >> h'_{22}$
Common Base	$\dfrac{\Delta + h_{11}/Z_L}{h'_{22} + 1/Z_L}$	h_{11}	$1/Z_L >> h'_{22}$
Common Collector	$\dfrac{\Delta'' + h''_{11}/Z_L}{h''_{22} + 1/Z_L}$	$h''_{21}Z_L + h''_{11}$	$1/Z_L >> h'_{22}$

APPENDIX VII

Approximate Transistor Input Impedance in Terms of t Parameters

Configuration	Input Impedance	Validity Conditions
Common Emitter	$r_b + r_e/(1-\alpha)$	$Z_L << r_c(1-\alpha)$
Common Base	$r_e + r_b(1-\alpha)$	$Z_L << r_c(1-\alpha)$
Common Collector	$r_b + (r_e + Z_L)(1-\alpha)$	$Z_L << r_c(1-\alpha)$

APPENDIX VIII

Transistor h Parameters Modified by Feedback

(Refer to Fig. 16)

Unmodified Parameters	Modified Parameters	
	Series Feedback	Shunt Feedback
h_{11}	$\{h_{11} + R(1 + h_{21})\}/A$	h_{11}/B
h_{12}	$(h_{12} + R\,h_{22})/A$	$(h_{12} + h_{11}/Z_L)/B$
h_{21}	$(h_{21} - R\,h_{22})/A$	$(h_{21} - h_{11}/Z_L)/B$
h_{22}	h_{22}/A	$\{h_{22} + (1 + h_{21})/Z_L\}/B$

$A = 1 + h_{22}R \qquad B = 1 + h_{11}/Z_L$

APPENDIX IX
Output Impedance in Terms of h and t Parameters

Configuration	h Parameters	t Parameters
Common Base	$\dfrac{h_{11} + Zg}{\Delta + h_{22}\,Zg}$	$r_c \dfrac{r_e + Zg + r_b(1 - \alpha)}{r_e + r_b + Zg}$
Common Emitter	$\dfrac{h'_{11} + Zg}{\Delta' + h'_{22}\,Zg}$	$r_c \dfrac{r_e + (1 - \alpha)(r_b + Zg)}{r_e + r_b + Zg}$
Common Collector	$\dfrac{h''_{11} + Zg}{\Delta'' + h''_{22}\,Zg}$	$r_e + r_c(1 - \alpha)\dfrac{r_b + Zg}{r_c + Zg}$ $\simeq r_e + (r_b + Zg)(1 - \alpha)$ if $Zg << r_c$

SHIFTING THE ALIGNMENT OF THE MELBOURNE-GEELONG TRUNK CABLES

*I. C. SMITH, A.M.I.E. Aust.**

INTRODUCTION

Telephone communication between Melbourne and Geelong is by two cables laid together for a route length of about 46 miles. The cable is armoured and laid directly in the ground for most of the length except in Melbourne and Geelong suburban areas and the township of Werribee, where cables are unarmoured and laid in ducts. The cable sizes are 3/40 + 126/20 and 2/40 + 96/20, known as T1 and T2 respectively, are quad local type (except the 40 lb. pairs which are twin) and contain both voice frequency and carrier circuits. Reference 1 gives details concerning the reasons for providing circuits in this way and the construction of the cables.

These cables are laid parallel with the Princes Highway between Melbourne and Geelong as shown in Fig. 1 and located with respect to the Highway as in Fig. 2. Although efforts were made when the cables were installed to ensure that they were clear of future road works (2), when the Country Roads Board (The C.R.B.) was preparing plans for the duplication of the Highway, the designing engineers found that at various points, totalling about three miles, the cables would lie within the formation of the proposed Highway. The Postmaster-General's Department was therefore requested to shift the cable to allow road works to proceed. For simplicity reference hereafter to "the cable" may be taken to mean both cables unless there is some specific statement to the contrary.

HISTORICAL

The Melbourne-Geelong cables were installed in 1937-38 to replace an aerial route which had reached the end of its useful life and to meet the growing traffic requirements between the two centres. The cable was laid at a mean depth of 18" (depth varied between 12" and 24", but mostly 18") by mole plough in a trench 20" wide and 30" deep previously dug by a ditcher and backfilled. The rocky nature of the country made ploughing directly into the ground impracticable and this, together with the need to avoid open trenches on a stock route led to this unusual method of cable laying (2).

Initially the cables contained voice frequency circuits only, operating on a 4-wire basis, i.e., T1 carried the Melbourne-

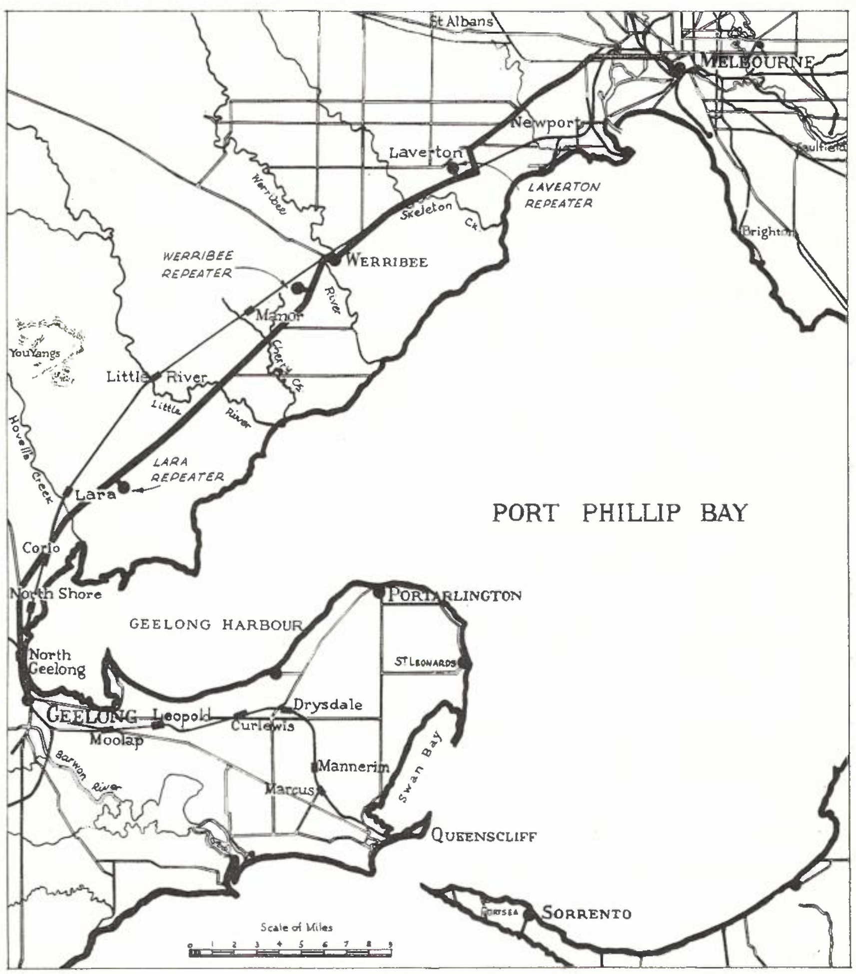

Fig. 1.—Route of Melbourne-Geelong Cable.

*Mr. Smith was formerly a Group Engineer at Geelong and is now attached to Methods and Training Section, Central Office.

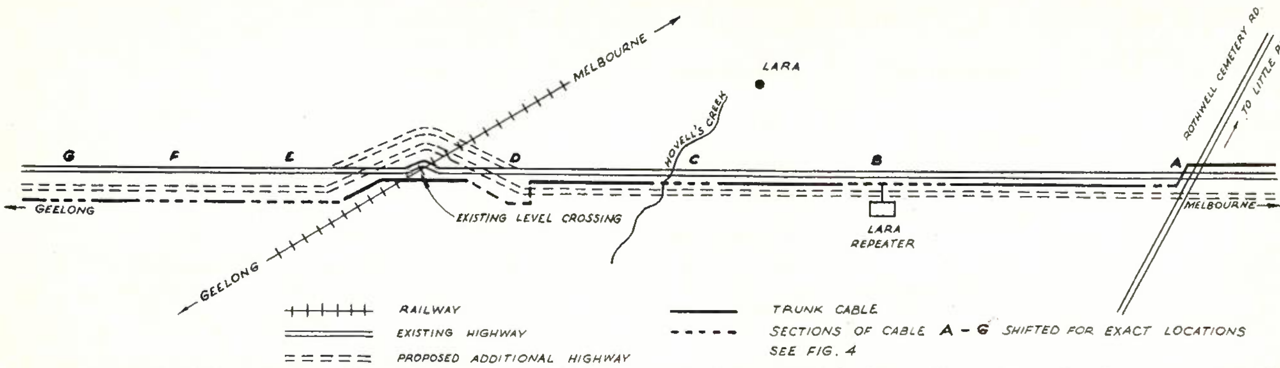

Fig. 2.—Relationship between Trunk Cable and the Princes Highway between Geelong and Werribee and approximate location of sections shifted.

Geelong and T2 carried the Geelong-Melbourne "leg" of each circuit with terminal receiving amplifiers at either end; 70 pairs in T1 and 70 pairs in T2 were loaded between Melbourne and Geelong. Of the 56 unloaded pairs in T1, 30 were originally intended for short trunk services along the route of the cable and these were later loaded, making a total of 100 loaded pairs in T1 and 70 loaded pairs in T2 between Geelong and Melbourne. Twenty-six pairs in each cable were left unloaded. The five 40 lb. programme circuits were all capacity balanced and loaded between Melbourne and Geelong with the exception of one circuit in T2. The 20 lb. pairs in each cable were jointed systematically both within loading sections and between loading sections to ensure uniform low capacity unbalance between quads. In addition, capacity unbalance within quads (side to side) was minimised by within quad balancing every 1,500 ft. Voice frequency loading provided was 88 millihenrys at 9000 ft.

When the demand for circuits between Melbourne and Geelong increased beyond the capacity of the 4-wire circuits during the war years, selected 4-wire circuits were converted to 2-wire circuits, with "22" type (two way two wire) voice frequency repeaters installed at the Werribee Post Office. In the immediate post-war period the demand for further circuits became pressing so the use of 12 channel systems operating on a four wire basis on the unloaded pairs was planned. This involved constructing repeater stations at Laverton, Werribee and Lara to give the 11 mile repeater spacing necessary to keep the attenuation at the highest frequency of operation (60 Kc) within the limits of available gain and ensure adequate signal to noise ratio. In accordance with usual practice, the unloaded pairs were "frogged" between cables. "Frogging" means that the group of carrier bearer circuits occupy pairs alternatively in T1 and T2, changing the pairs from one cable to the other at each repeater station. This procedure which was adopted to reduce the effect of progressive energy transfer along the length of the cable is illustrated in Fig. 3.

It will thus be appreciated that the arrangement of circuits in these cables is rather complicated making alterations to the working cables difficult.

At the date of cutover of the Melbourne-Geelong cable there were 88 telephone and 5 programme channels working. By contrast the cable now contains in addition to the 5 programme circuits, 205 telephone channels as follows:—

Melbourne-Geelong	160
-Western Victoria	30
-Tasmania	15

These figures include the circuits carried by nine 12 channel systems. They also include fourteen Melbourne-Geelong and eighteen Geelong-Melbourne transit circuits. As a matter of interest a forecast of 182 circuits for 1957 was made when the cable was installed in 1937.

ASSESSMENT OF THE PROBLEM

During the course of the design of the highway duplication by the C.R.B., the alignment of the cable was surveyed with a "Tedco" cable locator and after agreement that an accuracy of cable location of plus or minus one foot would be satisfactory, the C.R.B. used this information to evolve their final design. At this stage, it became apparent that the cable would have to be shifted at various points along its length. The information for all sections is summarised in Figure 4.

The problem was therefore to devise methods to shift the cable to the new positions specified, at the lowest cost. At various points it would be necessary to joint in lengths of new cable and at other points to joint out lengths of existing cable. Since the cable was required to operate continuously, all jointing and shifting operations would have to be performed on working circuits without causing even momentary interruption which would affect voice frequency signalling and teletype circuits superimposed on some of the telephone circuits. Jointing and testing schedules would be necessary to minimise jointing errors in the complicated inter-connection of cable pairs and ensure that if by any chance a jointing error was made, it would be quickly detected by the testing. Thus although the purpose of the operation was to relocate the cable at reasonable cost, the over-riding requirement was to ensure that continuity of service was maintained in this vital communication link.

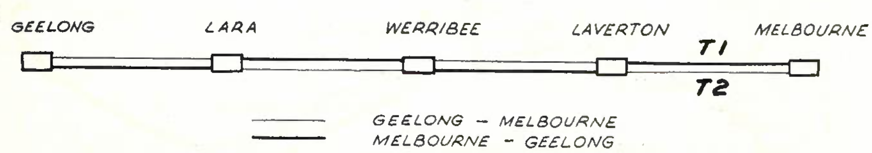

Fig. 3.—Layout of "frogging" of unloaded bearer pairs for four wire carrier circuits in Melbourne-Geelong cables.

Examination of drawings and schedules, and inspections on site revealed the extent of the problem which is summarised in Figs. 2 and 4. Shifting operations could thus be divided into the following categories—

1. Shift up to 38 ft. laterally from existing location at normal depth to a new location at normal depth (length involved approximately 2 miles).
2. Shift up to 2 ft. laterally from location at normal depth to new location 6 feet below ground surface level.
3. From existing location at normal depth, deepen up to 4 ft. 6 ins. below ground surface level. No lateral shift required.
4. Shift up to 63 ft. laterally from location at normal depth to new location up to 16 ft. below existing ground surface level.
5. Relocate length of cable at creek crossing.
6. Prior to lateral shifting outside Lara repeater, joint in four lengths of cable to extend all cables leading into the repeater.
7. Joint in lengths of cable to make a crossing under new highway.

By far the greatest length involved shifting cable from the existing location at normal depth to the new location at normal depth so this seemed to offer the greatest possibility for evolving a method to reduce excavating costs. A C.R.B. heavy tandem road grader was tried for this work and found to be very satisfactory for excavating the new trench. Besides possessing adequate power and weight, the blade supported within a long wheel base afforded stability and control not available in excavators of the bulldozer type where the blade is not

Chainage (feet)	Mean Distance from Melbourne (miles)	Vicinity	Length of Section		Maximum Distance to be shifted (feet) Right or Left Direction: With back to Melbourne.	Reason for Shifting	Work Required	Section on Fig. 2.
			Feet	Miles				
167,200 to 167,600	32.5	Immediately on Geelong side of Rothwell Cemetery Rd.	400	.076	10 R	To provide for acceleration lane near intersection	Unearth trench and relay at normal depth. Shift manhole 10 ft. sideways. Slack available.	A
177,327 to 182,313	34.1	Lara Repeater	4,426	.838	38 R	To follow deviation of new road necessary to avoid repeater station.	Unearth trench and relay T1 & T2, also 10 pr. subs. cable at normal depth. Lengthen each cable 70 ft. (35 ft. on each side of repeater), provide new 6-duct road crossing and manhole in median strip.	B
192,991 to 194,300	36.7	Top of hill near Hume & Hovell Memorial on to Geelong side of Hovell's Creek.	1,309	.248	63 R	To follow new highway which will cross Hovell's Ck on new bridge. 16 ft. to be removed from top of hill.	Extensive cutting works proposed. First shift cable alignment to new position between 193,600 and 194,300. This involves removal and relaying cables in Hovell's Ck. and on the banks either side. Sufficient slack available. After C.R.B. cutting completed, shift cable over remainder of distance, and relay at normal depth. Joint out slack.	C
201,600 to 202,360	38.9	North of Corio Overpass	760	.144	33 L	Road crossing from median strip to position east of new road.	Unearth, trench and relay at normal depth, cut in new lengths of cable and lay in troughing to form undercrossing for new highway. Provide one manhole and one No. 6 pit.	D
206,480 to 208,300	39.5	Opposite Percy St. on Geelong side of Corio rail crossing.	1,820	.345	11 L	To remove cable outside 10 ft. shoulder of highway into median strip between highway and service road.	Unearth, trench and relay at normal depth. Adequate slack available.	E
209,800 to 213,100	40	Geelong side of School Road.	3,300	.625	2.5 L	Ditto	Unearth, trench, relay & lower cable to about 6 ft. in places. Adequate slack available.	F
214,200 to 216,200	40.7	Geelong side of Harpur Road.	2,100	.398	6 L	Ditto	Unearth, trench and relay at normal depth, excepting in section south of Harpur Rd. where lower 3 ft. 6 in. Adequate slack available.	G

Fig. 4

so located. The speed at which it could operate was impressive, being about 700 ft./hour for a 1 ft. 10 in. deep "V" shaped trench (Fig. 15(a)). In addition to its digging capabilities the grader is unrivalled for filling trenches. However, as the cable was found to weave about unpredictably in the ground both vertically and horizontally, use of the grader to excavate it from its existing location involved a high probability of damaging the cable.

A device to constantly indicate the relationship between the grader blade and the cable to enable the blade to be adjusted accordingly was therefore required. This would minimise the chance of damaging the cable and at the same time permit excavating as close as practicable to the cable. The Research Section was therefore asked to develop an electronic locator which would indicate the depth of the cable at any point and so enable safe excavation over the cable. It was thought that such a device could be attached to the grader blade and be used to guide the operation.

CABLE LOCATION

The location technique which resulted is based on the principle that if an alternating electric current of known value is passed through a cable sheath buried in the ground, the resultant magnetic field produced around the cable may be detected by a suitable coil, amplifier and indicator system on the surface of the ground. Suitable geometrical relationships between three coils and calibration of a detector makes accurate location and depthing of the cable practicable. Fig. 5(a) shows the coil and detector, and Fig. 6, the power supply. The detector circuit is shown in Fig. 5(b).

Although originally it was thought that the locator could be attached directly to a grader and used to control excavation over the cable, (Fig. 15(b)) immediately this was tried it became apparent that such a procedure was impracticable for the following reasons:—

1. The depth and location of the cable in the ground varied unpredictably so that even with the locator operating satisfactorily, excavating over the cable was difficult to control.
2. Even if the excavation were taken to within 2 ins. of the cable, it was necessary to complete the excavation by hand methods before the cable could be removed. This was not only expensive and laborious but the use of picks and other hand tools in close proximity to the cable in the early stages caused more damage than mechanical excavating.
3. Operating with a 12 ton machine over the cable was extremely dangerous as sudden subsidence of ground under the wheels was likely to thrust the grader blade downwards and probably damage the cable.

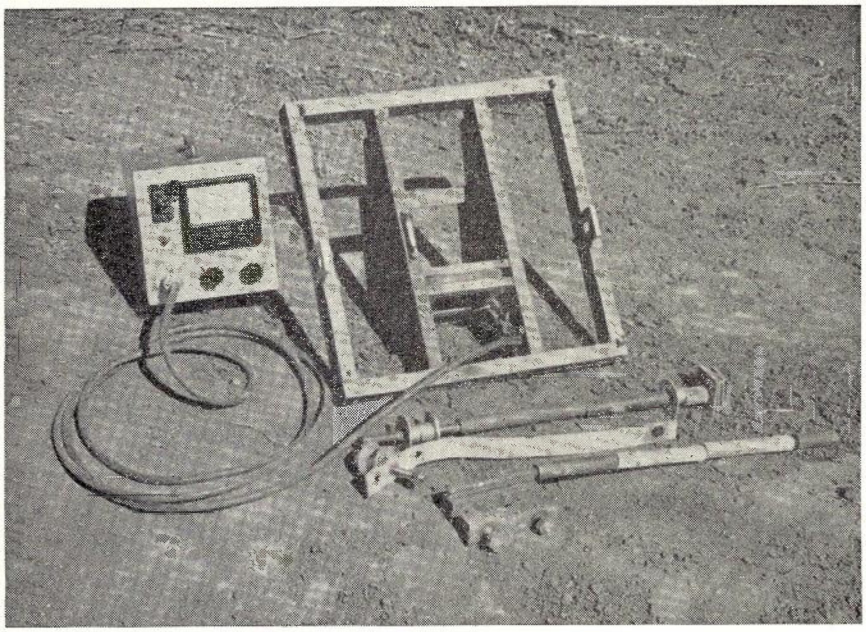

Fig. 5 (a).—Amplifier Detector, Locating Coil System and Clamps for mounting coil system on Grader.

4. Digging over the cable with the grader removed any survey pegs which had been placed in the ground above the cable to indicate its location.

In view of these disadvantages, survey and pegging of the location and depth of the cable every ten feet along its length before any excavation was attempted was decided and the equipment designed for attachment to the grader adapted for this purpose. In addition, to avoid the dangers inherent in operating the machine over the cable, the decision to dig only to the side of the cable (Fig. 15(c)) was made. This technique is explained in detail later.

Locating Equipment

The cable locating equipment consists of:—

1. Power Supply. The power supply is derived from a genemotor operating from 24 volts D.C. batteries and delivering, via a transformer, 3 amps at 10 volts 60 cycle A.C. applied to the sheath of one cable between two points (Fig. 6). The return circuit from the end of the cable section remote from the source is made via insulated wire along the ground spaced a sufficient distance from the cable so that the magnetic field over the cable is not appreciably affected by the field in the return wire. Close adjustment of sheath current is by a variable feeding rheostat.
2. Locating Coil System. The coil system consists of three coils mounted as shown in Fig. 7. Coil A is used in conjunction with Coil B to indicate the location of the cable and with Coil C to indicate the depth. They are connected to the detector by a flexible lead.
   For location of the cable, Coil A is held horizontal, Coils A and B are connected in series and a switch is fitted to enable Coil A to be short circuited. The current induced in both coils is thus fed to the detector. When the centre of Coil A is immediately over the cable, the current induced in it is zero and the meter indication is not changed when it is short circuited. When the coil system is in other positions the current from Coil B is either in or out of phase with that from Coil A and so either adds to or subtracts from the total current shown on the meter, indicating that the coils must be moved to one side of the cable or the other. By operating the short circuiting switch the system can be quickly adjusted to indicate the location of the cable as shown by no variation of the meter when the switch is in either position.
   For depth measurement Coils A and C are connected in series and for a given relationship between the coil system and the cable, the detector may be calibrated provided the cable is to the right of plane YY (Fig. 7). A number of curves showing relationship between the coil system and the cable, and meter indication are illustrated in

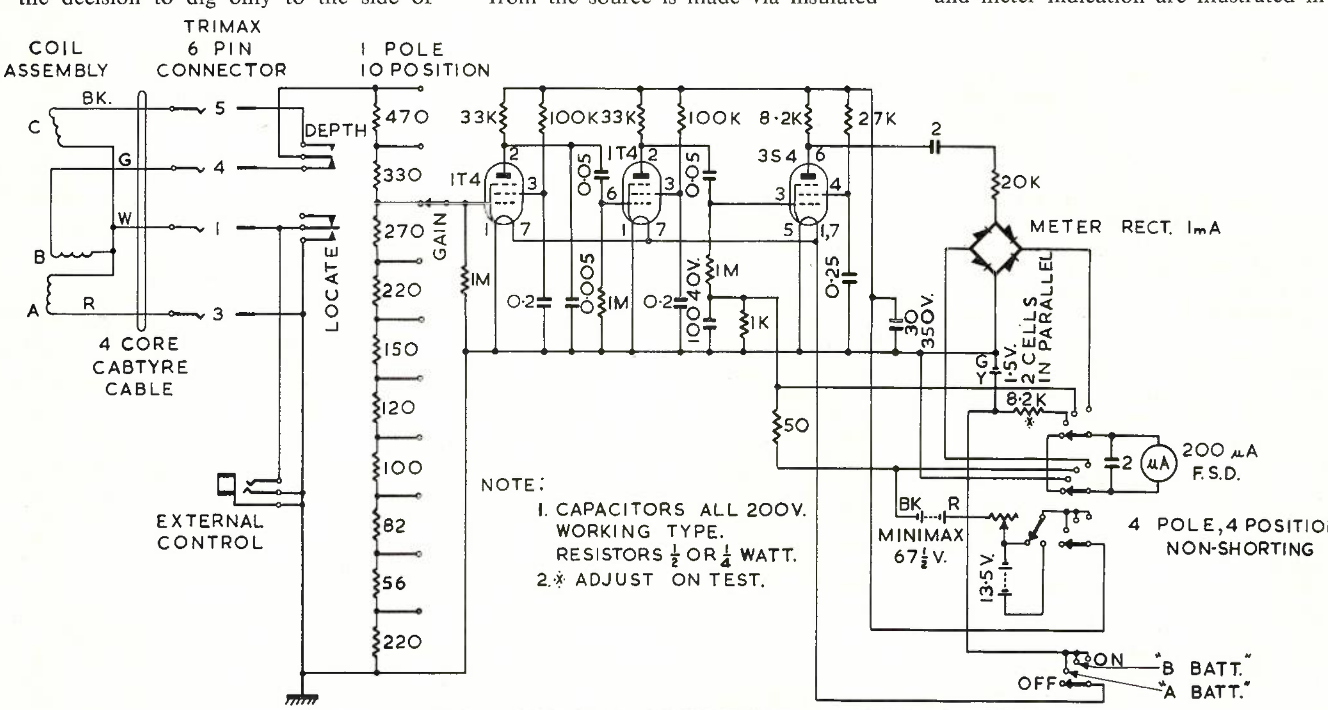

Fig. 5 (b).—Circuit arrangement of Amplifier Detector and Locating Coil System.

Fig. 8. For surveying the route of the cable for location and depth, the location is first determined and the coil system held horizontally is moved across the cable so that the right hand edge of coil A is vertically over the located position for the cable. As the depth of the cable increases, the signal received by the coil system decreases and so consequently does the meter deflection. A curve relating the depth of cable to meter deflection may be derived from the curves in Figure 8 but in practice a special calibration was made.

The use of the coil system to guide excavation is illustrated in Figs. 9, 10 and 13. Referring to the curves in Fig. 8, the curve corresponding with d = 5½ is flat between X = 8 to X = 24. This means that when the coil system is mounted on the grader blade with plane YY in line with the cutting edge as in Fig. 9, when the plane YY of the blade is 5½ in. from the cable, a constant deflection will be indicated on the meter independent of the position of the cable along plane YY provided X lies between 8 in. and 24 in. By suitably positioning the coil system relative to the blade, the effective working edge can be included within these dimensions. For the coil system to function effectively to indicate the location of the cable, when attached to the grader, it was necessary to approach from one side of the cable, gradually deepening the "V" trench until the correct depth and distance from the cable was reached (Fig. 10). This method was highly satisfactory for controlling the proximity of the grader blade to the cable whilst excavating.

3. Detector. The detector consists of an amplifier whose output is rectified and fed to a microammeter. By means of a key on the unit the coil connections may be switched to provide "locate" and "depth" facilities as described in the previous section. Other controls provide for variation of the gain of the amplifier and for calibration check and adjust to compensate for ageing of the amplifier batteries. The circuit is illustrated in Fig. 5(b). A discussion of the theoretical considerations underlying the operation of the locator and information concerning its construction is contained in Reference 3.

Preliminary Survey

To facilitate preliminary survey of the cable, the vehicle shown in Figures 11 and 12 was constructed from bicycle parts by a local cycle manufacturer. This provided for holding and levelling the coil system, for moving it laterally across the line of the cable, and also for moving it along the line of the cable. The detector was mounted in front of the operator who could push the locator along and steer it. In this way the cable was located about every 10 ft. and pegs were driven in the ground at the locations so determined. When checked by potholes the pegs were generally found to be accurate within plus or minus one inch. The cable was located in lengths of about a quarter of a mile, (dictated by jointing pit spacing) although the maximum distance determined by the resistance of the return lead was somewhat greater than this.

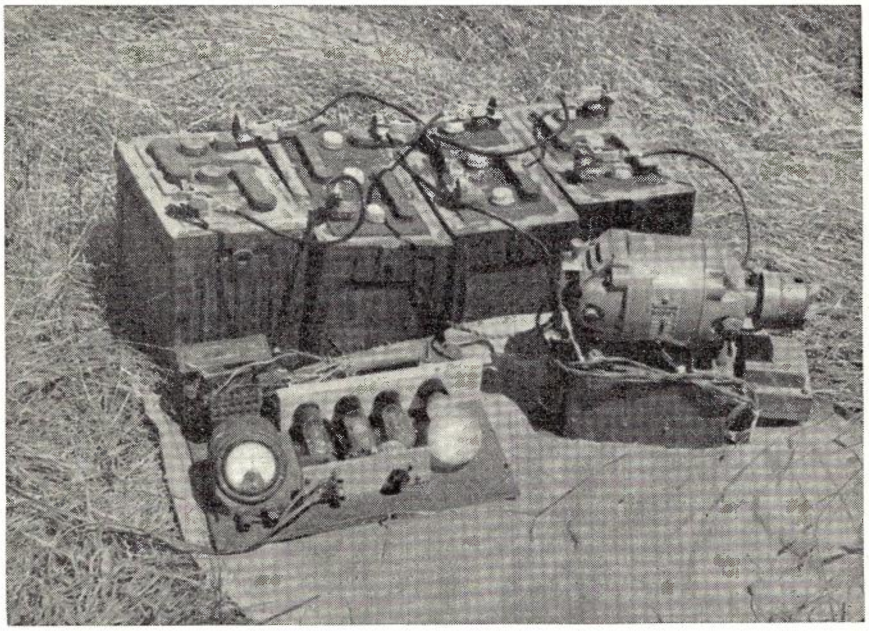

Fig. 6.—Power supply—Batteries, Generator, Transformer, Current Stabilizing and Feeding Circuit and Ammeter.

Control of Excavation

To control excavation with the aid of the locator, the coil system was mounted on the grader, and the locator operator with the detector sat beside the grader driver and advised him constantly of the distance by which the grader blade was clear of the cable (Fig. 13). Where the cable weaved from side to side, this was indicated so that the trench could be altered accordingly. Thus when the mechanical excavation was complete, the depth of earth covering the cable was reasonably uniform and the cable could be easily broken free by hand methods (Fig. 16).

Development of Locating Techniques, and Difficulties Experienced

When the locator equipment was delivered and tested, it was found that the accuracy achieved in the laboratory was difficult to reproduce in the field. Extended trials were therefore made to perfect the equipment and method of operation. Originally a petrol driven alternator was used for the power supply but despite elaborate equipment, sufficient stability of current was not possible. A

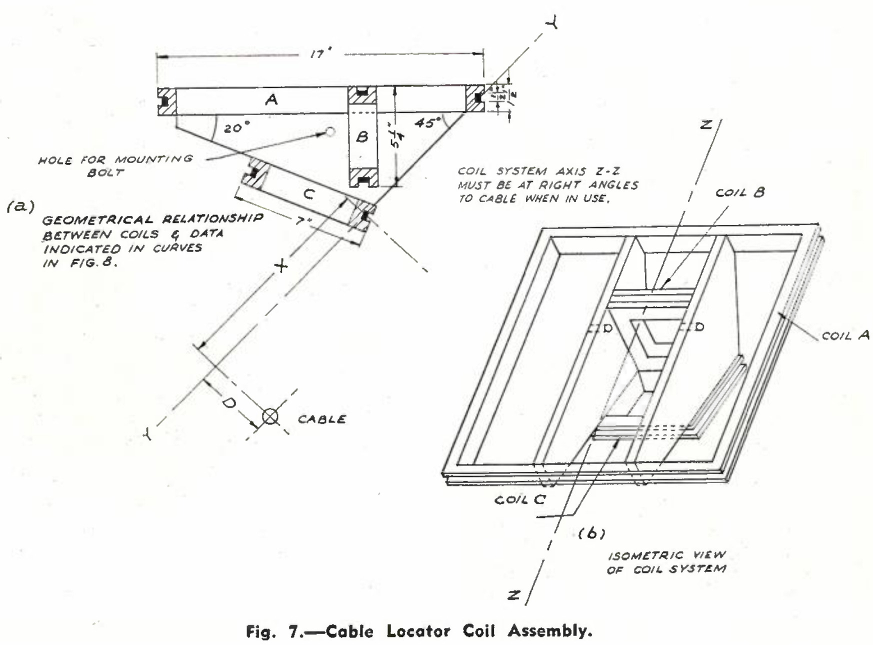

Fig. 7.—Cable Locator Coil Assembly.

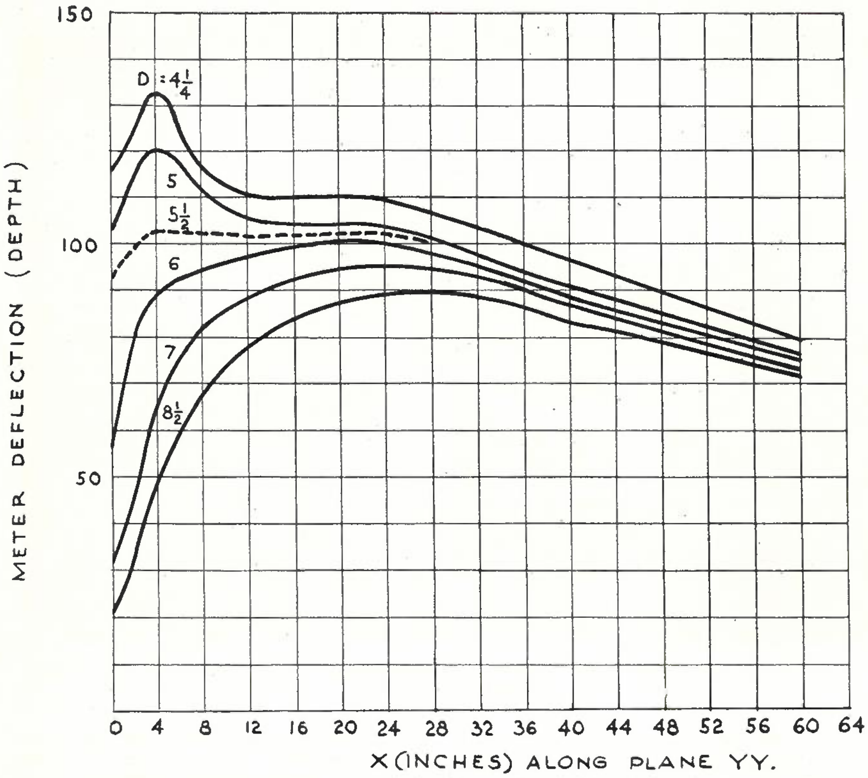

Fig. 8.—Relationship between meter indication and variables "D" and "X" (Fig. 7).

vibrator delivering 50 cycles/sec. was then obtained, but the output of this combined with current of the same frequency induced in the cable sheath by power lines paralleling the cable produced beat effects in the detector which made the meter readings useless. Finally, the genemotor was connected, and by setting the speed to give 60 cycles/sec. the beat of 10 cycles/sec. was high enough to be damped out by the meter movement without the need for elaborate stabilisation. Furthermore, the current output was sufficiently stable.

Some difficulties too were experienced in developing the locating technique. In the initial stages of the tests the ground was dry due to almost drought conditions and the locator appeared to operate quite well and hold calibration. However, after some rain had fallen, the calibrations were found to be different. The conclusion was reached that the loss of calibration was caused by variation in the shielding effect of the ground with moisture content. Thereafter the calibrations were checked frequently from day to day and from section to section along the route as the conductivity of the soil appeared to vary also. These difficulties affected the depth indication but not the location.

Variation in the calibration of the equipment due to the voltage decrease with battery discharge caused some early troubles, but these were overcome by providing a "Calibrate" switch on the instrument and a means of adjusting the voltage of the high tension supply to the amplifier circuit. This provided a sufficiently accurate means of compensating for discharge of both the high tension and filament supplies.

Fig. 14 illustrates the method of feeding current to the cable for either survey or excavation. A small difference between the located and actual position of the cable was found to vary along the length of the cable, increasing sharply as the points of connection of the return leads were approached. This variation was due to interference from the magnetic field in the return lead, both in the length parallel to the cable and in the sections at right angles to the cable. The variation of error was practically eliminated by strictly ensuring that the return lead was parallel to the cable and that the connecting sections of the return were, in fact, at right angles to the cable. Error was totally eliminated by arranging symmetrical return leads on either side of the cable as shown in Fig. 14, although location was sufficiently accurate without this elaboration which entailed considerable additional work in laying out, particularly as crossing a busy highway was involved.

When the locator was attached to the grader to guide digging, there was no need for such care in laying the return

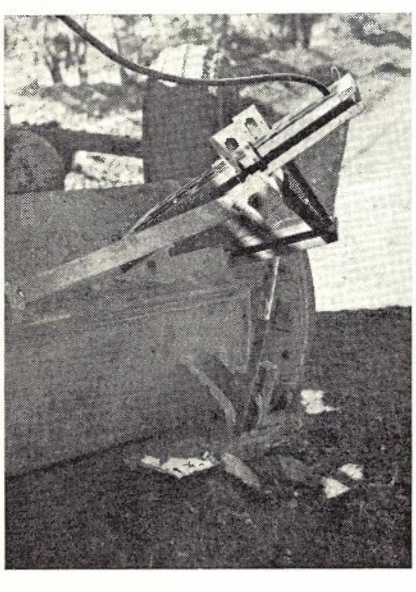

Fig. 9.—Coil System mounted on Grader Blade.

Fig. 10.—Grader Excavating beside Cable, under control of the Locator.

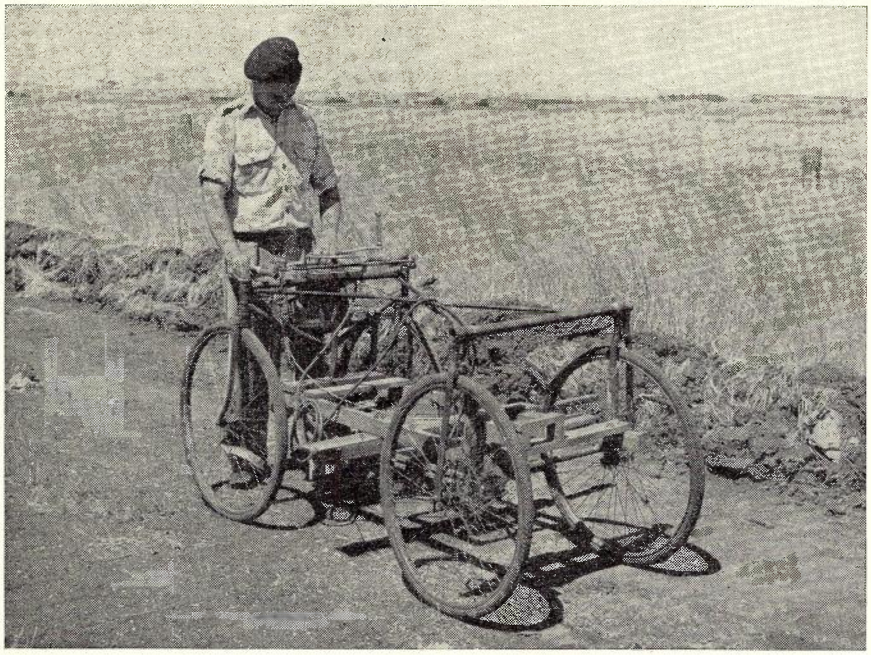

Fig. 11.—Preliminary Survey of Route of Cable.

lead. With the grader cutting an appreciable distance from the cable, accuracy was not critical and where the cut was close (within eight inches) the effect of the return lead was negligible. The value of the current passing through the cable was, however, at all times critical and close attention to ensuring that it was kept constant was necessary. If the current was allowed to fall below normal, the locator interpreted this as representing a greater distance between the grader blade and the cable than there was in fact, leading to danger of damaging the cable.

SHIFTING OPERATIONS

Various methods were developed for excavating the cable to cater for the various circumstances encountered. After numerous trials the method of grading a trench down to below the depth of the cable, approaching it from one side, was developed (Fig. 15(c)). This method proved comparatively safe and the grader was fairly easy to control; furthermore, it conformed with the best practicable technique of using the locator to guide excavating. The method was, however, only applicable where there was adequate space for the grader to operate and the ground was free from rock. There was, of course, always the chance of hitting a stone, but as the cable had been laid in the trench as explained, there were not very many stones or rocks encountered close to the cable in the sections where this method was used. The main disadvantage of excavating the cable in this way was that it was necessary to commence the graded trench about 15 ft. to one side of the cable in order to keep the slope of the trench towards the cable within reasonable limits. If this slope were too great, control of the grader became difficult and if the blade struck a stone or was otherwise deflected it would dig into the side of the trench and possibly damage the cable. Another disadvantage of digging beside the cable, is that the ground was disturbed over a large area. Even though especially consolidated it was not considered as stable or of such high bearing strength for road purposes as if it had been left undisturbed.

With the aid of the cable locator it was possible to dig to within 2 in. of the cable with reasonable safety. However, to guard against accidents caused by occasional rocks in the bottom of the trench, the usual distance was 5 in. after which the cable could be broken from the side of the trench with hand tools. (Fig. 16). Offsetting the cable survey pegs 9 inches and excavating to the side of the cable allowed the pegs to remain intact and indicate the route of the cable for the entire excavating operation.

Despite the considerable amount of mechanical excavation, appreciable hand work was also necessary, mainly at either end of the excavations, road crossings, deep sections, where rock was encountered, where space was too confined or when conditions were too wet for mechanical plant. The use of hand picks was discontinued very early in the operation when considerable damage was inflicted on the cable. No matter how hard the men tried to avoid hitting the cable some

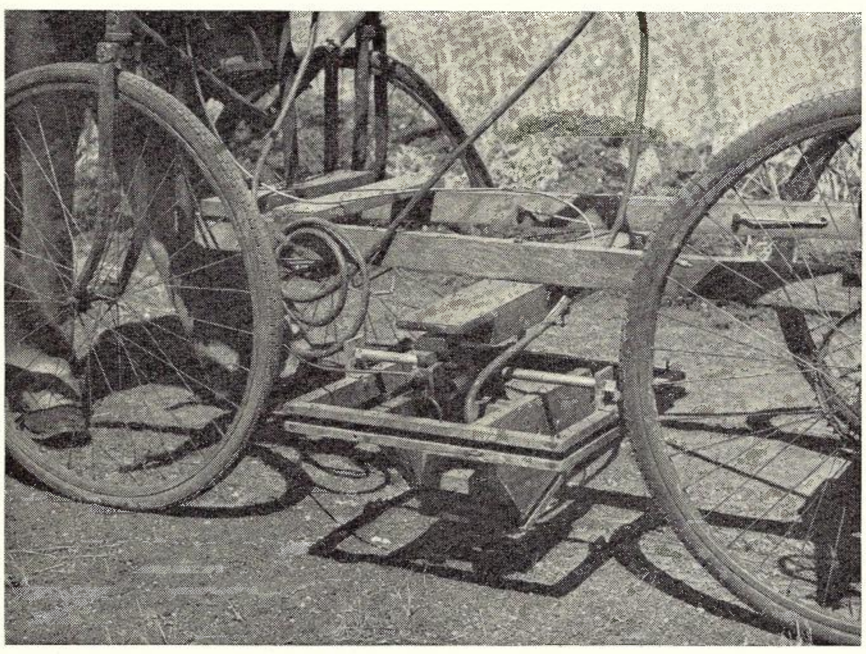

Fig. 12.—Coil Mounting on Survey Vehicle.

Fig. 13.—Control of the Grader using the Cable Locator.

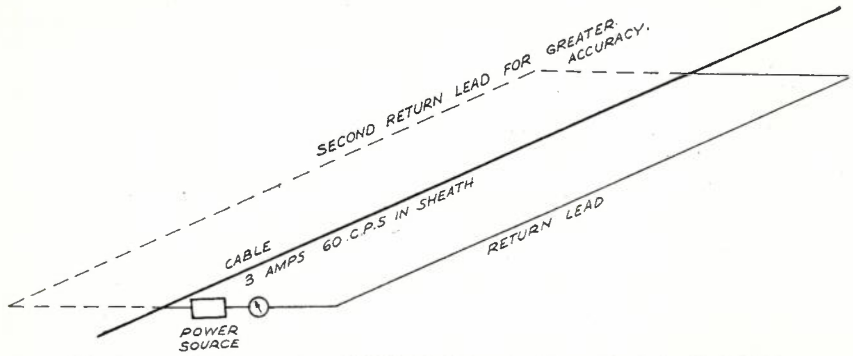

Fig. 14.—Method of feeding current to cable for either survey or excavation.

damage to the sheath seemed inevitable. The use of pneumatic spaders was found to be the most effective alternative as they could be closely controlled and operated safely beside the cable. On each occasion that a length of cable was shifted, it was loosely covered with earth for at least 24 hours before finally covering to the full depth. This enabled the gas pressure alarm system to detect any punctures made in the sheath during the shifting operation. Fortunately these were very few.

Shifting Techniques Developed

Where the cable was to be shifted a distance not exceeding 3 feet, the excavation was made in the manner illustrated in Fig. 15(d). This meant that for the short distance involved there was no need to dig a second trench as the depth of the trench used for excavating the cable was sufficient at that point to accommodate the cable in the new location. If the new location of the cable was within the width of the excavation made for removing the cable, it was sometimes convenient to deepen the cutting at the point where the cable was to be shifted (Fig. 15(e)). This was not always practicable however, due to factors such as obstructions and lack of clearance restricting excavation to only one side of the cable. Where the distance exceeded 8 feet, a second trench was needed along the new alignment (Fig. 15(f)). It was found desirable to excavate the spoil outwards so that there was no chance of the spoil from either trench interfering with the other trench. Figs. 16 and 17 illustrate the cable shifting operations in progress.

For a lateral shift to a depth of up to 4 ft. 6 in. it was possible to use the Barber Greene Runabout Ditcher, which is capable of digging a trench almost 4 feet 6 in. deep and 11 inches wide. The technique was to dig the new trench with the ditcher and excavate the cable using the grader technique. The cable could then be carried over and dropped into the new trench and filled by the grader (Fig. 15(h)).

In two sections the only requirement was to lower the cable where the surface level was to be altered. This proved to be one of the more difficult tasks as it is not possible to excavate beneath a cable unless this is done from the side of an adjacent trench. However, by digging a trench to the maximum depth of a runabout ditcher and keeping about 4 in. from alignment of the cable, it was possible to break the cable from the wall of the trench and lower it to the required depth over a lateral shift of not more than 6 in.., which was acceptable. By digging the trench deeper than required, the earth which fell into the bottom of the trench as a result of unearthing the cable did not have to be removed by hand, and this considerably shortened the job. (Fig. 15(g)).

For Section C, (Fig. 4) the road works involved removing the top of a hill in which the cable was located. As the cost of relocation of the cable to a depth of 16 feet would have been prohibitive, it was necessary to wait until the contractor had removed the 16 foot cut to within 4 feet of the cable. The cable was then excavated and placed in a new trench at the correct level after which the contractor was able to complete his excavation. The relocation of the cable in the creek (Section C) was dealt with by the usual technique employed when laying submarine cables beneath a creek or river. Firstly the trench was dug in the new position with a bucket dredge and

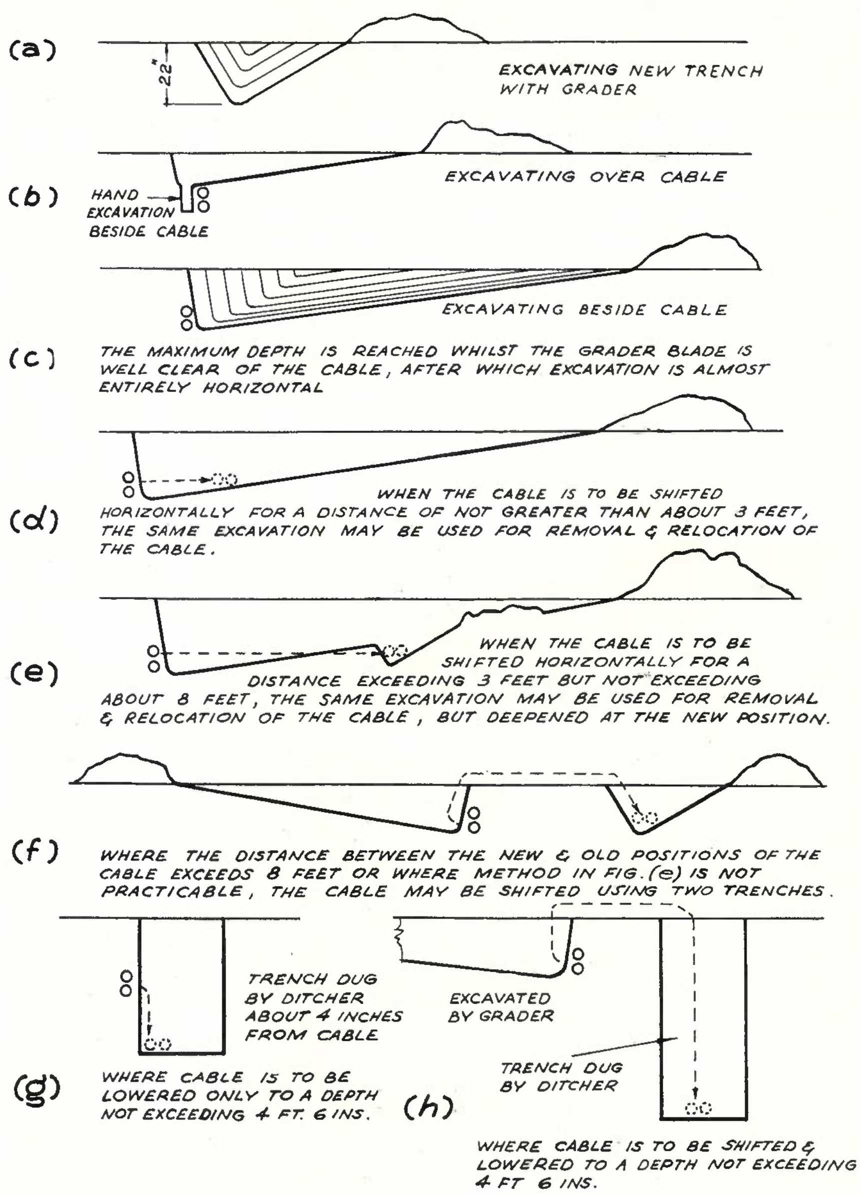

Fig. 15.—Techniques used for excavating for the various cable shifting operations.

Fig. 16.—Removing Cable from the side of the Trench following mechanical excavation by Grader.

afterwards the cable was removed from its existing position and transferred. This operation was one of two that caused faults which will be described later.

Section F involved shifting the cable a short distance laterally and locating at a depth of 6 feet below ground surface level. The use of a tractor mounted back-acting excavator was contemplated and would have been successful had a considerable amount of rock not been encountered. In this case the use of pneumatic equipment was required to complete the excavation. This operation was further complicated by the need to lower V.F. loading coils which were located in a manhole about halfway along the section. Considerable work was required for this operation which involved—

1. Digging out the rocks with pneumatic equipment.
2. Digging a large hole for the new manhole 4 ft. 6 ins. below the existing manhole.
3. Protecting the loading pots and joints and breaking the old manhole.
4. Building the joints and loading pots into a protective and supporting crate.
5. Pouring the base of the new manhole.
6. Lowering the crated loading pots and joints and several chains of cable either side down to the new level.
7. Removal of crating, building of new manhole and filling in trench.

In this section it was also necessary to move the cable clear of two culverts, one a single cell, the other a three cell, which had been built over it. This entailed first supporting the culverts by placing steel beams into each of the cells and taking the weight by means of wooden wedges. After stability of the culverts and the safety of staff had thus been assured the cable was removed.

Throughout the entire project wet weather was a constant hazard as a comparatively small amount of rain could result in difficult working conditions. Whilst unpleasant for the men to work in mud, it was possible, but quite impossible for machinery, as mud in the area concerned is notably slimy and adherent. As far as possible, survey and excavating operations were co-ordinated according to the weather and availability of the grader which was required by the C.R.B. for other work. Although some flooding of the excavations occurred before sections of cable could be shifted, necessitating the use of pumps, the long dry summer which followed enabled the work to move ahead smoothly.

During the early stages of the work when the grader was being used for experiments in digging over the cable, the grader sank to its axles in a section of ground which had appeared quite solid, but proved to have become water-logged. The alarming situation of a 12-ton grader bogged to the axles and resting on the cable thus presented itself rather suddenly. After some experiment the front of the grader was lifted as high as possible by means of its blade and after chocking with wooden blocks, the grader was towed out. In the process of emerging, the grader flattened the armouring of one of the cables but did no other damage.

The condition of the armouring was found to vary considerably along the length of the cable. In some places the armouring and even the exterior jute serving was intact but in other places protection down to the second layer of armouring was corroded away.

Fig. 17.—Placing the Cable in the new trench. (Following operations shown in Fig. 16).

Treatment of Slack

The amount of slack available to permit shifting was very difficult to predict. There were, of course, some sections where the amount of slack available was obvious, e.g., where the cable crossed a road parallel with the cable, but in many sections shifting was required where the cable was laid ostensibly in a straight line. In these cases. the working slack was available from the amount of cable contained in the twists and irregularities imparted during the laying process. This slack which was necessary to enable the cable to be shifted from one alignment to another was usually unpredictable but most sections yielded a yard or so. Surprisingly little slack enabled considerable manoeuvre of the cable. Fortunately there was no difficulty in laying the cable absolutely straight in the new location so that all the slack removed from the irregularities in the first position were available for changing the alignment. In two instances where the cable was changing alignment there was some doubt concerning the amount of slack which would be available. It was therefore necessary to calculate the rate of divergence from the old to the new position to ensure that the change in alignment was possible with the slack probably available. The following example illustrates the principle.

As illustrated in Fig. 18, if change of alignment of ten feet is made via OA, the move will require 24 feet of slack, OB will require 1 foot, whilst via OC, only 6 inches will be needed.

At the end of a shift, there was usually a yard or two of slack. Short lengths were disposed of by "snaking" the cable vertically over a few yards of trench. This method proved simple, cheap and satisfactory as it did not involve stowing the cable outside the alignment of the trench or jointing out lengths. Where long lengths of slack were recovered—e.g., where the cable was lowered 16 feet—the excess was removed by jointing out, which was necessarily more expensive than burying.

USE OF ASBESTOS CEMENT TROUGHING

An innovation was the use of asbestos cement troughing for making a road crossing. This troughing is used extensively for containing power cables by the Victorian Railways Department, where it is known as trunking. It had previously been supplied to the P.M.G. Department by the Railways Department for use in connection with other cable re-location work and proved so suitable that extension of its use for similar work was indicated. Fig. 19 gives details of the cross-section. The troughing was laid on sand packing with concrete setting at butt joints between lengths and also halfway along each length for greater stability. If required, the lids may be sealed by bituminous or similar compound to make the ducts so formed water-tight and resistant to penetration by the roots of plants.

Asbestos cement troughing is suitable for use where cable is to be re-located without jointing—i.e., where it is not possible to pull it into pipe ducts—and possesses the following advantages:—

1. It is easy to handle and lay.
2. It is manufactured to sufficiently close tolerances for use as telephone cable duct.
3. The interior finish is sufficiently smooth.
4. The shape of the duct is ideal for accommodating telephone cables.
5. The material, although being fairly light is also sufficiently strong to withstand without damage the normal rigors occasioned by laying. When laid, the ducts are not readily damaged by men walking on or in them, either with or without the covers.
6. As the troughing is made in 6 foot lengths, it may be laid fairly quickly.
7. The troughing can be made sufficiently stable and may be sealed.

FAULTS

In view of the extent and diversity of the operations, the high degree of freedom from fault during this period is a tribute to the care with which the staff concerned performed their work. On two occasions, however, one of the two cables was rendered inoperative, depriving the route of circuits and affecting telephone traffic for about 4 hours on each occasion. The first of these faults occurred when a section of cable crossing Hovell's Creek (Section C) was being cleared by means of a bucket dredge. This fault was troublesome and tedious to repair as it was located under water. Fortunately, gas pressure in the cable kept water from saturating all but about a foot of the cable in the vicinity of the fault, which proved to be a cut in the sheath about six inches long. The other fault was caused when one of the cables was severed by a Runabout Ditcher during the course of the very exacting work of lowering the cable without lateral shift. Repairs were effected and the operation completed without further incident.

JOINTING AND TESTING TECHNIQUES

Considerable thought was given to the methods of jointing and testing during the various jointing operations to ensure that no circuits were interrupted. As mentioned earlier, the cables had been systematically jointed when they were installed, and for this reason the use of jointing schedules relating channel or circuit numbers to pairs in order of rotation at the various jointing points along the cable was essential so that the pairs could be tested soon after they were jointed. The information was obtained from a combination of cable and channel records, and copies supplied to the testing officers at the terminal exchanges, the jointers, and to technicians at intermediate repeaters where this was indicated. Two-way loud speaking communication sets were connected across one cable pair so that jointers and testing officers could speak to each other at any time.

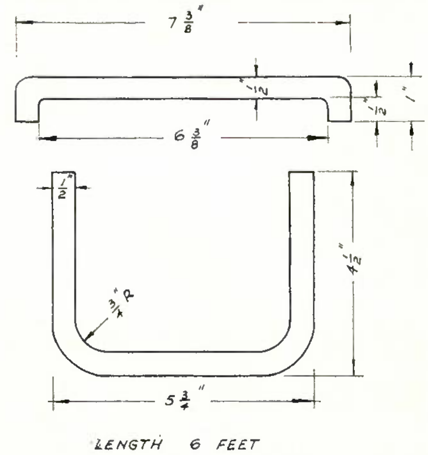

Fig. 19.—Cross-section of asbestos cement troughing.

All jointing operations involved either cutting in or cutting out lengths of cable. In doing this, the jointers maintained the circuits by:—

1. Testing pairs in the new cable for insulation between wires and to ground prior to starting the joint.
2. Testing continuity of each wire in the new lengths of cable (where this applied) before jointing.
3. Applying a D.C. test with non-earthed battery to positively identify the wire at the two points before cutting out or cutting in cable. This could be done without opening or otherwise interrupting the circuit.
4. Jointing through only one wire of a pair at a time, without opening the circuit, using bridging clips or cut-over jointing techniques as appropriate.

By adopting this procedure no mistakes were made in the jointing and no circuits were inadvertently opened.

Before jointing commenced, transmission, insulation, and direct current polarity tests were applied to ascertain whether there were any faults already existing. This testing was considered necessary as earlier experience indicated that appreciable time could be wasted in attempting to rectify faults assumed to have been caused by the current jointing operations, but which were later proved to have existed prior to the start of the work. The tests indicated a small number of minor faults throughout the length of the cable which did not affect service so were listed for attention later. The checks applied whilst jointing was in progress consisted of a listening test when circuits were working followed by a simple transmission loss test when the

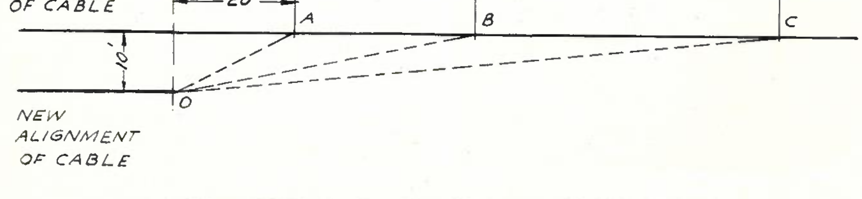

Fig. 18.—Amount of slack required for change of alignment.

circuits became free. Thus a minimum of circuit time was lost during the operation. Before the joints were finally sealed, direct current polarity tests were conducted over the repeater section concerned as a check to ensure that no crosses had been inserted in the conductors.

CONCLUSION

In all, some three miles of working trunk cable were relocated. The work had to be co-ordinated with road making operations involving the Main Road Authority and its contractors. Many unique engineering problems were solved and some new techniques and equipment developed.

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ELECTRICAL NOISE IN AUTOMATIC TELEPHONE EXCHANGES

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INTRODUCTION

Among the factors which determine the intelligibility of speech, one of the most important is the magnitude of the acoustic noise which is the accompaniment of most conversations held under the conditions of everyday life. When a mon-aural communication link, such as a telephone channel, is interposed between the speaker and listener the artificiality thus introduced enhances the effect of noise and the telephone engineer is faced with the problem of avoiding any further degradation of the intelligibility through the introduction of excessive electrical noise into the circuit.

It has been said that if a subscriber is aware of noise on his circuit the irritation experienced will tend to reduce the efficiency of his communication over the circuit. However, long before the noise is of sufficient magnitude to cause irritation, loss of intelligibility due to the noise will have already occurred. The process known as "masking" is too complex to be discussed at length here, but details may be found in the considerable literature devoted to the subject (1).

Electrical noise, for the purposes of analysis and control may be classified as follows:

(a) Microphone noise,
(b) Induced line noise,
(c) Crosstalk
(d) Contact noise,
(e) D.C. supply noise.

The limitation of noise from each of these sources involves widely different measures of control and in some cases, particularly (b) and (d), there is still insufficient knowledge available for more than the simplest measures to be adopted. This article discusses in detail the problem of D.C. supply noise originating within automatic exchanges and describes the various solutions which have been proposed.

THE BASIC LOCAL CALL CIRCUIT

Every call from a subscriber, after it has been established, passes over a circuit which basically may be represented as in Fig. 1. The commonest form of transmission bridge by means of which direct current to actuate the microphone is conveyed to the telephone is the Stone bridge shown. The battery terminals of the bridge are connected to the D.C. supply system which is represented as having a complex impedance Z_s at audio-frequencies. A number of noise sources are connected to the supply, their effect being denoted by a noise e.m.f., E. The double-wound coil is constructed to have a comparatively high impedance which we may denote by Z_c. Hence the noise voltage, V, developed across the line is

$$V = \frac{Z}{Z + Z_c + Z_s} E,$$

where Z is the impedance presented by the line and telephone. In practice, Z is small compared with Z_c, and Z_s very much smaller still, so

$$V = \frac{Z}{Z_c} E.$$

Expressed in decibels with practical values for Z and Z_c of 600 ohms and 9,500 ohms, at 800 c/s respectively, we find that the attenuation offered by the bridge to the supply noise is about 26 db.

Consider now the situation at the telephone. This will be at its worst when the local line has the maximum permissible attenuation and the acoustic noise or room noise is well above average. For planning purposes, the room noise is assumed to be at a sound level of 60 db, which would cover 90% of all situations.

The sound level at a point in a sound-field is the reading in decibels of a recognised sound level meter such as one complying with the American Standards Association specification Z24.3-1944. The scale of db has a zero reference level of 0.0002 dyne per square cm. sound pressure which is approximately the absolute threshold of hearing.

The room noise is transmitted via the microphone and sidetone path of the telephone to the receiver where it tends to mask the incoming signal. It is therefore unnecessary to reduce the supply noise voltage V to such an extent that the corresponding voltage V_r at the receiver is very much less than the voltage due to room noise since the two will act together to mask the received speech.

Equivalent Acoustic Noise Level at the Receiver: The sidetone attenuation of modern telephones (such as the type 300) varies according to the length and type of local transmission line and junction. However, the transmission performance of the circuit falls continuously as the local line increases in

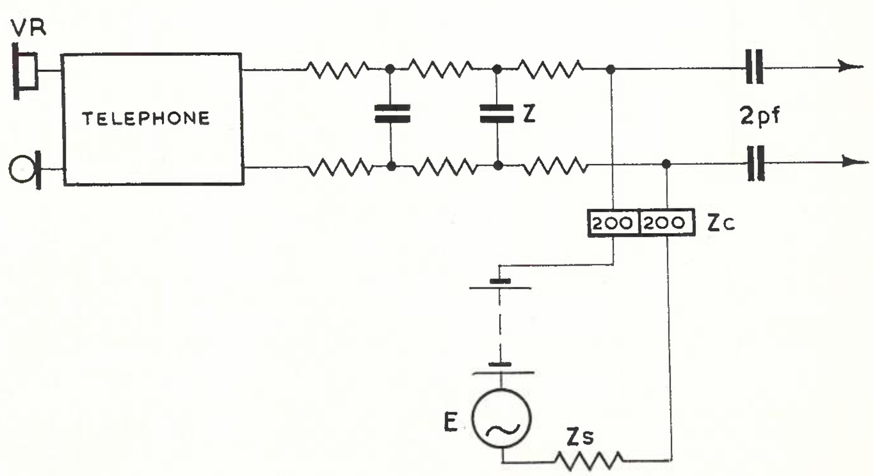


Fig. 1.—Basic Local Call Circuit.

*Mr. Bryant is Divisional Engineer Telephonometry and Acoustics, Research Section at Headquarters.

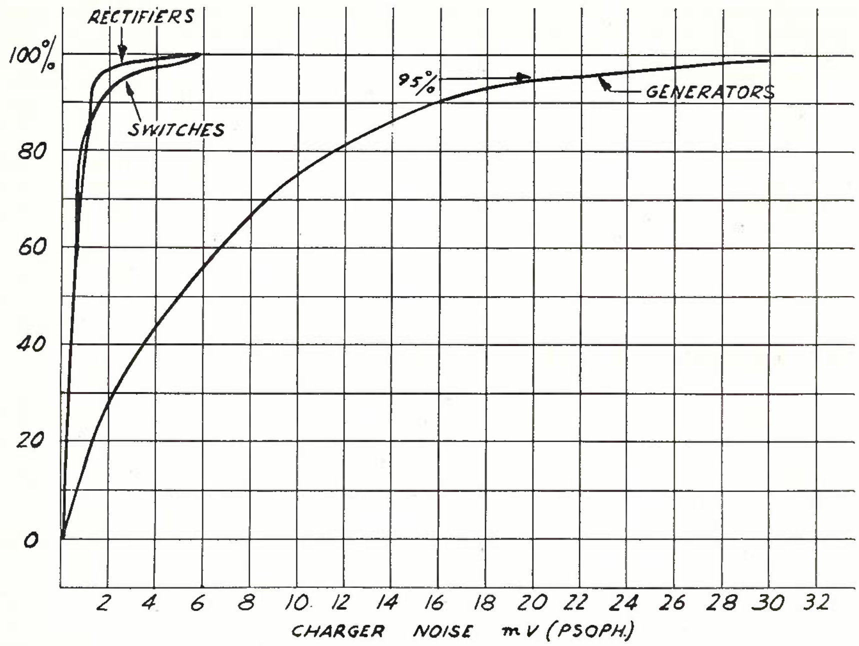


Fig. 2.—Ogives of Switch and Power Noise During a Survey of Melbourne and Suburban Exchanges.

length and for long lines in cable the sidetone attenuation approaches a constant value, affected only slightly by variation in length and type of junction. For planning purposes, the standard grade of transmission performances is provided by two type 300 telephones connected via local lines consisting of 2.84 miles of 10 lb. cable, 200 + 200 ohm, 50V transmission bridges and a junction having an impedance of 600 ohms, non-reactive and an attenuation of 15 db, the telephones operating in 60 db of room noise. In general, connections provide a transmission performance not worse than standard and this circuit is therefore a suitable reference for the determination of permissible electrical noise levels with the possible exception that a more realistic junction might be adopted. Using a network representing a junction containing a mixture of 20 lb. and 10 lb. conductors the psophometric noise voltage at the receiver has been determined for a room noise of 60 db sound level. With average microphones and receivers this voltage is 0.13 millivolt.

It may be argued that this reference voltage should be reduced by 10 db since the average room noise is about 50 db sound level. However, transmission performance provided by the standard circuit is high and would rise only slightly if the room noise were reduced. A factor of greater importance in modifying this voltage is the possible improvement in sidetone reduction of subscriber's telephones. However, at present induction coil and circuit design have barely been able to keep up with that of new receivers having vastly improved sensitivity.

Attenuation of Exchange D.C. Supply Noise: Once again it is convenient to refer to the standard circuit with such modifications as the practical situation requires. The attenuation of a noise voltage measured psophometrically at the battery connections of the transmission bridge and at the receiver terminals amounts to 32 db. With a shorter local line the attenuation falls to 30 db. However, there is an increase in noise voltage across the receiver amounting to as much as 3 db when the circuit between the two subscribers traverses several exchanges due to the addition of noise from these several sources. From these data it may be concluded that, for average calls, the noise voltage at the battery input to the transmission bridge corresponding to 0.13 mV at the receiver would be 27 db above this figure, i.e., 2.9 mV, psophometric.

SOURCES OF NOISE ON THE D.C. SUPPLY

Only those sources of noise which arise in the charging and switching plant are discussed. The charging plant is considered under the separate headings of motor-generator sets and static rectifiers.

Motor-Generator Sets: Generators are, in present practice, normally floated continuously across the operative battery. Such a generator produces a steady charging potential and becomes the virtual source of power for the exchange. In addition, fluctuating noise components are produced which are widely distributed in frequency. Higher frequency components, some of which, arising from the current commutation, are harmonically related to motor speed, fall within the telephone transmission frequency band. The magnitude and nature of motor-generator noise are discussed in relation to the type of machine by Harbottle, et al., (2). In many cases, the large magnitude of generator noise necessitates the use of filter networks in the power supply circuit.

Rectifiers: Rectifiers have superseded motor driven generators since the development of the modern, large capacity selenium rectifier (3). Their use is identical with that of the rotary generator. Chargers of this type produce noise components harmonically related to the power frequency and the psophometric values of these components are usually much lower than is the case with the motor generator units. Comparative values may be obtained from Fig. 2 which relates to Melbourne exchanges. From these curves it may be seen that 95% of rectifier installations produce a noise e.m.f. across the battery of less than 1.8 mV while 74% of motor-generator installations exceed this value. These rectifiers are also equipped with filters and the figures show the extent of the steps now taken to limit charger noise.

Switching Plant: Since a common power supply is used to operate switching equipment and to provide feeding current to the subscriber's telephone, noise transients, arising from the operation of numerous relay contacts, interruptor contacts and selector switches, may pass through the transmission bridge. The electrical circuits contain

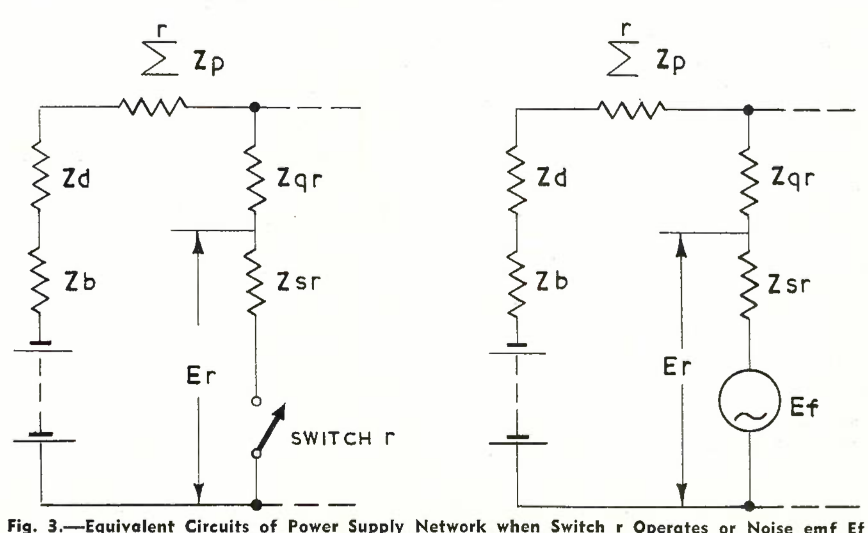


Fig. 3.—Equivalent Circuits of Power Supply Network when Switch r Operates or Noise emf Ef is Generated from Energy Stored in the Circuit.

relay fields, magnet windings, spark-quench capacitors, resistors and contacts which give to the noise a complex and highly variable character.

Noise arising from switch operation is characterised by the occurrence of high amplitude peaks, rather than, as in the case of charger noise a moderately steady level, dependent upon load. In medium-sized exchanges under light load conditions it is possible to record the noise peaks due to the operation of individual switches. For example, the equivalent e.m.f. of a train of pulses caused by a bimotional switch stepping vertically was 0.6 milli-volt psophometric and that resulting from uniselector operation 0.8 millivolt.

In the past, considerable study has been given to the problem of reducing charging plant noise and to the specification of safe limits for this noise. Less attention has been paid to noise arising in the exchange switching equipment. For this reason noise level measurements have always been measured at the battery where the switch noise has already undergone some attenuation. At a point near the switching equipment the situation is rather different and switch noise has been found to be much more important than was previously thought. To take an example, a typical modern exchange had at the equipment room end of the supply busbars the following noise voltages.

Charger noise:—
- rectifier without filter, 7.4mV.
- rectifier with filter, 0.5mV.

Switch noise:—
- peak, 5.8mV.
- average, 1.8mV.

It is apparent that noise from each of these sources must be considered in conjunction with the design of the exchange D.C. distribution system. The ogive of Fig. 2 relating to switch noise indicates the distribution of noise voltages from this source over a large number of exchanges in the Melbourne area. The noise voltage was measured at the battery with the power conversion equipment not operating. It was found that 95% of the exchanges had switch noise at this point in the supply network not greater than 2.2 millivolts. As will be shown later, the switch noise at any point on the supply is dependent on the supply impedance and it is difficult to choose consistently any particular point for all exchanges at which to measure the switch noise. At the time the measurements were made interest lay in determining the magnitude of the charger noise and the switch noise was observed primarily in order to estimate the noise due to the former alone. However, as an indication of the levels at the equipment, the average switch noise at the battery multiplied by the average attenuation of the noise due to the main busbars (17 db) gives 4.2 millivolts for the group of exchanges surveyed; 5% of these exchanges would have noise levels exceeding 15 millivolts.

IMPAIRMENT DUE TO NOISE

The psophometer (4) or circuit noise meter (in America the noise measuring set) has long been adopted as a standard instrument for measuring line noise caused by induction from electric power supply systems. It is less suitable as a device for measuring the fluctuating electric voltages which characterise switch noise. However, because of its universal adoption, ready availability and careful specification the psophometer is widely used for measuring all types of electrical noise.

It is desirable, in order to set limits to the amount of noise transmitted to line from the exchange supply system, to ascertain by subjective methods the relationship between the impairment caused by the noise and its psophometric magnitude. Experiments have been conducted by the C.C.I.F. (now C.C.I.T.T.) to establish this relationship (5). It was found that room noise reduced the transmission performance of the modern type of commercial telephone used in the tests by 6.9 db, i.e., the intelligibility of speech over the telephone when room noise was introduced was equal to that over the telephone without room noise, but with the level attenuated uniformly by 6.9 db. Using rectifier noise added to this room noise a further 3db reduction was obtained by a psophometric voltage twice that produced by the room noise, showing that rectifier noise is less effective than room noise in causing impairment. It has been found that room noise has a spectrum which exhibits a mean rate of fall with increasing frequency of 4.8 db per octave. The spectrum of the ripple on an unfiltered 3-phase rectifier falls at 13 db per actave and the slope is probably increased when a filter is inserted. Thus rectifier noise is spectrally quite unlike room noise. The C.C.I.F. tests showed that a psophometric voltage of 2 millivolts rectifier noise would in conjunction with 0.13 millivolts of room noise cause an additional impairment of 0.5 db.

Switch Noise: In the C.C.I.F. experiments, "white" random noise was used to determine the relative impairment. In this case a 6 db reduction in transmission performance was obtained with an added white noise of psophometric voltage equal to that of the room noise. Above this voltage the impairment rises steeply, but below it the white noise becomes less effective and the impairment is roughly proportional to the root sum of squares of the voltages.

When switch noise is sufficiently intense (i.e. as a result of a large number of switches, etc., operating continuously in the exchange) it is, within the effective bandwidth of the telephone, similar to white noise. The impairment produced by switch noise of low intensity, that is when bursts of noise interrupt the speech at relatively infrequent intervals, has not yet been determined. However, for a noise consisting of impulses of roughly constant amplitude and short duration compared with the mean repe-

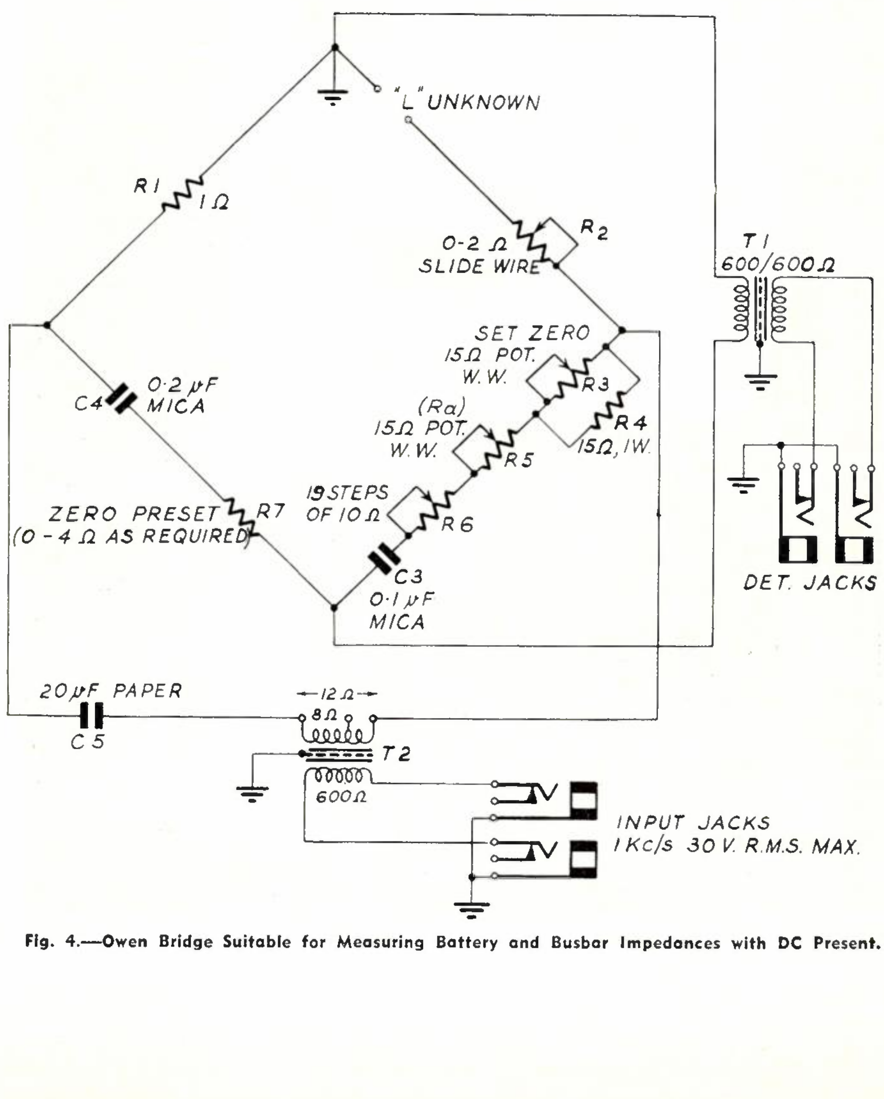


Fig. 4.—Owen Bridge Suitable for Measuring Battery and Busbar Impedances with DC Present.

tition rate it is reasonable to assume that the impairment due to such a noise would be approximately proportional to its psophometric value.

THE POWER SUPPLY NETWORK

The use of a common switch-operating and transmission battery, together with the finite though small power supply impedance, causes the feeding of noise components into the line circuit.

However, exchange busbar layout and charge and discharge arrangements vary considerably so that a simple representation of the total power supply network corresponding to that of Fig. 1 is not possible. Furthermore, batteries vary in size from a few hundred up to several thousand ampere-hour capacity and may be single units or two or more parallel-connected units. Comparison between exchanges using paralleled or single batteries is interesting since the former have a lower battery inductance but may require more complicated busbar systems.

An analysis of the various arrangements of batteries and power supplies existing in exchanges at the present time is given in Appendix I. With regard to the attenuation of noise produced by the power-conversion equipment, the analysis shows that a worthwhile reduction in noise can be obtained by the use of separate charge-discharge leads if the battery is situated some distance from the equipment. While the use of parallel batteries results in lower combined battery impedance, the reduction in the impedance of the whole power supply may be insignificant since the busbars themselves constitute the larger part of the total impedance.

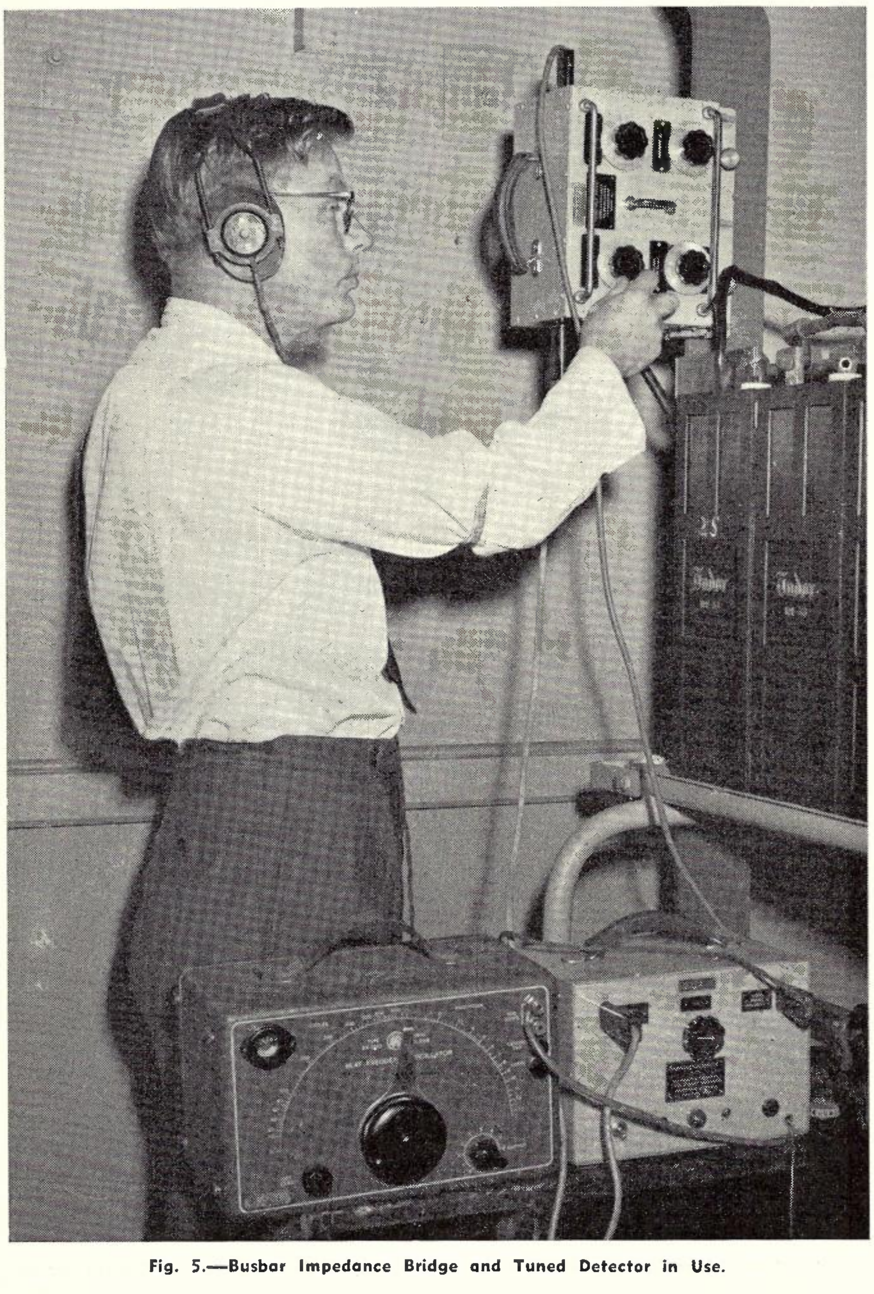


Fig. 5.—Busbar Impedance Bridge and Tuned Detector in Use.

In the consideration of noise generated by the switching equipment two sources may be recognised. One represents the noise potential difference across the impedances of the battery and main busbar due to the fluctuations in current drawn from the power supply by the exchange equipment. The other is the noise e.m.f. which has its origin in the discharge of electrical energy stored in capacitors and iron-cored coils. Owing to the different arrangement when these components are charged and discharged, the action of the spark mechanism at contacts, and the use of spark-quench devices, the nature of this e.m.f. is different from that due to magnetising current fluctuations.

The switch noise due to the operation of a typical switch r may be estimated as follows. In Fig. 3 is shown an equivalent circuit in which Z_b is the battery impedance, Z_d the main busbar and Z_p the incremental impedances along the submain at junctions with the suite distribution busbars represented by Z_q. Suppose a current I_r flows due to the operation of switch r. The change in potential at the switch is:

$$E_r = I_r(Z_b + Z_d + Z_{qr} + \sum_1^r Z_p).$$

At switches (and transmission bridges) more remote from the battery than r there is only slight attenuation of the disturbance created by r. However, there is considerable attenuation between r and the battery. For example, at switch No. 1 we have

$$E_1 = \frac{E_r\ (Z_b + Z_d + Z_{p1})}{Z_b + Z_d + Z_{qr} + \sum_r Z_p},$$

and at the battery,

$$E_b = E_r\ \left(\frac{Z_b}{Z_b + Z_d + Z_{qr} + \sum_r Z_p}\right).$$

If some typical values are inserted in these equations, viz.: $I_r = 1$ amp., $Z_b = 0.03$ ohm, $Z_d = 0.04$ ohm, $Z_{qr} = 0.08$ ohm for $r = 20$, and $Z_{p1} = 0.05$ ohm we obtain the unweighted values $E_r = 0.15$ v, $E_1 = 0.12$v and $E_b = 0.03$v.

Suppose now that the discharge of a capacitor or collapse of magnetic field in a relay or magnet core gives rise to a noise e.m.f. E_f in series with the switch impedance Z_{sr}. Then, since Z_{sr} is much greater than the other impedances shown in Fig. 2 we may write

$$E_r = \frac{E_f}{Z_{sr}}\ (Z_b + Z_d + Z_{qr} + \sum_r Z_p).$$

As in the previous case, the noise voltage at any point is dependent on the busbar impedances and is severely attenuated at the battery.

Impedance of the Supply Network: The elements which together constitute the impedance presented to the terminals of the transmission bridge consist of the conversion equipment (rectifier or D.C. generator), the battery, the main and subsidiary busbars (transmission lines) and switching equipment, the impedance of which will depend on the amount of apparatus actually in operation or busy. The manner in which these elements combine has already been discussed; their relative magnitudes remain to be determined more precisely.

The impedance of batteries is treated comprehensively by Harbottle (2), who gives empirical formulae for the impedance in terms of its component resistance, capacitance and inductance. However, it is found that for most battery installations the capacitive component of the reactance is negligible and the resistance small compared with the inductance for batteries having capacities greater than about 400 ampere-hours. For batteries of 200 ampere-hour capacity the resistance of a 50 V. installation is about 0.02 ohm. A formula for inductance, by Harbottle, is repeated here for convenience:

$$L = 4 \times 10^3 \left[(a + b) \log_e \frac{2ab}{r} - a \log_e (a + d) - b \log_e (b + d)\right] \text{ microhenries.}$$

In this equation, a and b are the lengths of the sides of the rectangle formed by the cells, d is the length of the diagonal, i.e. $\sqrt{(a^2 + b^2)}$, and r is the effective radius of the conductor (r can be taken as 0.02 $AH^{0.7}$, where AH is the capacity of the battery. All lengths are in centimetres.

The inductance of the busbars is discussed in more detail in Appendix II, but reference may be made to the graphs of Figs. 10 and 11 for the inductance per hundred feet run of twin, parallel, rectangular bars at any spacing up to 20 cms. (approximately 8 inches). The graph of Fig. 13 shows the inductance of 1 to 6 pairs of V.I.R. cable having 7/0.064″ stranded conductors and arranged as a compact go-and-return circuit. As mentioned later, cable of this type is a substitute for busbars of conventional form.

The A.C./D.C. conversion equipment may possess an inductance, as seen from the charge busbars, of up to 2 millihenries, for generators and down to a few microhenries for static rectifiers which contain filters in the output. In the latter case, the impedance presented to the busbars consists essentially of a series choke having an inductance of .10 to 100 microhenries and a shunt capacitor having a capacity of the order of 10,000 microfarads (e.g., see Fig. 9 of Ref. 4).

The switching equipment imposes a static load impedance upon the D.C. supply network which has a widely varying effect. In the 2000 type group selector, for example, in the unoperated condition, the vertical and rotary magnets are connected across the supply via the spark quench capacitors. At audio frequencies the shunt impedance is effectively that provided by the two magnets in parallel. A thousand such magnets would present a combined inductance of 170 microhenries. The static load obtained from a number of switches is distributed over several busbars and this distribution may be extremely complex. The resistive component of the busbar and static load impedance is negligible compared with the reactive component and therefore only the latter need be considered in impedance calculations involving all elements of the supply network.

Measurement of Impedance: The inductance at any point in the supply network is usually measured with the exchange operating. The measuring technique must therefore allow for the presence of the 50 volts D.C. potential and, since the inductances are extremely small, it must provide adequate precision of measurement in the range 1 to 30 microhenries. An Owen bridge of the form shown in Fig 4 is suitable. Since a large number of measurements had to be made during the course of an investigation into supply impedances this bridge was made up as shown in Fig. 5. The bridge is arranged to clamp directly onto the earthed (or low potential) bus and a strap-connection is clamped on to the other busbar. Measurements can be made with a D.C. potential between the busbars of up to 300 volts. The unknown inductance (from 0 to 41 microhenries) can be read directly on two calibrated dials; one in 19 steps of 2 microhenries and the other a continuously variable dial calibrated at 0.1 microhenry intervals over a range of 0 to 3 microhenries. The magnitude of the series resistance of the unknown impedance can be obtained from the resistance balance dial which is calibrated in 0.1 ohm steps over a range of 20 ohms. However, as indicated above, the resistance to be measured in busbar impedance measurements consist of a few milliohms for the battery to which the series resistance of the busbars adds only a small percentage. The bridge in its present form is therefore not suitable for the measurement of resistance although only slight modification would be necessary to incorporate this feature.

The tuned detector used with the bridge is also shown in Fig. 5. The detector is necessary owing to the low output voltage from the bridge near balance, to the design of the bridge which has a bridge arm ratio of 800 to 1 and to the high noise level present during impedance measurements on busbars. A gain of 67 db at 1 kc/s is therefore provided by the detector together with a filter of the parallel T type in the negative feedback loop, giving a rejection greater than 10 db at 200 c/s either side of 1 kc/s. The detector is used with a pair of headphones while the bridge is fed with 1 or 2 watts from the oscillator.

SURVEY OF SUPPLY IMPEDANCES IN MELBOURNE EXCHANGES

After analysis of the noise measurements mentioned above had shown that switch noise in many exchanges was of an unsuspected magnitude, it was decided to investigate the supply impedance of as many exchanges as possible in order to determine how reduction of the impedance might be effected and to what extent this would influence the switch noise level.

In the following table a condensed list of the results obtained from this survey is given. The symbols heading the columns denote the following inductances:

L_b, the internal inductance of the battery, and

L_d, the inductance of the discharge bus between battery and power board.

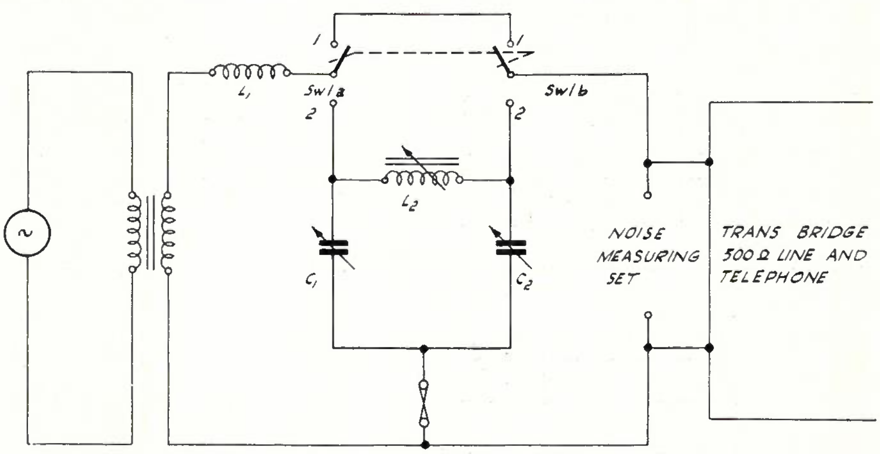
In some cases this constitutes a common charge/discharge path.

L_p, the inductance of the main bus, i.e., between power board and distribution point in the equipment room.

$L_b + L_d$, the common inductance between the charger and main bus for exchanges with common charge/discharge busbars.

$L_b + L_c$, the inductance of the charging circuit for exchanges with independent charge leads, where L_c is the inductance of the charge leads.

$L_p + L_d$, the total inductance of the main busbars between the battery and the distribution point.



L_1 = 20·9 μH (BATTERY AND BUS-BAR IMPEDANCE)

Fig. 6.—Rack Supply Filter Schematic Circuit.

The impedance in ohms at 800 c/s may be obtained by multiplying the inductance in microhenries by 0.005. All inductances in the table are in microhenries. Averages are given below each column.

Type IA Exchanges	L_b		L_b and L_d		L_b
	Batt 1	Batt 2	Batt 1	Batt 2	
	2.9	2.5	10.1	12.7	9.9
	5.1	4.6	13.8	14.3	4.7
	3.8	3.6	17.0	—	2.0
	6.0	—	24.0	—	2.0
	4.2	3.6	16.2	13.5	4.6

Type II A Exchanges	L_b		L_b and L_d		L_p
	Batt 1	Batt 2	Batt 1	Batt 2	
	2.9	—	12.5	—	13.0
	4.0	2.7	9.6	10.5	9.7
	4.4	3.8	13.1	22.0	8.3
	6.1	7.6	12.8	21.8	17.2
	3.3	3.4	13.0	10.0	13.0
	7.4	9.7	11.8	17.3	11.7
	4.3	4.1	12.4	9.2	14.0
	4.6	5.2	12.2	15.1	12.4

Type III A Exchanges	L_b and L_d		L_b		L_p
	Batt 1	Batt 2	Batt 1	Batt 2	
	3.3	3.9	—	9.4	18.0
	3.5	2.8	12.8	7.8	7.6
	7.6	5.9	15.8	12.4	—
	4.8	4.2	14.3	9.9	12.8

Type III B Exchanges	L_b		L_b and L_d		L_p
	Batt 1	Batt 2	Batt 1	Batt 2	
	6.0	5.5	16.6	15.4	18.9
	6.0	5.0	9.0	9.5	15.2
	4.8	4.6	13.0	17.4	19.2
	3.8	—	7.8	—	11.8
	5.2	5.0	11.6	14.1	16.3

Type III C Exchanges	L_b		L_b and L_c		L_p and L_d
	Batt 1	Batt 2	Batt 1	Batt 2	
	1.9	2.3	9.8	15.2	15.2

Type III D Exchanges	L_b		L_b and L_d	
	Batt 1	Batt 2	Batt 1	Batt 2
	3.7	3.9	7.5	7.5

In this table the various types of exchange layout are denoted as follows:

I A. Exchanges with batteries less than 400 ampere-hour capacity and common charge/discharge leads.

II A. Exchanges with batteries 400-1,000 ampere-hour capacity and common charge/discharge leads.

III A-D. Exchanges with batteries 1,000-3,000 ampere-hour capacity and
A, common charge/discharge leads.
B, separate charge/discharge leads.
C, separate charge/discharge leads, parallel batteries.
D, common charge/discharge leads, parallel batteries.

The separate values for each battery when parallel batteries are used are indicative of the effect of the additional busbar required to connect them and the presence of the mutual inductance between charge and discharge paths.

The data show how battery inductance increases with size of battery and how paralleling batteries reduces the total battery inductance, without, however, affecting the total discharge path impedance appreciably. The larger installations also have much more complex busbar arrangements which on the average cause increased inductance to be found between power board and distribution point.

SUMMARY OF EFFECTS OF BUSBAR LAYOUT ON SUPPLY NOISE

It has been established that the relative noise voltages throughout the power supply network are governed by the impedances, consisting essentially of inductive reactance, of which the network is composed. The following considerations are of primary importance in the design of the power supply network.

(a) The charger noise is proportional to the common coupling impedance between the charger and the exchange loops. This impedance consists of the battery impedance plus that of any common charge/discharge leads.

(b) Both the charger and the exchange have comparatively high impedances compared with the common coupling impedance. The series impedances of the charge and exchange loops do not, therefore, attenuate the charger noise significantly.

(c) The switch noise at a point on the main or exchange busbars is proportional to the impedance of the power supply network at that point.

(d) The switch noise is attenuated, before reaching the battery, by the series impedance of the exchange loop.

From these criteria it is apparent that both charger and switch noise may be minimised by reducing the common coupling impedance to a minimum. Common charge-discharge leads should be avoided and the battery arranged for minimum impedance. The series impedances of the main and exchange busbars should always be made as small as possible by minimising their length and spacing. The battery room should be sited as close as possible to the equipment room.

The impedance of the battery may be reduced by the operation in parallel of batteries of smaller individual capacity. However, the increase in area for maintenance of the additional cells is an accompanying disadvantage. If each parallel battery is to be switched, the additional busbar required reduces the advantages offered by paralleling. It is probable that real benefit (from the point of view of noise) can only be obtained by the use of paralleled batteries if separate charge leads are used, when the elimination or reduction in size of filters is possible, and main and exchange busbars are of low impedance, in which case a reduction in switch noise results.

LOW INDUCTANCE BUSBAR CONSTRUCTION

During the period when the investigations described above were being carried out, considerable interest in the reduction of busbar impedance was shown by State Engineering Branches. As a result several exchanges, in New South Wales, Queensland and South Australia, were constructed using various measures which directly or indirectly had the effect of reducing the supply impedance. These included close-spaced busbars, wide busbars, multiple cables, multiple batteries, and batteries installed in the equipment room.

A detailed examination of eight such

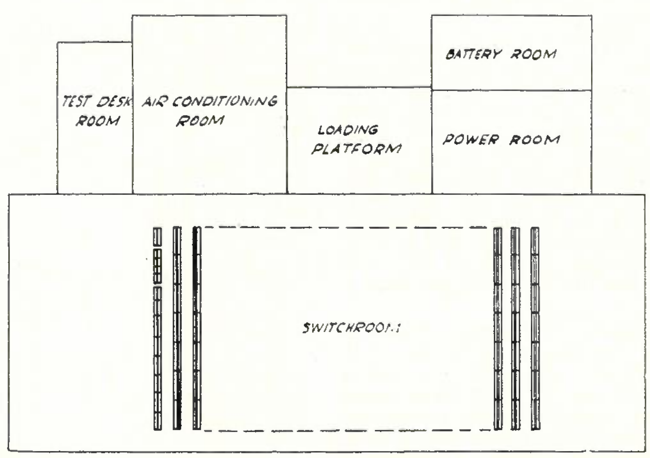


Fig. 7.—Layout of an Exchange with Ultimate Capacity for 5000 Lines.

installations was carried out in an attempt to estimate the effectiveness of the various measures employed.

Willoughby Exchange N.S.W.: In this exchange the busbars in the main discharge circuit are arranged in close-spaced pairs in order to reduce their impedance. Insulating strips having cross-sectional dimensions 2½ x 1 inch type are used between these bars in a "sandwich" type of construction. This feature, together with the use of four batteries—connected in parallel at the power board—results in a very low value of coupling impedance. A comparison between the measured and calculated impedances and also the impedance which would be obtained with bars at standard spacing is given in the following table for various points in the exchange.

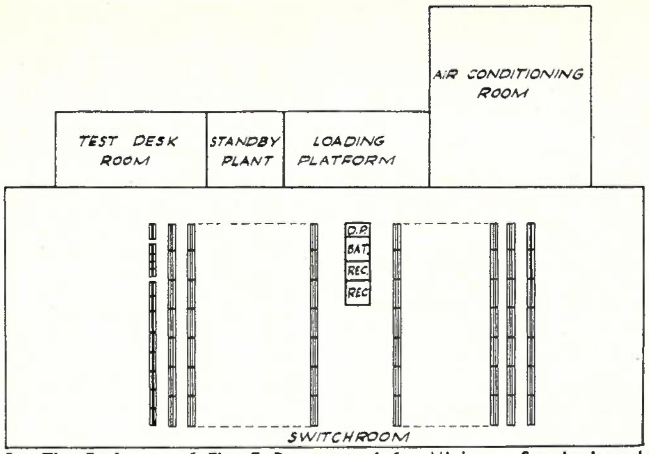


Fig. 8.—The Exchange of Fig. 7 Re-arranged for Minimum Supply Impedance.

Busbar length or point at which measurement was made.	Inductance (microhenries)		
	Calculated	Measured	Standard Spacing
Battery No. 1, disconnected from load.	—	2.8	—
At beginning of main discharge bus from power board (2 batteries in parallel).	—	2.6	6
Inductance of discharge bus from Battery No. 1 to power board (12′ of close-spaced bars).	1.7	1.4	4.9
Inductance of discharge bus from Battery No. 3 to power board (17′ of close-spaced bars).	2.0	—	5.5
Total common coupling impedance, at power board. 4 batteries in parallel.	1.1	1.6	2.0
Impedance on sub-main (standard spaced) at furthest point in the exchange.	2.4	2.2	3.5
Impedance on rack 3A at furthest point in the exchange.	13.6	13.2	14.7

The close-spaced construction extends only from the batteries to the submain distribution to the suites. The main busbar has a run of 22 feet from the power board. The figures show how rapidly the impedance builds up owing to the comparatively wide spacing of submain, suite and rack busbars.

The reduction in coupling indicated in the table is not as great as might be expected with the use of such close spacing of the busbars. It is unlikely that the noise reduction obtained by the use of such construction would exceed 6 db compared with standard construction. However, this is largely because in this particular installation the battery self-inductance is the dominant factor in the impedance of each of the four battery circuits. The use of close-spaced bars would be very desirable if only a single pair were used to feed the four batteries (commoned in the battery room), or if a long discharge path could not be avoided.

Noise levels throughout the exchange were quite low. Even with the main filter capacitor removed from the rectifier the power noise was only 0.1 mV. psophometric.

Chermside Exchange, Queensland: Relatively short lengths of busbar are used in this exchange to connect the two batteries to the power board. Although the bars are apparently close-spaced (½ inch) there are substantial loops in which the positive and negative branches carrying the same current are actually in different parts of the circuit. In this case the rule that the circuit should enclose a minimum area is not observed. The main busbar is spaced at ½ inch between bars and the 22 feet of run has a calculated inductance of 1.1 microhenries. The remainder of the distribution consists of a submain (24.5 feet of 2 x ½ inch bars spaced 4 inches apart) and standard suite and rack distribution. This part of the circuit has an inductance of 14.6 microhenries. Measured inductances were according to the following table.

Point at which measurement was made	Inductance (microhenries)
Battery No. 1, idle	2.5
Same, including discharge bus to board	3.6
Batteries No. 1 and 2 parallel at board	1.9
Battery No. 1 on load, measured at end of main busbar	6.0
With Battery No. 1 only, maximum inductance in exchange (Rack III B).	18.2

The noise measurements indicated that the noise levels throughout the exchange were moderate except at a point (Rack III B) at which the inductance of the supply was maximum, where the maximum switch noise reached 1.6 mV. psophometric. The power noise was low and even when the capacitors in the filter were removed reached only 1.2 mV. psophometric.

Chapel Hill Exchange, Queensland: A battery cubicle is situated in the equipment room of this exchange with the rectifying equipment adjacent tc it. There is therefore no contribution by busbar inductance to the coupling impedance provided by the battery. An improvement in layout of the latter could possibly be made with a corresponding reduction in the inductance from the present value of 1.8 microhenries. The total inductance at the powerboard was found to be 5.8 microhenries. The noise levels at this point were correspondingly low (0.1 mV for power noise and 0.7 mV for switch noise).

North Adelaide Exchange, S.A.: In this exchange, four batteries, permanently connected in two pairs in the battery room, may be combined at the discharge board to give a single battery of low impedance. The installation employs close-spacing (¾ inch) separation of 4 x ½ inch bars for the battery discharge and main busbars and standard arrangements for the remainder of the distribution. The length of run of close-spaced busbar is roughly 76 feet. Results of measurements of inductance and noise in this exchange are summarised in table on page 180.

Point at which measurement was made	Inductance (microhenries)	Switch Noice (mV psophometric)
Battery switching panel	2.4	0.9 (peak) 0.4 (average)
End of main busbar	7.7	1.6 (peak) 0.9 (average)
End of suite furthest from supply	17.8	2.3 (peak) 1.3 (average)

As in the case of Willoughby Exchange, the results obtained at North Adelaide lead to the following conclusions:—

(i) close spacing of main and battery discharge busbars produces only slight noise improvement unless the battery inductance is very small or the discharge path is long.

(ii) Close-spacing must be used throughout the discharge circuits (include submain) for adequate attenuation of switch noise. The spacing used in the North Adelaide of the 4 x ½ inch bar gives an inductance per foot only a third of that of standard bars.

Wakefield Exchange, S.A.: This exchange is unique in that it consists of an old and new section and part of the main busbar run is close-spaced with ¾ inch separation between bars and part is multiple cable. Conventional 4 x ½ inch busbars extend, at close-spacing, some 60 feet; the main then continues in two pairs of 2 x ½ inch bars for 43 feet in standard arrangement (4 inch centre-to-centre) and is finally extended into another equipment room in 6 pairs of 19/0.064 stranded VIR cable for a distance of 83 feet, and in one pair of similar cable for a further 52 feet. Two batteries may be connected singly or in parallel to the exchange by a switch at the power board. The separation of the equipment into two rooms causes the static load of the exchange equipment on the supply system to have an appreciable effect on the measured inductance as indicated in the following table:

Condition	Inductance (microhenries)		
	At Battery	At commencement of multiple cables	At further rack in the exchange
Supply system	3.8	12.9	25.1
Supply system plus static exchange impedance (no load)	3.5	8.1	18.5
Same with light load	3.3	7.8	17.8
Same with heavy load	3.0	6.9	14.7

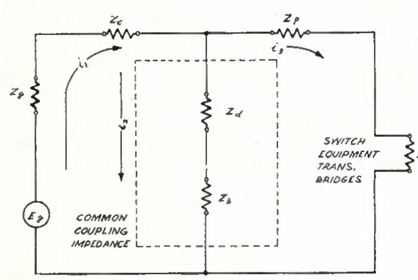


Fig. 9.—Equivalent Circuit of Simple Battery and Common Charge - Discharge Busbar Arrangement.

It will be apparent from this table that under the conditions prevailing in this exchange, i.e., a very long main busbar feeding two equipment rooms, the impedance of the switching equipment operates to reduce the effective impedance of the supply. As the load increases more relays are energised at a given moment and the impedance of the equipment falls, causing the effective impedance of the supply to decrease accordingly.

Corresponding to the high impedances the switch noise levels are also high. The power noise at the discharge panel is 1.1 mV psophometric but less than this figure when the two batteries are operated in parallel. At the furthest rack in the exchange from the power supply, the peak switch noise is 4 mV and the average noise 1.6 mV psophometric. The investigation of this exchange suggests that in circumstances where the length of busbars cannot be reduced, the reduction of switch noise without the use of special filters or a "quiet" battery supply may not be possible.

Conclusions of Survey: The impedance of the battery in the exchanges described in the last section is larger than that of the discharge bus to the switchboard (i.e., not including the main bus). This was not true, in general, of the exchanges previously investigated in the Melbourne survey. Further reduction of the magnitude of the coupling impedance between the power conversion equipment and the exchange would be largely dependent on the reduction of the battery impedance. The impedance of the battery has little effect on the level of switch noise on the rack supply because it is swamped by the impedance of the distribution system. Thus, the magnitude of the battery impedance is important only with respect to the power noise. In most cases the power noise in these exchanges was up to 26 db below the recommended limit (2mV). Power conversion equipment could be designed to work into an impedance of 5 microhenries, with substantial economy in choke and capacitor provision in the filter if the battery charge and discharge bars are designed to have minimum inductance.

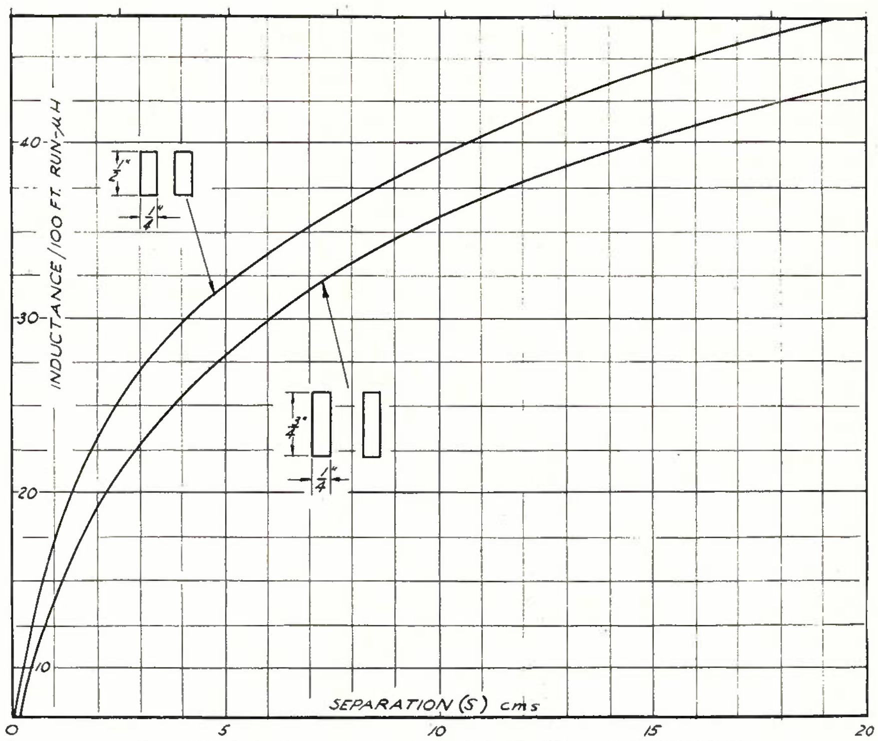


Fig. 10.—Inductance of ½" and ¼" and ¾" and ¼" Parallel Busbars.

If no other method is used to suppress or attenuate switch noise it is necessary that the common impedance of the distribution system should be as low as possible. For practical reasons of construction it is unlikely that the suite and rack distribution design could be radically altered to provide lower inductance. Main and submain busbars can be readily designed to give low inductance compared with that obtained with standard spacing.

DESIGN OF BUSBAR SYSTEMS

In the design of busbar distribution systems, inductance of busbars can be reduced by ensuring that they are as short as possible and are really close-spaced. The Chermside exchange, and other exchanges which have not been described here, contain examples of busbars that are physically close-spaced but do not contain topologically a minimum area since the positive and negative branches are in different parts of the circuit.

In the past, the busbar system has been designed in accordance with the requirements of voltage drop only (Ref. 3, p. 152), but the reduction of inductance to a minimum should be a parallel objective. These objectives are not closely related since the inductance of a pair of parallel conductors is dependent on their linear dimensions and spacing while their resistance is dependent on the cross-sectional area and is independent of the spacing. In both cases there is a linear relationship with the length of run of the conductor-pair so that minimum length of run should be an initial aim in order to keep inductance and resistance small as well as provide for economy in the use of copper. An interesting comparison, from this point of view, may be obtained from the following table which shows the inductance and cost of busbars of different construction and also the amount of space occupied by each:

In this table, the relative cost is that of the copper only; the cost of installation is not included. The total installed cost of multiple cable could well be less than the total cost of a busbar installation as fewer joints would be required and the amount of copper provided could be more closely tailored to meet the needs of the various sections of a particular installation.

Total Permissible Inductance: It has been shown earlier in this article that, on the evidence obtained from experiments conducted by the C.C.I.T.T., switch noise causes greater impairment of transmission performance than power noise. It has been proposed that the maximum switch noise level should not be allowed to exceed 1.4mV. From observation of the inductances and noise voltages in a large number of exchanges it has been found that the noise current (weighted psophometrically) should not exceed 20 mA. The maximum inductance of the power distribution system is determined, from these figures, as follows:

$$Z = \frac{V}{I} = \frac{1.4 \times 10^{-3}}{20 \times 10^{-3}}$$
$$= 0.07 \text{ ohms}$$

Type of Busbar

	2″ x ½″ Std. Spacing (4″)	2″ x ½″ close-spacing (¼″)	45 pairs 7/.064″ VIR cable	17 pair 19/.064″ VIR cable	5 pairs 37/.083″ VIR cable
Inductance (microhenries per 100′)	27	5.7	0.5	1.0	2.5
Total occupied area (sq. in.)	10	2.5	9.0	8.6	8.4
Cost of conductor (rel. to 2″ x ½″ busbar)	1.0	1.0	1.93	1.53	1.67

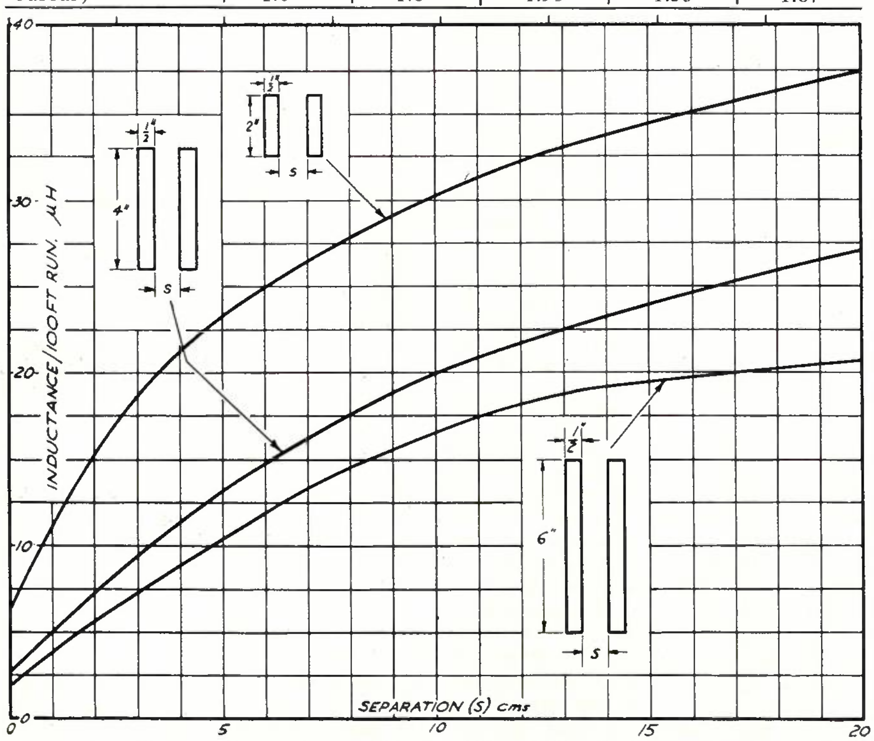


Fig. 11.—Inductance of 2″ x ½″, 4″ x ½″ and 6″ x ½″ Parallel Busbars.

$$L = \frac{Z}{2\pi f} = \frac{0.07}{2 \times 3.14 \times 800}$$
$$= 14 \text{ microhenries}$$

This figure includes the shunt effect of the exchange load, which may be determined from figures previously given. It would appear from experience with the exchanges already incorporating close-spaced busbars that a sufficiently low inductance can be obtained if this form of construction is carried right up to the suites, i.e., from the batteries to the power board and through the main and submain busbars in the equipment room. There is no need, and practical difficulties make undesirable, to employ close-spacing for the busbars branching off the submain which feed each suite of racks (the suite busbars).

RACK SUPPLY FILTERS

American techniques for the reduction of electrical noise in exchanges include the use of "decentralised" filters in the power supply to individual racks of equipment. This method is claimed to have several advantages over the use of a single, large filter in the main power supply feed. The small filters are fitted only where needed, while the central filter has usually to accept the total load current of the exchange and is consequently large and costly. The individual filter may be used to isolate the transmission bridges from the switch noise generated in neighbouring racks of equipment.

The rack supply filter is intended primarily to attenuate switch noise which arises in adjacent racks. However, with such filters, the possibility of removing other filtering from the supply could also be considered, e.g., a reduction in size of the battery or more practically the omission of the filter in the output of the power rectifier. In existing conditions a load of 20 amps would be required to be carried by a filter supplying a rack and incurring a potential drop not exceeding say 0.2 volt. The filter must be of suitable dimensions for mounting in line with, or close to, the distribution busbars at the top of the rack. These conditions impose limits on the size of inductors and capacitors which may be used in such a filter.

A simple half-section filter was constructed, tested and installed on a rack in Camberwell exchange, Victoria, in order to obtain a practical evaluation of the performance of such a filter. Practical values for the elements of the filter were chosen by tests carried out in the circuit shown in Fig. 6. The actual values used in the filter constructed were L_2 = 0.38 millihenry, C_1 = O and C_2 = 4,000 microfarads. The impedance presented by a rack of switches is many times greater than the output (image) impedance of the filter at all frequencies above about 200 c/s. The cut-off frequency for the basic half-section is given by:

$$Z_1 + Z_2 = 0,$$

$$\text{i.e., } j\omega L = \frac{j}{\omega C}.$$

Inserting the numerical values we have

$$j\omega\ 380 \times 10^{-6} = \frac{j}{\omega\ 4000 \times 10^{-6}}$$

$$\text{i.e., } \omega^2 = \frac{10^{12}}{380 \times 4000}$$

$$\text{Hence } f_c = \frac{10^6}{2\pi\sqrt{380 \times 4000}} = 130 \text{ c/s.}$$

It has been stated (6) that American practice is to feed racks of equipment through smoothing units consisting of a 2000 microfarad condenser and a choke of 1.9 or 0.8 millihenries according to whether a current capacity of 10 or 25 amps is required. For the larger unit the cut-off frequency of the half section is 126 c/s. Provided the rack impedance in this case is also much larger than the impedance of the filter it may be assumed that the reduction in noise obtained with the filter installed at Camberwell is similar to that of the American prototype. Under the conditions indicated in Fig. 6 the filter gave an insertion loss of 7 db at 200 c/s increasing uniformly to 28 db at 1 kc/s. Measurements at the exchange showed that the reduction achieved by the filter in noise entering the rack is about 25 db psophometric. When the rectifier is operating with its own filter disconnected the hum content of the total noise entering the rack is the most important element and is not reduced to a satisfactory level by the filter; a further 9 db reduction is then required.

The filter reduces only the switch noise originating outside the rack but there is an apparent reduction of noise caused by a switch operating on the filtered rack compared with the unfiltered condition, since without the filter this noise is added to the noise coming in to the rack from elsewhere. It was estimated that the average noise level reaching the rack at which measurement was being made was equivalent to the continuous operation of eight switches simultaneously.

AN EXAMPLE OF EXCHANGE LAYOUT

The layout of a typical exchange of medium capacity is shown in Fig. 7. This example is taken from an actual exchange recently constructed. The capacity of the exchange is 5000 lines, but it has been equipped initially for 1000 lines. Two batteries are provided together with a 100 amp rectifier to carry the present load.

Calculation of the inductance of the power supply and distribution gives a nominal 32 microhenries including 8.8 microhenries for suite and rack busbar inductance. Inclusion of the shunt inductance (for the 1000 lines) in the calculation reduces the maximum effective inductance to 22.3 microhenries, a reduction of 30%.

The problem of reducing the inductance to an acceptable value (14 microhenries) may be approached in several ways. For example, if close-spaced bars were used, the inductance of the batteries, main and submain busbars could be reduced to 9.7 microhenries which, with the 8.8 microhenries of the suite and rack busbars, gives a total nominal impedance of 18.5 microhenries. The effect of the static load of the exchange would be large enough only to give a maximum effective inductance of 14.6 microhenries. The busbars would be 2 x ½ inch, separated ½ inch, and this close-spacing of the submain busbars would present some problems in providing for the connection of the ¾ x ¼ inch busbars feeding the suites.

A possible alternative would be the adoption of a different layout which would bring the batteries closer to the midpoint of the distribution in the equipment room. In this way the maximum effective inductance might be brought as low as 16 microhenries. The use of battery cubicles and rectifiers in the switch room would perhaps be the most satisfactory method of achieving a maximum effective inductance of less than 14 microhenries. The exchange layout might then appear as in Fig. 8. The battery is mounted centrally in the exchange in order to keep the submain busbars short and also because in the initial stages the exchange is not fully equipped. The equipment is arranged so that switches and relays equipped with transmission bridges are situated as close to the battery as possible.

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APPENDIX I
CIRCUIT ANALYSIS OF D.C. DISTRIBUTION

The simplest form which the power conversion and distribution system can take is that incorporating a single battery and common charge-discharge busbars. The electrical equivalent circuit is that of Fig. 9. In this circuit E_g represents the noise e.m.f. of the conversion equipment which has an internal impedance Z_g. The impedance Z_b of the battery and Z_d of the charge-discharge leads constitute the common coupling impedance between the convertor and the load. Z_c and Z_D are the impedances of the charge lead to the switchboard and the main bus to the load. The impedance of the load, Z_s, consists of the switching equipment. The noise currents due to the e.m.f. E_g are i_1, i_2 and i_3 as shown and, by Kirchhoff's Laws,

$$i_2 + i_3 = i_1,$$

$$i_2 (Z_d + Z_b) = i_3 (Z_D + Z_s),$$

$$\text{and } i_1(Z_g + Z_c) + i_2(Z_d + Z_b) = E_g.$$

$$\text{Hence } i_1 = i_2 \left(1 + \frac{Z_d + Z_b}{Z_D + Z_s}\right).$$

$$\therefore i_2 \left\{ (Z_g + Z_c) \left(\frac{Z_D + Z_s + Z_d + Z_b}{Z_D + Z_s}\right) + Z_d + Z_b \right\} = E_g,$$

whence

$$i_2 = \frac{(Z_D + Z_s)\, E_g}{\left\{ (Z_g + Z_c)(Z_D + Z_s + Z_d + Z_b) + (Z_d + Z_b)(Z_D + Z_s) \right\}}.$$

Here Z_b, Z_c, Z_d and Z_D are very much smaller than the impedances Z_g and Z_s, so we may write

$$i_2 = \frac{Z_s}{Z_s\, Z_g},\ E_g = \frac{E_g}{Z_g}.$$

The noise p.d. across the battery is

$$E_b = \frac{Z_b}{Z_g} \cdot E_g,$$

$$\text{i.e. } \frac{E_b}{E_g} = \frac{Z_b}{Z_g}.$$

Similarly the noise p.d. across the battery and charge/discharge leads is

$$E_{bd} = \frac{(Z_b + Z_d)}{Z_g} E_g,$$

$$\text{i.e., } \frac{E_{bd}}{E_g} = \frac{Z_b + Z_d}{Z_g}.$$

The noise voltage, E_{bd}, is effectively that appearing across the impedance Z_s offered by the switching equipment since this is large compared with Z_b, Z_d or Z_D.

When separate charge-discharge leads are used the common coupling impedance is reduced to that of the battery above. There may, however, be mutual inductance between portions of the charge and discharge bars. If this can be neglected, i.e., if the bars are spaced more than a few inches a calculation similar to that made above shows that

$$\frac{E_b}{E_g} = \frac{Z_b}{Z_g}$$

and E_b is the noise p.d., due to the power converter, existing at the switching equipment. The maximum reduction in power noise which can be obtained by using a separate charge-discharge system is therefore $20 \log \frac{Z_b + Z_d}{Z_b}$ db compared with the common charge-discharge system. Batteries are often used in pairs and may be connected permanently together in the battery room or by a switch at the switchboard. Common or separate charge-discharge leads may be used.

For two or more batteries connected in parallel at the battery end of a common charge-discharge bus the impedance Z_b of the simple case discussed initially is made up of the impedance of the separate batteries and interconnecting leads. For example, if there are two batteries of impedance Z_{b1}, Z_{b2} and an interconnecting bus of impedance Z_{bb} we have

$$\frac{E_{bd}}{E_g} = \frac{Z_b + Z_d}{Z_g} =$$

$$\frac{Z_{b1}\,(Z_{b2} + Z_{bb})}{Z_{b1} + Z_{b2} + Z_{bb}} + Z_d \,.$$

If the batteries are identical and Z_b can be neglected then the noise p.d. appearing across the impedance Z_s is:

$$E_s = \frac{\left(\frac{Z_{b1}}{2} + Z_d\right)}{Z_g} E_g \,.$$

If the batteries are to be switched there are separate leads from them to the switchboard. If these leads are used for both charging and discharging the common coupling impedance contains these elements also. As in the previous example, if the batteries have associated leads Z_{d1}, Z_{d2} the coupling impedance is

$$\frac{(Z_{b1} + Z_{d1})\,(Z_{b2} + Z_{d2})}{Z_{b1} + Z_{b2} + Z_{d1} + Z_{d2}}$$

which for two batteries with similar busbar arrangements becomes

$$\tfrac{1}{2}(Z_{b1} + Z_{d1})$$

However, in practice, the coupling impedance is often greatly increased by mutual inductance between the discharge leads.

The last case to be considered is that of several batteries connected in parallel in the battery room and served by separate charge-discharge leads. If there is no mutual impedance the common coupling impedance is given by

$$\frac{Z_{b1}\,(Z_{b2} + Z_{bb})}{Z_{b1} + Z_{b2} + Z_{bb}},$$

for, say two batteries. For two similar batteries connected by a short bus (Z_{bb} negligible) this expression reduces to $\frac{1}{2} Z_{b1}$ and the power noise transmitted to the switching equipment is

$$E_s = \frac{Z_{b1}}{2Z_g} \,.\, E_g.$$

APPENDIX II

INDUCTANCE OF BUSBARS OF ROUND AND RECTANGULAR CROSS-SECTION

General Principles: The electromotive force induced in a circuit A when the current in a neighbouring circuit B is changed is proportional to the rate of change of the linkages of flux set up by the current in B with the conductors of circuit A.

If the circuits are linear, the magnetic induction in A is directly proportional to the current in B causing it, and the electromotive force induced is $-M\frac{dI}{dt}$ where M is the Mutual Inductance of the circuit. With permeability constant and equal to unity M is dependent only on the geometry of the circuit.

A conductor by itself will also have induction due to its own current. The electromotive force induced is $-L\frac{dI}{dt}$,

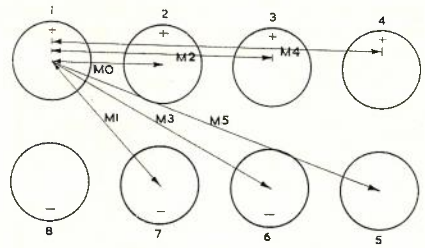

Fig. 12.—An Arrangement of Eight Conductors Symmetrically in Unipolar Layers, Showing the Six Independent Mutual Inductances.

where L is the Self Inductance of the circuit. As before, if the permeability is unity, L is determined by the geometry of the conductor.

L and M are measured in henries in the M.K.S. system.

Geometric Mean Distance: Two conducting filaments of length l, l^1 have a mutual inductance given by

$$M = \int_0^{l^1}\int_0^{l} \frac{\cos\theta}{r}\, ds\, ds^1,$$

where r is the distance between two elements ds, ds^1 in the two filaments, and θ is the angle between their directions. The integration is carried over all possible pairs of elements.

If the filaments are parallel, distant d apart, and of equal length, the integration is readily carried out with a result which may be approximated to

$$M = 2l\,(\log_e \frac{2l}{d} - 1),$$

provided d is small compared with l.

A wire or bar may be considered to be composed of filaments. The inductance of a wire or pair of wires will therefore be of the form

$$N = 2l\,(\log_e \frac{2l}{R} - 1),$$

in which

$$\log_e \frac{2l}{R} = \frac{1}{n}\left(\log_e \frac{2l}{d_1} + \log_e \frac{2l}{d_2} + \,..\, + \log_e \frac{2l}{d_n}\right)$$

where $d_1, d_2, \ldots d_n$, are the distance between all pairs of filaments. We may write

$$\log_e R = \frac{1}{n}(\log_e d_1 + \log_e d_2 + \,..\, + \log_e d_n),$$

and the value of R satisfying this expression is known as the Geometric Mean Distance.

An extension of the concept to a single area allows the expression of a geometric mean distance of an area from itself to be used. This may be written as

$$S \log_e R = \int_s \log_e r.\, da,$$

where r is the radius vector from the point, inside or outside the area, chosen as origin. It will be readily seen that, if there are several areas, A.B.C. ...., with geometric mean distances from another area S of R_A, R_B, R_C, etc., then the geometric mean distance from S of the sum of these areas is

$$\log_e R = \frac{A \log_e R_A + B \log_e R_B + C \log_e R_C + }{(A + B + C \ldots)}$$

Inductance of a Return Circuit of Two Parallel Bars: The self-inductance of a conductor of geometric mean distance r from its axis may be written as

$$L = 2l\,(\log_e \frac{2l}{r} - 1)$$

and the mutual inductance of a pair of conductors having a geometric mean distance R with respect to each other as

$$M = 2l\,(\log_e \frac{2l}{R} - 1)$$

A return circuit consisting of two conductors has a total inductance

$$L_T = 2(L - M)$$

where $L = L_1 = L_2$ are the self inductances of the conductors and M is the mutual inductance.

Inserting the expressions given for L and M into this equation gives

$$L_T = 4l\,(\log_e \frac{2l}{r} - \log_e \frac{2l}{R}).$$
$$= 4l\,(\log_e R - \log_e r)$$
$$= 0.004\,(\log_e R - \log_e r)$$

microhenries per cm if R, r are in cm.

Graphs are given in Figs. 10 and 11 for five common types of busbar.

Inductance of a Return Circuit of Three Parallel Bars: This analysis is given as an example of the procedure that may be followed in analysing complex arrangements of busbars. The circuit consists of two outer bars in parallel and an inner bar which constitutes the return conductor.

Suppose unit current flows into the circuit, then the current in each outer conductor is one-half and the induced e.m.f. in each of the two outer conductors is proportional to

$$\tfrac{1}{2} L - M_1 + \tfrac{1}{2} M_2 \,,$$

where L is the self inductance of either outer conductor, M_1 is the mutual inductance of the inner and outer conductors and M_2 is the mutual inductance of the two outer conductors. Hence this expression represents the inductance of each of the outer conductors which together are in series with the inner conductor, the inductance of which by the same reasoning is

$$L - \frac{1}{2} M_1 - \frac{1}{2} M_1 .$$

The inductance of the complete circuit is therefore

$$L_T = \frac{1}{2} \left(\frac{1}{2}L - M_1 + \frac{1}{2} M_2\right) + (L - M_1) = \frac{1}{4} (5L - 6M_1 + M_2).$$

Note that the numerical coefficients of the inductances, having regard to sign, total zero and L_T is therefore linearly dependent on the length of the conductors as in the two-bar case. The values of L_1 M_1 and M_2 in terms of geometric mean distances are substituted as before. Hence,

$$L_T = \frac{1}{2} (6 \log_e R_1 - \log_e R_2 - 5 \log_e r).$$

Values of $\log_e R_1$, $\log_e R_2$ and $\log_e r$ in abhenries per cm may be calculated by the method given in ref. 7.

Inductance of Multiple Cable: The inductance of two conductors forming part of the same circuit has been shown to be $2L \pm 2M$ where L is the self inductance of each conductor and M is the mutual inductance between the conductors. Each conductor can be thought of as having a total of $L \pm M$ which together with that of its neighbouring conductor produces a total inductance of $2L \pm 2M$.

It is necessary to study the manner in which inductors combine when additional conductors are placed next to the first, e.g., in two layers, first with unipolar, positive and negative layers, and second with alternate positive and negative conductors in each layer.

The inductance of each individual cable is treated separately. The inductance consists of the self inductance L and the mutual inductances M_0, M_1, M_2 etc. with the other cables. For example, suppose there are eight cables arranged as in Fig. 12, in two unipolar layers, numbered 1 to 4 in the top layer and 8 to 5 in the bottom layer. The mutual inductances between cable number 1 and the remaining cables are indicated in the diagram.

All other mutual inductances can be identified with one or another of these because of the symmetry.

The inductance to be assigned to cable 1 is $(L + M_2 + M_4 - M_1 - M_3 - M_5)$; the inductance of cable 2 due to itself and the other cables is $(L + M_0 + M_2 - 2M_1 - M_3)$. Since the positive cables 1 and 4 owing to symmetry have the same inductance, likewise cables 2 and 3, the total parallel inductance, L_D, due to these four, parallel connected cables can be found from:

$$\frac{1}{L_D} = \frac{2}{L + M_2 + M_4 - M_1 - M_3 - M_5} + \frac{2}{L + M_0 + M_2 - 2M_1 - M_3} .$$

As these four, positive, parallel cables are connected in series to the four negative cables of similar inductance the total inductance of the circuit will be $L_T = 2L_D$.

The inductances are evaluated by means of the formulae

$$L = 0.002\, l \left(\log_e \frac{2l}{p} - \frac{3}{4}\right) \mu H,$$

$$M = 0.002\, l \left(\log_e \frac{2l}{d} - 1\right) \mu H,$$

where l is the length of run, d the centre to centre distances between cables, and p the radius of the copper in the cable (all in cms).

It has been found that, for rectangularly arranged cables the minimum inductance is achieved with alternate positive and negative cables in each layer and the maximum occurs when there are layers of alternate polarity, each layer being unipolar. The inductance decreases as the number of pairs is increased, but more and more slowly. These results are summarised in the curves of Fig. 13.

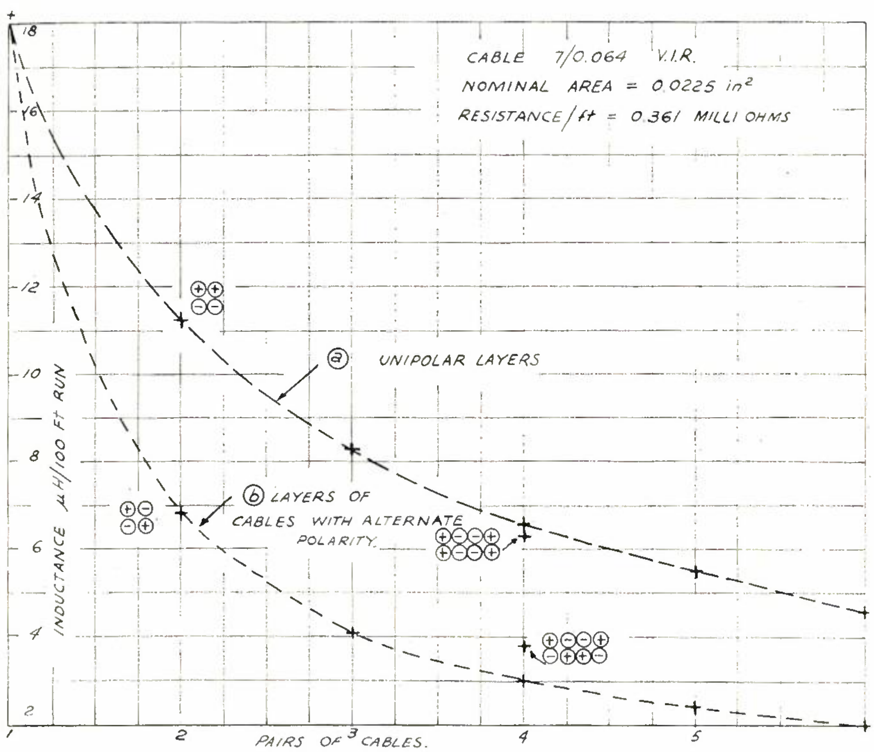

Fig. 13.—Inductance of Multiple Pairs of Cable.

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METHODS OF NUMERICAL FILTER DESIGN—PART III

*E. RUMPELT, Dr.Ing.**

6. DESIGN OF IMAGE ATTENUATION WITH THE HELP OF TEMPLATES

The trial and error process of finding a realisable image attenuation characteristic, which fits a specified minimum attenuation curve, requires a substantial part of the total time necessary for the design of a filter and it is therefore desirable to facilitate this process as much as possible. This can be achieved by a graphical design method which permits the use of templates for drawing the attenuation curves of the individual sections. The necessary preliminary condition for such a method is that the image attenuation function of an m-derived section can be transformed by a suitable frequency transformation into the following general form:

$$\alpha_s = F(\gamma - \gamma_\infty) \quad \ldots\ldots (6.1)$$

Where γ is the new frequency parameter and γ_∞ the transformed attenuation peak frequency.

6.1 DERIVATION OF TEMPLATE FUNCTION FOR LOW-PASS FILTER DESIGN

By introducing the frequency transformation

$$\gamma = \tfrac{1}{2}\ln(1 - 1/\Omega^2) \quad \ldots\ldots (6.2)$$

into Eq. (5.23), that is, by substituting in this equation e^{γ} for $\sqrt{1 - 1/\Omega^2}$ and e^{γ_∞} for $\sqrt{1 - 1/\Omega_\infty^2}$ the function of the image transfer constant of an m-derived low-pass filter section becomes:

$$\Theta_s = \alpha_s + j\beta_s = 2\coth^{-1} e^{(\gamma - \gamma_\infty)} \quad (6.3)$$

According to Eq. (6.2) γ is real for $\Omega > 1$, that is, in the stop-band of the low-pass filter section. Here the image phase-shift is either 0 or π.

In the first case:

$$\alpha_s = 2\coth^{-1} e^{(\gamma - \gamma_\infty)} \quad (\text{for } \gamma > \gamma_\infty) \quad \ldots\ldots\ldots (6.4)$$

In the second case, because of $\coth \frac{1}{2}(\alpha_s + j\pi) = \tanh \frac{1}{2}\alpha_s = 1/\coth \frac{1}{2}\alpha_s$

$$\alpha_s = 2\coth^{-1} e^{(\gamma_\infty - \gamma)} \quad (\text{for } \gamma < \gamma_\infty) \quad \ldots\ldots\ldots (6.5)$$

Eqs. (6.4) and (6.5) may be expressed by a single function when writing:

$$\alpha_s = 2\coth^{-1} e^{|\gamma - \gamma_\infty|} \text{ (nepers)} \quad \ldots (6.6)$$

and this function may also be written in the following form:

$$\alpha_s = 20\log\frac{e^{|\gamma-\gamma_\infty|} + 1}{e^{|\gamma-\gamma_\infty|} - 1} \text{ (db)} \quad \ldots (6.7)$$

In Eqs. (6.6) and (6.7) the image attenuation is only dependent on the distance between the frequency parameter γ and the transformed attenuation peak frequency γ_∞ and, above a linear γ-scale, the attenuation characteristic can therefore be drawn with the help of a template. The template function is given by the Eqs. (6.6) and (6.7) and is shown in Fig. 1.

The reference point of the template curve is γ_∞, that is, the point of infinite attenuation. The curve is symmetrical with respect to a vertical line through this point and the template need therefore be made for half the curve only. It gives the image attenuation characteristic of a full section. For single half-sections an auxiliary template is required with halved attenuation values.

*Dr. Rumpelt is a Divisional Engineer in the Research Section at Headquarters. This article was presented originally as a series of lectures at the Royal Melbourne Technical College and is published with the kind permission of the College.

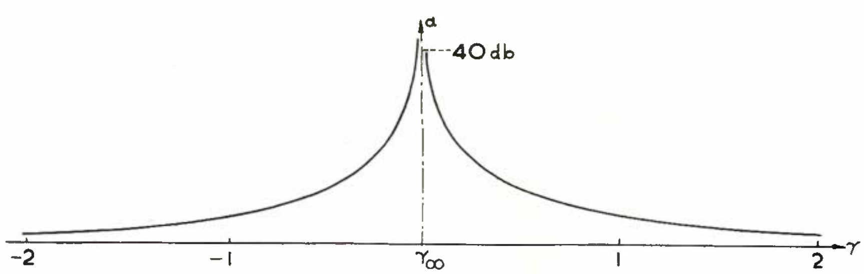

Fig. 1 — Characteristic of Attenuation Template.

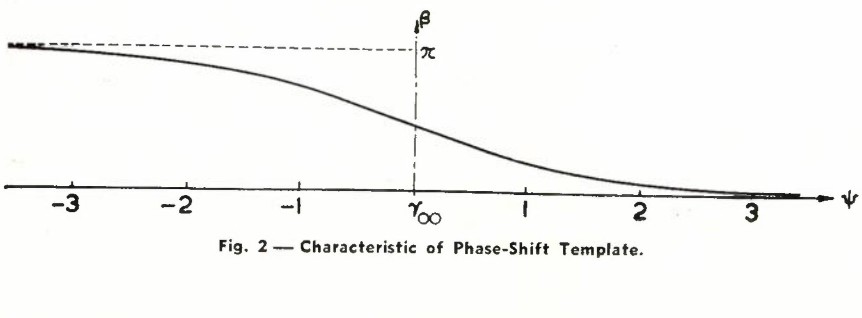

Fig. 2 — Characteristic of Phase-Shift Template.

6.2 Design Procedure with Templates.

The attenuation design with the help of these templates follows the same general line as indicated in paragraph 5.4.2. The calculated minimum attenuation characteristic is plotted against a linear γ-scale which must be identical with the γ-scale of the templates. The transfer of this attenuation characteristic from normalised frequencies, Ω, to the γ-scale is carried out with Eq. (6.2). If the low-pass filter is to be realised in ladder structure and if m-derived or double-derived image impedances are required at the filter terminals, the attenuation peaks of the matching sections are transformed to the γ-scale with Eq. (6.2) or with the relation:

$$\gamma_\infty = \ln m \quad \ldots\ldots\ldots\ldots (6.8)$$

where m is the factor of m-derivation. For each attenuation peak the appertaining attenuation characteristic is drawn with the main template or with the auxiliary template, depending on whether the peak is contained in both matching sections or only in one of them. These curves remain unchanged during the subsequent attenuation design.

Hereafter the number and the positions of the peaks of the additional attenuation curves are estimated which are required to cover the minimum attenuation curve. The peaks are marked on the γ-scale and the appertaining curves are drawn with the proper template. By adding all these curves the total image attenuation characteristic is obtained and must everywhere remain above the minimum curve. The summation is carried out with a pair of dividers for a series of selected points along the γ-axis.

The first choice of attenuation peaks does not normally yield a satisfactory result. The use of templates, however, makes it very easy to estimate the change of the total attenuation by a given shift of any one of the individual attenuation peaks and consequently, together with the quick drawing of the individual curves, the trial and error process is much facilitated by the use of templates.

After the final positions of the attenuation peaks on the γ-axis have been determined, the total image attenuation characteristic of the low-pass filter may be drawn and transferred to the physical frequency scale with the inversion of Eq. (6.2) multiplied by f_c:

$$f = \frac{f_c}{\sqrt{1 - e^{2\gamma}}} \quad \ldots\ldots\ldots (6.9)$$

The m-values or Ω_∞-values of the various sections are obtained from the γ_∞-values by means of the relation:

$$m = e^{\gamma_\infty} \text{ or } \Omega_\infty = \frac{1}{\sqrt{1 - e^{2\gamma_\infty}}} \quad \ldots (6.10)$$

These values and the results previously obtained concerning the image impedances give sufficient information for calculating the component values of the individual sections.

6.3 Use of Special Graph Paper for the Template Design

The transformation from the Ω-scale to the γ-scale, and vice versa, when plotting the required minimum attenuation above the γ-axis and transferring the designed attenuation characteristic and the attenuation peaks back to the Ω-axis, may be avoided if a graph paper is produced with a grid which along the abscissa axis has a distorted Ω-scale fitting to the linear γ-scale of the templates. The formula to be used for calculating the distorted Ω-scale is Eq. (6.2). The correlation between Ω- and γ-values is roughly shown in Table II.

The insertion phase-shift of filters designed as described previously is not linear. Especially in frequency ranges near cut-off frequencies there is a substantial phase distortion and if a linear phase-shift is important for a satisfactory performance of a filter, its phase-shift must be corrected by means of a phase equalising network. For this purpose it is necessary that the insertion phase characteristic in the pass-band of the filter be first determined.

7.2 Difference between Image and Insertion Phase-Shift.

The insertion phase-shift of a filter deviates from its image phase-shift for the same reason that the insertion loss deviates from the image attenuation, that is, because of mismatch between the image impedances and the terminating resistances. The insertion phase-shift is the imaginary part of the expression in Eq. (3.1). In pass-bands, where the image impedances are resistive, only the first and the last term have an imaginary component, the first being the image phase-shift, β, the last being

$$\arg\left[1-\frac{(1-p_1)(1-p_2)}{(1+p_1)(1+p_2)}e^{-j2\beta}\right]$$

With increasing frequency the image phase-shift of a filter increases monotonously, and according to the above expression, the insertion phase-shift ripples around the image phase-shift, the maximum deviation between the two being

$$\pm \text{ ar sin}\frac{(1-p_1)(1-p_2)}{(1+p_1)(1+p_2)}$$

For a reasonably good match between image impedances and terminating resistances this deviation may be neglected. For example: $0.8 < (p_1, p_2) < 1.25$ gives a maximum deviation of only 0.7° or 0.012 radians.

TABLE II.

Ω	1.01	1.02	1.03	1.05	1.10	1.20	1.50	2	3	∞
$-\gamma$	1.963	1.624	1.429	1.188	0.876	0.593	0.294	0.144	0.059	0

7. The Phase-Shift of Filters

In many applications of electric wave filters the phase change of signals passing through the filters has no marked influence on the intelligence conveyed and may be disregarded. There are, however, cases where a phase distortion in signal transmission is very disturbing and must be kept within small limits. This occurs always when the shape of the amplitude versus time characteristic of a signal is essential for the intelligence contained in it, for instance, in the transmission of video signals. In such cases the transient response of filters in the transmission path needs some consideration.

7.1 Influence of Phase-Shift on the Transient Response.

The transient response of an electrical transmission system is described by its output response when a unit-step voltage or its derivative, a voltage pulse of infinitely short duration, is applied to its input. Theoretically, all frequencies from zero to infinity are contained in such signals and any suppression of parts of the frequency spectrum by a filter must cause some distortion. The latter, however, is a minimum if all sine waves of the various frequencies still present at the filter output have the same phase relations to one another as they had at the filter input. In other words, all sine waves should be subject to the same time delay when passing through the filter.

This condition is fulfilled if the insertion phase-shift of the filter in its pass-band rises linearly with frequency, starting at zero frequency with zero or an integral multiple of π. If the pass-band does not start at zero frequency the phase characteristic in the pass-band must be part of a straight line which passes through zero or an integral multiple of π at zero frequency. For relatively small frequency bands it is sufficient if the derivative of the insertion phase-shift with respect to the angular velocity, that is, the differential delay, is constant.

7.3 Determination of the Image Phase-Shift of Low-Pass Filters.

The image phase-shift in the pass-band of a composite filter is the sum of the image phase-shifts of the individual sections. The image phase-shift of a section is determined by its image attenuation characteristic, which in turn is determined by the attenuation peak frequencies.

The image phase-shift of an m-derived low-pass filter section as a function of the normalised frequency is given by Eq. (5.25):

$$\beta_s = 2 \text{ ar cot}\frac{1}{m}\sqrt{1/\Omega^2-1}$$

The factor m in this formula is determined by the attenuation peak frequency through Eq. (5.24). For constant-k sections this factor is 1.

After the attenuation peak frequencies or the corresponding factors m of the various sections in a low-pass filter have finally been established during the image attenuation design, the image phase-shift of the filter can be calculated with the help of Eq. (5.25). This may be done by evaluating the expression

$$\sqrt{1/\Omega^2-1}$$

for a series of Ω-values, dividing each result by the m-values of the various sections and determining from a table for circular cotangens the corresponding arguments, which then are multiplied by 2 and added up. For single half-sections the multiplication by 2 is omitted.

The labour involved in this calculation can be much reduced if a tabular or graphical representation of Eq. (5.25) is used with a series of m- or Ω_∞-values as parameters. It is also possible to make a graphical design with the help of templates.

7.4 Design of the Image Phase-Shift with the Help of Templates

By introducing the frequency transformation

$$\psi = \tfrac{1}{2}\ln(1/\Omega^2-1) \quad \ldots.(7.1)$$

into Eq. (5.25), i.e., by substituting e^{ψ} for $\sqrt{1/\Omega^2-1}$ and by using for m the relation of Eq. (6.10) the function of the image phase-shift of an m-derived low-pass filter section becomes:

$$\beta_s = 2 \text{ ar cot } e^{(\psi-\gamma_\infty)} \quad \ldots.(7.2)$$

This relation has the general form of Eq. (6.1) which means that the image phase-shift of a section can be drawn with the help of a template if it is plotted above the linear ψ-scale. The function of the phase-shift template is Eq. (7.2) and the characteristic is shown in Fig. 2.

The reference point of the template curve is γ_∞ at which the phase-shift is $\pi/2$. The curve is skew-symmetrical with respect to this point. At large positive and large negative values of ψ the curve asymptotically approaches 0 and π, respectively. For single half-sections an auxiliary template is required with halved phase-shift values.

When designing the image phase-shift of a composite low-pass filter with these templates the γ_∞-values of the various sections are marked on the ψ-scale. For constant-k sections is $\gamma_\infty = 0$. For m-derived sections is $\gamma_\infty < 0$. For each γ_∞-value a curve is drawn with the proper template. By adding up all these curves with a pair of dividers at a series of points along the ψ-axis the total image phase-shift characteristic is obtained, which then may be transferred to physical frequencies by means of the inverse transformation:

$$f = \frac{f_c}{\sqrt{1+e^{2\psi}}} \quad \ldots\ldots(7.3)$$

The labour of the transformation step from ψ to Ω may be avoided if the design is carried out on a special graph paper with a grid which along the abscissa axis has a distorted Ω-scale fitting to the linear ψ-scale of the phase-shift templates. This Ω-scale is calculated with Eq. (7.1). The correlation between Ω- and Ψ-values is roughly shown in Table III:

TABLE III

Ω	0.10	0.20	0.40	0.60	0.80	0.90	0.95	0.98	0.99
ψ	2.30	1.59	0.83	0.29	−0.29	−0.72	−1.11	−1.59	−1.95

7.5 Determination of the Differential Delay of Filters

The differential delay of a filter in its pass-band may be determined in a graphical way from the phase-shift curve when the latter is plotted against a linear frequency scale. At any frequency where the differential delay is required a tangent is drawn to the phase-shift curve. For a rise of this tangent by the phase-shift $\Delta\beta$ (in radians) at a frequency increase of Δf the differential delay is:

$$\frac{d\beta}{d\omega} = \frac{1}{2\pi}\frac{\Delta\beta}{\Delta f}\text{(seconds)} \quad ..(7.4)$$

The differential delay may also be calculated from the image phase-shift functions of the individual filter sections by differentiating them with respect to the angular velocity and summing them up.

In the case of a low-pass filter the derivative of the image phase-shift function of a full section, Eq. (5.25), is:

$$\frac{d\beta_s}{d\Omega} = \frac{2m}{[1-(1-m^2)\,\Omega^2]\sqrt{1-\Omega^2}} \quad ....(7.5)$$

where for each section m is as usually determined by the attenuation peak frequency through Eqs. (5.24) or (6.10). For single half-sections the factor 2 must be omitted.

With $\dfrac{d\beta}{d\omega} = \dfrac{d\beta}{d\Omega}\dfrac{d\Omega}{d\omega}$ and $\dfrac{d\Omega}{d\omega} = 1/\omega_c$

the differential delay of the whole filter is given by:

$$\frac{d\beta}{d\omega} = \frac{1}{\omega_c}\sum_{s=1}^{n}\frac{d\beta_s}{d\Omega} \quad ..(7.6)$$

With the help of Eq. (7.5) the derivative of the image phase-shift function of a low-pass filter full-section may be represented as a function of the normalised frequency Ω in tabular form for a series of m- or Ω_∞-values as parameters. The values of $d\beta_s/d\Omega$ in Eq. (7.6) can then be read from this table and they need only be summed up and divided by ω_c to get the differential delay of the whole low-pass filter at any frequency in the pass-band.

TABLE IV—Functions of Attenuation and Phase-Shift Templates.

x	0	0.005	0.010	0.015	0.020	0.025	0.030	0.040	
α_s	∞	52.1	46.0	42.5	40.0	38.1	36.5	34.0	db
β_s	$\pi/2$								radians
x	0.05	0.06	0.08	0.10	0.15	0.20	0.25	0.30	
α_s	32.1	30.4	28.0	26.0	22.5	20.0	18.1	16.5	db
β_s				1.47		1.37		1.27	radians
x	0.40	0.50	0.60	0.70	0.80	0.90	1.00	1.20	
α_s	14.1	12.2	10.7	9.5	8.4	7.5	6.7	5.4	db
β_s	1.18	1.09	1.00	0.92	0.84	0.77	0.70	0.58	radians

x	1.40	1.60	1.80	2.00	2.20	2.40	2.60	2.80	3.00	
α_s	4.4	3.6	2.9	2.4	1.9	1.6	1.3	1.1	0.9	db
β_s	0.48	0.40	0.33	0.27	0.22	0.18	0.15	0.12	0.10	radians

$x = |\gamma - \gamma_\infty|$ for the attenuation template;

$x = \psi - \gamma_\infty$ for the phase-shift template. For negative values of x the phase-shift is $\pi - \beta_s$ where β_s is the phase-shift for the corresponding positive values of x.

A TRANSISTOR AMPLIFIER WITH HEAVY FEEDBACK FOR 12-CHANNEL OPEN WIRE CARRIER SYSTEMS

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1. INTRODUCTION

The Research Laboratories of the Australian Post Office have designed a minor repeater for 12-channel open-wire telephone systems, suitable for out-door mounting anywhere in Australia. Its component part of special interest is its line amplifier. The type of system places severe requirements on the line amplifier performance, as will be seen from the detailed specification to be set out presently, and it is generally accepted as axiomatic that feedback in the order of 40 db must be provided. The performance is customarily specified in relation to a reference level which — to give a brief if not very accurate definition — corresponds to a level of 1 mW transmitted from the trunk switchboard. A transistor amplifier seemed desirable, partly because low power consumption and operating voltage would allow D.C. power feeding over the open-wire pair, partly also for the accumulation of first-hand experience in, and possibly contribution to, transistor circuit techniques.

To list the specifications: Transmission frequencies, 92 to 143 kc/s reference level at the output, 0 dbm or better, with an overload margin of at least 16 db; gain, to follow within 0.1 db a prescribed response which rises from 16.5 db at 92 kc/s to 23 db at 143 kc/s; input and output must match a nominal impedance of approximately 600 ohms with a return loss of 40 db; no intermodulation product falling into the transmission band must exceed −78 db relative to reference level when it is due to two tones of reference level transmitted simultaneously; the amplifier must be open and short circuit stable; the specification must be met, and no damage occur, with ambient temperatures from −20°C. to +80°C.

It has been possible to meet the specification with the circuit to be described later, and to raise the reference level actually to +5 dbm. Some details of performance of an experimental assembly are shown in Figs. 9, 10 and 11, to which reference will be made again later.

This paper briefly states rules for the design of single feedback loops, which may be welcomed by readers not so familiar with the subject, then compares at some length the merits of the three basic transistor connections; then shows how the results were incorporated into the design mentioned before, and, finally, comments on hybrid coil feedback.

2. SYNOPSIS OF RULES FOR FEEDBACK AMPLIFIER DESIGN

The general rules for the design of feedback amplifiers may be found in Bode's classical book, Reference 1. The main points for a single loop design can be summed up as follows.

In order to provide n decibels of feedback in a low pass amplifier, we must be able to control the gain over n/10 octaves, plus one additional octave, of frequency beyond the transmission band (Bode, p. 471). Thus, if an amplifier is to have 40 db of feedback and is to amplify from low frequencies up to 150 kc/s, its gain must be controllable up to 150 kc/s $\times\ 2^5$, or approximately 5 Mc/s. In the present case, useful amplification is needed only in a band of approximately 90 to 150 kc/s, that is in a band of relative width 0.5, centred around 120 kc/s. In this case of a "band pass" amplifier, the gain must be controllable from $120/(0.5 \times 2^5)$ to $120 \times (0.5 \times 2^5)$ or approximately 7.5 to 1500 kc/s. (Bode, Chapter 10, para. 7 to 9.)

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*Mr. Thies is a Group Engineer in the Research Laboratories, Central Office. This article is substantially a reprint of a paper which was presented at the 1959 Radio Engineering Convention of the Institution of Radio Engineers, Australia and which appeared in the February, 1960 issue of the Proceedings of the Institution. It is published with the kind permission of the Institution.

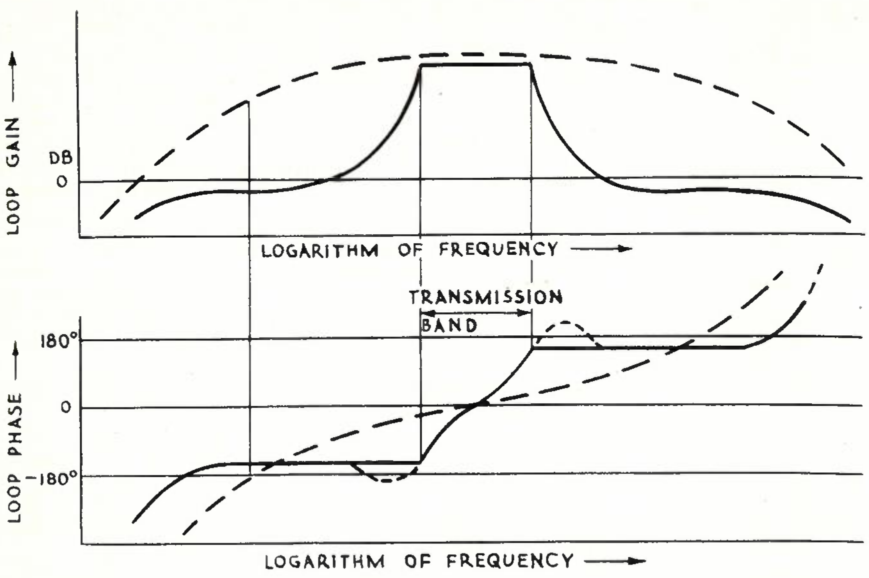


Fig. 1 — Bode Diagrams.

Thus, even with a relatively narrow useful bandwidth, a feedback amplifier must, in the first instance, be a wideband amplifier. Simply feeding back part of the output power to the input would result in Bode and Nyquist diagrams as shown by dashed lines in Figs. 1 and 2. Such a structure would not be stable except with small amounts of feedback. To obtain stability together with maximum feedback, it follows from the general law which relates the minimum phase shift to the slope of the gain-frequency characteristic that the loop response must be shaped somewhat as shown by the full or the dash-dotted curves in Figs. 1 and 2. As much phase-shift as compatible with stability must be provided at all frequencies outside the transmission band, and this must be achieved by minimum phase shift devices.

The full-line characteristic is usually said to be unconditionally stable, and the dash-dotted one, conditionally stable. This terminology is only conditionally correct, as it infers that the loop gain may be partially lost while the phase shift is maintained. The dash-dotted characteristic may be perfectly stable if it is obtained by cathode networks in valve circuits or emitter networks in common emitter transistor circuits and if a loss of mutual conductance or current gain respectively is coupled with a sufficient loss of phase shift. On the other hand, non-linearities coupled with limiting can make the full-line characteristic overload unstable. See also the end of Section 4.

Whether a given basic amplifier circuit is capable of modifications to provide a desired amount of feedback can be estimated from the final asymptote of the gain-frequency characteristic (Bode, Chapter 18, para. 7). The smaller the slope of the asymptote and the higher the frequency at which it intersects with the zero gain line, the larger the feedback which can be provided. The basic amplifier circuit must therefore not owe its wideband characteristics to stagger-tuned circuitry, as this goes necessarily together with a sharp cut-off characteristic.

3. TRANSISTORS IN A FEEDBACK LOOP

We now turn to a discussion of the merits of the three possible transistor connections, common base (CB) common emitter (CE) and common collector (CC). Certain approximations must be made so that useful results can be obtained. They will be found permissible with present-day transistors operating at elevated frequencies in wide-band amplifier stages, but the reader is advised to satisfy himself of their validity in any particular case to which he may desire to apply the results.

The discussions are based on the equivalent T circuits, Figs. 4 to 6. The values of Z_E, Z_B and Z_C and the current gain parameter "a" are identical for all three connections. They are related to the hybrid parameters of the CB connection as shown in Fig. 3. For general considerations we assume that:

$a = \alpha$, $1 - a = 1 - \alpha$ .... (1)

$a = a_0/(1 + jf/f_\alpha)$, f_α is the alpha cut-off frequency .... (2)

$f_\beta = f_\alpha(1 - a_0)$, f_β is the beta cut-off frequency .... (3)

$Z_E =$ constant .... (4)

$Z_C = 1/j2\pi fC_c$, $C_c =$ collector capacitance .... (5)

$Z_E/Z_C << a$, $Z_B/Z_C <<$ .... (6)

$Z_B =$ constant .... (7)

(7) is generally correct with alloy junction transistors. For grown junctions a better approximation is (Reference 2):

$Z_B = Z_{B0}/\sqrt{1 + jf/f_\beta}$ ..... (8)

Consider first single-stage amplifiers, Figs. 4, 5 and 6. With the CE stage we have allowed for an external emitter impedance Z'_E, for reasons which will become apparent presently. We shall assume that $Z'_E/Z_C << a$, also that $(Z_E + Z'_E) < Z_L$.

Using also (6) we may re-write the transfer equations of Figs. 4, 5 and 6.

$$\text{CB} \quad \frac{I_2}{E_S} = \frac{a}{(Z_E+Z_S)(1+Z_L/Z_C) + Z_B(1-a+Z_L/Z_C)} \qquad (9)$$

$$\text{CE} \quad \frac{I_2}{E_S} = \frac{a}{(Z_E+Z'_E)(1+Z_L/Z_C)+(Z_S+Z_B)(1-a+Z_L/Z_C)} \qquad (10)$$

$$\text{CC} \quad \frac{I_2}{E_S} = \frac{1}{Z_E+Z_L+(Z_S+Z_B)(1-a+Z_L/Z_C)} \qquad (11)$$

To find the asymptotes of the voltage gain, we multiply these equations by Z_L and note that for high frequencies $a \rightarrow a_0/jf/f_\alpha$ (from 2), then $1 - a \rightarrow 1 + j0$. We may also assume that $Z'_E \rightarrow 0$ (which can be enforced by a parallel capacitance).

(These asymptotes are meaningful in that they permit to estimate the loop phase shift at not too high frequencies, prior to the insertion of gain and phase shaping networks, and from there to estimate the available feedback. In general, (2) and (9) and (11) contain inadmissible simplifications when frequencies near and beyond the α-cut-off frequency are considered, and (12) to (14) are no longer valid. Wording it differently, the low-frequency phase-shift behaves as if the asymptotes were given by (12) to (14).)

Then, after slight re-arranging:

$$\text{CB} \quad \frac{E_2}{E_S} \rightarrow \frac{a_0}{jf/f_\alpha} \times \frac{Z_L Z_C}{Z_L + Z_C} \times \frac{1}{Z_E + Z_S + Z_B} \qquad (12)$$

$$\text{CE} \quad \frac{E_2}{E_S} \rightarrow \frac{a_0}{jf/f_\alpha} \times \frac{Z_L Z_C}{Z_L + Z_C} \times \frac{1}{Z_E + Z_S + Z_B} \qquad (13)$$

$$\text{CC} \quad \frac{E_2}{E_S} \rightarrow \frac{Z_L Z_C}{Z_L+Z_C} \times \frac{1}{Z_E+Z_S+Z_B+Z_L Z_C/(Z_L+Z_C)} \qquad (14)$$

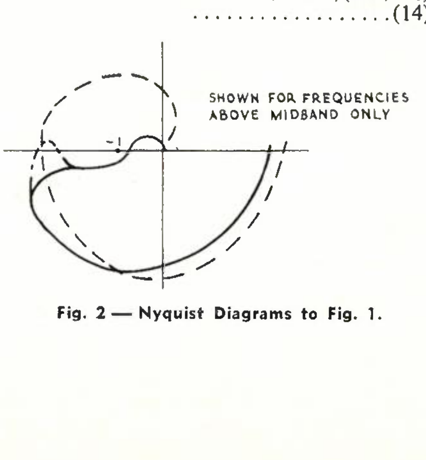


Fig. 2 — Nyquist Diagrams to Fig. 1.

It is immediately obvious that (12) and (13) are identical, that the collector impedance appears in parallel with the load impedance in all three cases, and that the asymptote of the CC stage has no slope due to current cut-off. In the transmission band of the CC stage, of course, $Z_L < Z_S$, if there is to be stage gain. If this holds at high frequencies, the last denominator in (14) will be very nearly $= Z_E + Z_S + Z_B$, and the three equations become identical but for the first factors of (12) and (13). Where source and load impedance levels are prescribed, insertion of numerical values will show which type of stage is preferable.

So far it appears that the CC stage is most suitable for feedback amplifiers. However, a cascaded CC amplifier requires interstage transformers, each adding, in general, a 12 db/octave slope to the final asymptote, because of leakage inductance and high-side capacitance. Moreover the stage gain is low, and we may expect to achieve the same or better results with fewer stages in the CB or CE connection.

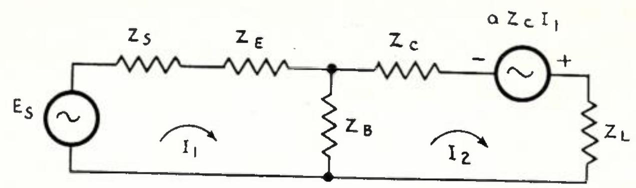
A cascaded CB amplifier also needs interstage transformers. An input transformer also is generally provided with the type of amplifier discussed here. So we next compare CB and CE stages where the level of the source impedance is freely available. Let both stages work into the same load impedance which, of course, is always possible. Now we have to compare output currents for given values of $E_S/\sqrt{Z_S}$. Multiplying (9) and (10) by $\sqrt{Z_S}$ and priming those parameters of the CE stage which may differ from the corresponding ones of the CB stage, we obtain after slightly re-arranging:

$$\text{CB} \quad \frac{I_2}{E_s/\sqrt{Z_S}} = \frac{1}{\sqrt{Z_B}} \times \sqrt{\frac{Z_S}{Z_B}} \times \frac{at}{1 + t(1-a) + (1+t)Z_L/Z_C} \qquad \ldots(15)$$

$$\text{CE} \quad \frac{I_2}{E_S/\sqrt{Z'_S}} = \frac{1}{\sqrt{Z'_B}} \times \frac{1}{\sqrt{Z'_S/Z_B} + \sqrt{Z_B/Z'_S}} \times \frac{at'}{1 + t'(1-a) + (1+t')Z_L/Z_C} \qquad \ldots(16)$$

where,

$$t = \frac{Z_B}{Z_E + Z_S} \qquad t' = \frac{Z'_S + Z_B}{Z_E + Z'_E} \qquad \ldots(17)$$



$$E_S = (Z_S + Z_E + Z_B)\,I_1 - Z_B\,I_2$$

$$0 = (-Z_B - aZ_C)\,I_1 + (Z_B + Z_C + Z_L)\,I_2$$

$$\frac{I_2}{E_S} = \frac{a + \dfrac{Z_B}{Z_C}}{(Z_E + Z_S)\left(1 + \dfrac{Z_B}{Z_C} + \dfrac{Z_L}{Z_C}\right) + Z_B\left(1 - a + \dfrac{Z_L}{Z_C}\right)}$$

Fig. 4 — Common Base Stage.

We restrict the comparison to the case where Z_S, Z'_S and Z_B are purely resistive. We then can achieve $t = t'$ and from an inspection of (15) and (16) both connections will have identical high-frequency response. Small t or t' will minimise the effects of frequency dependance of "a" and Z_C. With t chosen, Z_S is determined by (17). With t' chosen, Z'_S may be selected to meet any desired condition, after which Z'_E follows from (17). The selected condition might be to achieve the best stage gain, i.e., to make the divisor $(\sqrt{Z'_S/Z_B} + \sqrt{Z_B/Z'_S})$ of (16) a minimum. It obviously leads to $Z'_S = Z_B$,

hence $\sqrt{Z'_S/Z_B} + \sqrt{Z_B/Z'_S} = 2 \quad \ldots(18)$

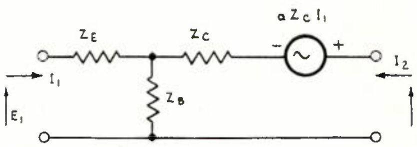
With $t = t'$ and $Z'_S = Z_B$, it follows from (15) and (16) that the stage gains are in the ratio $\sqrt{Z'_S/Z_B} : \frac{1}{2}$. Hence, with identical high frequency response, the CB stage provides the greater gain only if its source impedance is greater than $Z_B/4$. Otherwise the emitter feedback resistance Z'_E necessary to obtain the same response of the CE stage is small enough to allow a greater stage gain. With practical load impedances, the break-even point corresponds to a relatively low power (or strictly transducer) gain $20 \log_{10} [2I_2 \sqrt{Z_L Z_S}/E_S]$. In the output stage of the amplifier to be described later, $Z_L = 1400$ ohm, $Z_B = 1000$ ohm, $Z_E = 5$ ohm. Assuming for an estimate that $a = 1$, $Z_C = \infty$, it follows from (15) that the power gain for $Z_S = Z_B/4 = 250$ ohm is 13.5 db.

The matching condition (18) is not very critical, the gain being down 0.5 db if $Z'_S = 2Z_B$ or $\frac{1}{2}Z_B$, and 2 db if $Z'_S = 4Z_B$ or $\frac{1}{4}Z_B$.

If the assumption that Z_S, Z'_S and Z_B are pure resistances does not hold, the comparison becomes more complex. The general conclusion still holds that cascaded CB stages are preferable to (transformer coupled) CE stages only where low stage gains are provided.

As comparable gains can be obtained with transformerless coupled CE stages, it appears that a cascaded CE amplifier is the most suitable basic circuit for a feedback amplifier. The collector load impedances may be made equal to the base impedances of the following stages, or somewhat larger, so that the matching condition (18) discussed above is mostly met. Load cut-off frequencies due to stray capacitance will be very high and probably negligible. The gain frequency response may be shaped by emitter networks. Stage output and input impedances will be high compared with the load impedances so that there is little interaction between stages and we may design each stage independently, assuming for the time that adjacent stages have infinite impedances (see also below). This provides for a convenient and flexible design.

This procedure being followed, there is a counterpart to the full-shunt terminated low pass filter ("Wheeler" network) discussed by Bode as an interstage network in his chapter 17.3. Referring to our equation (10) and assuming that Z_S, Z_B, Z_C are resistive and Z_E is negligible, the second term in the denominator becomes an inductance at frequencies a few times the β cut-off frequency, while "a" in the numerator is still substantially a constant. The inductance may be considered as the full series reactance of a low pass filter, the remainder of which is to be provided by the external emitter impedance Z'_E. This corresponds to the interstage of a valve amplifier where the stray capacitance is considered as the full-shunt reactance of a low pass filter which is otherwise provided by an interstage network. Only impedances and conductances have been interchanged. The stage gain will be substantially flat up to the cut-off frequency of the low pass filter; from there on the phase shift will be 90°.



$$E_1 = h_{11}I_1 + h_{12}E_2$$

$$I_2 = h_{21}I_1 + h_{22}E_2$$

$$\Delta h = h_{11}h_{22} - h_{21}h_{12}$$

N.B. $h_{21} = -\alpha$

$$Z_E = \frac{\Delta h - h_{12}}{h_{22}} \qquad Z_B = \frac{h_{12}}{h_{22}} \qquad Z_C = \frac{1 - h_{12}}{h_{22}} \qquad 1 - a = \frac{1 - \alpha}{1 - h_{12}}$$

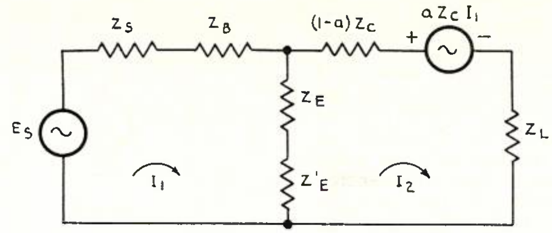
Fig. 3 — Hybrid and Equivalent-T Parameters.

Transistors being bilateral circuit elements, there is always some interaction between stages and the load impedance of the last stage will affect the input impedance of the first stage, the source impedance of the first stage will affect the output impedance of the last stage. The exact transfer equations even for a two-stage amplifier are very cumbersome and necessitate approximating simplifications for a useful interpretation. With numerical values known, this is not particularly difficult, but a general discussion is bound to be tedious and lengthy and will not be attempted.

However, parasitic impedances generally set such limits to interstage impedance levels of a feedback amplifier that interaction between adjacent stages is small, and interaction between non-adjacent stages is negligible. The cascaded CB amplifier furnishes a typical example. It is natural to simplify the consideration of two such stages by assuming that each transistor has unity current gain, so that the overall current gain equals the turns ratio of the coupling transformer, whilst there is additional power gain because of the large output to input impedance ratio. Now the input impedance of the driven stage is, above its β cut-off frequency, largely inductive and doubles for about every octave of frequency. It is, in effect, augmented by the leakage inductance of the transformer and is shunted by the parasitic transformer capacitance and the largely capacitive output impedance of the driving stage. The result is that the driving stage operates effectively into a parallel tuned circuit and that its current gain is increased accordingly. The parallel resonance is likely to be sharp and, unless the turns ratio is small, to occur at a frequency where the loop gain must still be controlled. So that a sharp peak in the loop gain is avoided, it will be necessary to shunt the transformer by a resistance of such magnitude that the high output impedance feature of the driving stage is lost and the source impedance of the driven stage is effectively determined by that shunt resistance, while the driving stage operates under short circuit output conditions. The criterion derived before, whether CB or CE connection is preferable, can now be applied to each stage separately.

The method used before of comparing CB and CE connection is so readily adapted to a comparison for other effects such as non-linear distortion that a digression in that direction will not be out of place. The usefulness of the method lies in that we need know only **that,** but not **how,** the parameters vary. We constructed a CB and a CE stage for a given load impedance so that both were equally affected by the variations of "a" and Z_C with frequency, and compared the gains of these two stages. We might just as well have considered variations of these parameters from other causes, such as ageing, replacement or amplitude.

Thus, from the foregoing comparison for high frequency response, we can state immediately that a CB stage hav-



$$E_S = (Z_S + Z_E + Z'_E + Z_B) I_1 \qquad - (Z_E + Z'_E) I_2$$

$$0 = (-Z_E - Z'_E + aZ_C) I_1 \qquad + [Z_E + Z'_E + (1-a) Z_C + Z_L] I_2$$

$$\frac{I_2}{E_S} = - \frac{a - \dfrac{Z_E + Z'_E}{Z_C}}{(Z_E + Z_E')\left(1 + \dfrac{Z_L}{Z_C}\right) + (Z_S + Z_B)\left(1 - a + \dfrac{Z_L + Z_E + Z_E'}{Z_C}\right)}$$

Fig. 5 — Common Emitter Stage.

ing, with regard to "a" and Z_C, the same non-linear distortion and stability of stage gain as a CE stage, has more gain that the CE stage only if its source impedance (Z_S) is greater than a quarter of the base impedance ($Z_B/4$) provided, of course, that both operate with the same load impedance, collector voltage and current, and that the source impedance level is freely available. As pointed out before, $Z_S > Z_B/4$ implies low stage gains.

To investigate with regard to emitter and base impedances Z_E and Z_B, we rewrite equations (9) and (10) as follows:

Again both stages have identical distortions and equal ratio of output current to signal voltage. Now the CB stage has the greater gain if $Z_S > Z'_S = Z_B$, that is for stage gains 6 db less than correspond to the previous condition $Z_S > Z_B/4$.

4. DESIGN OF THE MINOR-REPEATER AMPLIFIER

Choice of Transistors and of Number of Stages

The expected high ambient temperatures call for silicon transistors. The type used throughout has an α cut-off frequency of at least 8 Mc/s, a β of

$$\textbf{CB} \quad \frac{I_2}{E_S} = \frac{a}{Z_E(1+Z_L/Z_C) + Z_B(1-a+Z_L/Z_C) + Z_S(1+Z_L/Z_C)} \qquad \ldots\ldots\ldots\ldots(19)$$

$$\textbf{CE} \quad \frac{I_2}{E_S} = \frac{a}{Z_E(1+Z_L/Z_C) + Z_B(1-a+Z_L/Z_C) + Z'_E(1+Z_L/Z_C) + Z'_S(1-a+Z_L/Z_C)} \qquad \ldots\ldots\ldots\ldots(20)$$

Equal distortions due to Z_E and Z_B will occur if

$$Z_S(1+Z_L/Z_C) = Z'_E(1+Z_L/Z_C) + Z'_S(1-a+Z_L/Z_C) \qquad \ldots\ldots\ldots\ldots(21)$$

If, first, the source impedance is prescribed, $Z_S = Z'_S$, Z'_E can be chosen to meet (21). Then obviously both stages will have the same gain for the same distortions.

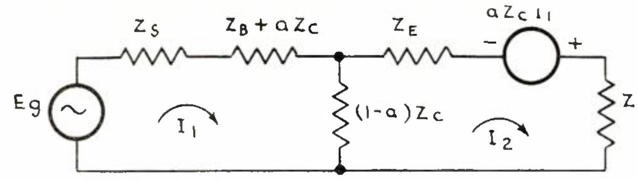
If, secondly, the source impedances are freely available, we can meet (21) without having recourse to an emitter feedback resistance Z'_E, i.e., we can let $Z'_E = 0$ and choose Z_S and Z'_S to meet (21). Both stages again have equal distortions and equal ratio of output current to signal voltage, but the CE stage operates from a higher source impedance, according to (21), and has therefore the better gain.

If, lastly, the matching condition (18) is adhered to, ($Z'_S = Z_B$) and the feedback impedance Z'_E has been chosen, there is always a source impedance Z_S for the CB stage so that (21) is met.

about 40, and can be loaded with 100 mW at 100°C. To obtain the desired over-load power of +21 dbm it was necessary to provide a parallel push-pull output stage of four transistors. With an external gain of some 20 db and a desired feedback of 40 db, more than 60 db of forward gain is required, and it was found necessary to allow three stages, each being given some emitter feedback so that the feedback characteristic is reasonably stabilised against varying transistor parameters. A two stage design would have had to rely to an undesirable degree on transistor parameters for the shaping of the feedback characteristic. The circuit is shown in simplified form in Fig. 7. Details of the output stage are shown in Fig. 8.

Circuit Design

This is based on the recommendations of Section 3. Stage gains are high enough to recommend common emitter connection in all cases. There is a certain deviation from the design principle in that collector networks A and



$$E_S = (Z_S + Z_B + Z_C)\, I_1 \qquad -(1-a)\, Z_C\, I_2$$

$$0 = \quad -Z_C I_1 \qquad + \left[(1-a)\, Z_C + Z_E + Z_L\right] I_2$$

$$\frac{I_2}{E_S} = \frac{1}{Z_E + Z_L + (Z_S + Z_B)\left(1 - a + \dfrac{Z_E + Z_L}{Z_C}\right)}$$

Fig. 6 — Common Collector Stage.

B are provided. Their impedances are high in the transmission band so that they absorb a minimum of power. At frequencies sufficiently outside that band, "B" provides the load for the inter-stage transformer and improves its bandwidth, while "A" provides the correct termination for the output hybrid coil. Both networks provide desirable phase shift and slope of gain. Where first-stage noise is a problem, a similar network can be provided instead of the resistor R1, improving the noise figure by about 3 db.

To obtain open and short circuit stability it is desirable to use some type of bridge feedback so that input and output terminals are decoupled from the feedback path. The "low side" hybrid coil type (Bode, p. 38) was chosen because here shunt inductance and nonlinearities of the transformers are inside the forward path of the feedback loop and small cores can be used. With high side feedback, shunt inductance and nonlinearities are outside the loop.

Network C provides the desired slope of the external gain in the transmission band. This slope equalises the dry-weather loss of the open wire section preceding the minor repeater.

Network D was obtained by a low pass to band pass transformation, the low pass filter being full-series terminated — see Section 3. As the desired bandwidth of 60 kc/s is considerably smaller than the β cut-off frequency of 200 kc/s, the equivalent inductance of the transistor had to be built out to achieve the desired stage response. The 680 ohm resistor of "D" is added to make the additional inductance inoperative at high frequencies — otherwise the inductance would simply degrade the transistor. A shunt capacitance could be provided with the resistor to further improve the stage response at high frequencies. There will then be a series resonance effect between that capacitor and the equivalent inductance of the transistor with a peak of the stage gain. Bode discusses the resulting type of loop response in his Chapter 18, paragraph 7. The exact nature of the peak depends so much on the individual transistor that the device was not used.

The complete loop response was then calculated, excepting the contribution of the first stage. The network E was then designed to approximate 150° of the total phase shift over as wide a band as possible, together with a reasonably flat gain in the transmission band. For final calculations, the exact denominator of the transfer equation, Fig. 5, was used. Also a check was made to verify that interaction could be disregarded.

The effective responses of the various networks may be read from Fig. 9. It is also seen from this graph that the overall feedback is 37 to 41 db, and that the total effective feedback of the output stage is 47 to 49 db.

Intermodulation performance is shown in Fig. 10. The transmission band having less than an octave of bandwidth, all harmonics, all second order products and all two-tone even-order products must fall outside that band. Thus, the lowest significant order of products is that of the $2F_1$-F_2 type. Inherently small second-order distortion is still beneficial because second-order products, particularly when they are subject to little loss in the feedback path, re-modulate with the fundamentals to give third-order products.

Fig. 11 shows impedance graphs. By varying the network impedances of the hybrid coils, the nominal impedances can be varied over a substantial range while the high return losses are maintained. Other performance data have been given in the introduction.

D.C. biasing and stabilising circuitry is conventional. Emitter to ground voltage of the first stage is 2V, collector to ground voltage, 8V. The second stage is directly coupled, with the full battery voltage on its collector. Emitter to ground voltage of the last stage is 1.3V; no-signal current is 5 mA per transistor.

The circuit had some forerunners. The first circuit to be completely calculated and set up provided a flat gain of 20 db and 34 db of feedback. It had a single push-pull output stage designed to deliver at least 16 dbm, but 18.5 dbm were obtained. The desired overload power then was raised to 21 dbm and the circuit redesigned for a parallel push-pull output stage, identical to that of the final circuit; 40 db of feedback was achieved.

Both circuits performed very similarly to the final one and were quite stable. Loop characteristics were of the dash dotted type of Figs. 1 and 2. The circuit was then modified to provide sloping gain, but was unstable once excited to overload. The fundamental oscillation frequency was 75 kc/s, its second harmonic being by far the strongest component of the complex wave. Limiting occurred at the harmonic frequency and the desired reduction of phase-shift at the fundamental did not materialise. The circuit was redesigned to obtain the full line characteristic. It was identical with the final circuit except that the desired response of the first stage was obtained by a much simpler emitter

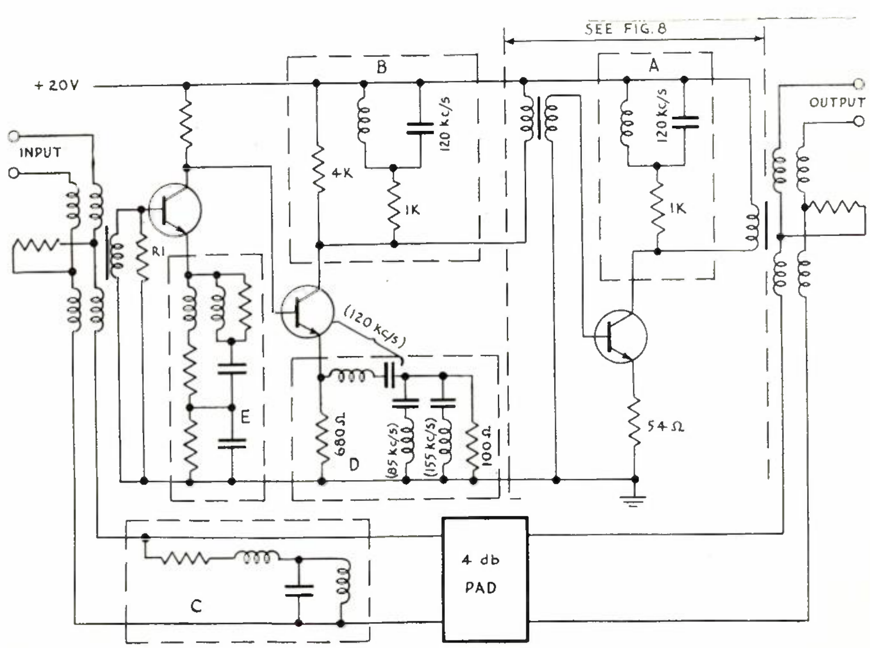


Fig. 7 — Simplified Circuit of Minor Repeater Amplifier, Showing the Parts which Determine the Feedback

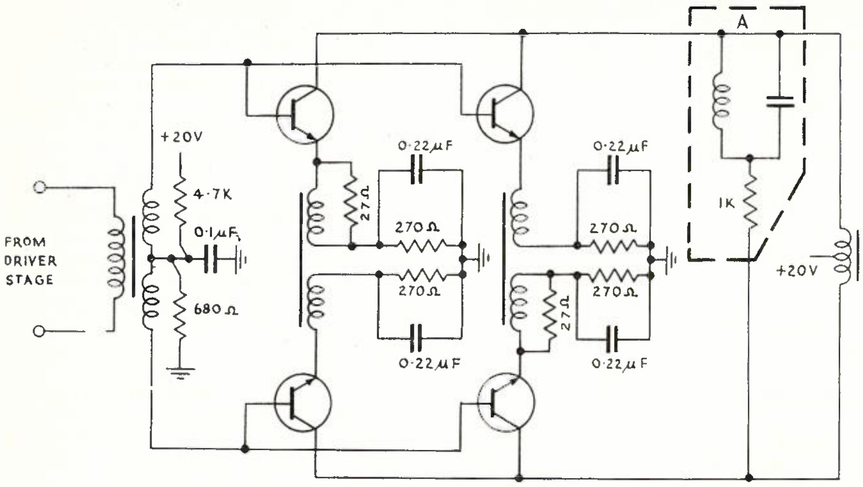


Fig. 8 — Parallel Push-pull Class AB Output Stage.

network, containing one inductor and one capacitor, and by a series tuned damped circuit in parallel with the collector load resistance. The circuit again oscillated after overloading. This time limiting of the oscillation by the small first stage collector impedance was to blame — the resonant frequency of the series tuned circuit being very near the frequency of oscillation. The trouble was successfully cleared by redesigning the first stage to its final form.

5. NOTES ON TWO RELATED TOPICS

Much of the following can probably be found in the literature, but it is assembled here for convenient reference.

5.1 On Hybrid Coil Feedback: The basic theory is found in Bode, Chapter 5.5. The transmission properties of the "low-side" hybrid coil are as follows:

First, let the hybrid coil be an ideal transformer. Let l, c, f be respectively the numbers of turns of the line, collector and feedback windings, and R the desired line impedance. For perfect balance the terminations are:

Feedback terminals: $R_f = R\dfrac{f}{l}$

Collector terminals: $R_c = R\dfrac{c^2}{l(f + l)}$

Network terminals: $R_n = R\dfrac{f}{f + l}$

With correct terminations, the transmission losses are:—

between collector and line terminals:

$$10 \log_{10}\frac{f + l}{l}\text{(db)}$$

between collector and feedback terminals:

$$10 \log_{10}\frac{f + l}{f}\text{(db)}$$

For the impedance Z looking into the line terminals, with collector and feedback terminals terminated in Z_c and Z_f respectively, and with the correct network impedance R_n:

$$\frac{Z - R}{Z + R} = \frac{l}{l + f} \times r_c \times \frac{1}{1 - r_c\, r_f\, f/(l + f)} \times \frac{1}{1 + \mu\beta}$$

where

$\mu\beta$ is the loop gain

$$r_c = \frac{Z_c - R_c}{Z_c + R_c} \qquad r_f = \frac{Z_f - R_f}{Z_f + R_f}$$

With a large amount of feedback, the amplifier impedance is almost wholly determined by the network impedance and the turns ratio of line and feedback windings, no matter how poorly the hybrid is balanced. Or we may say that the return loss is improved by the amount of feedback provided ($20 \log_{10} (1 + \mu\beta)$). To account for shunt and leakage inductance of a practical transformer, let L_1, L_c, L_f be respectively the inductances of its line, collector and feedback windings; let k_{lc}, k_{lf}, k_{cf} be the coupling coefficients, with obvious meanings of the subscripts; and let leakage coefficients σ_l, σ_c, σ_f be defined by $1 - \sigma_l = k_{lc}\, k_{lf}/k_{cf}$, etc.

The practical transformer is equivalent to an ideal one which has:

turns ratios of

$\sqrt{(1-\sigma_l)L_l} : \sqrt{(1-\sigma_c)L_c} : \sqrt{(1-\sigma_f)L_f}$;

a shunt inductance $(1-\sigma_c)L_c$ added to its collector terminals; then has leakage inductances $\sigma_l\, L_l$, $\sigma_c\, L_c$ and $\sigma_f\, L_f$ added to its respective terminals.

Winding resistances appear unaltered in series with the respective windings of the substitute ideal transformer. Winding self-capacitances appear across the respective terminals of the practical transformer, that is, "outside" the leakage inductances. Winding-to-screen and similar stray capacitances need a consideration of the geometry of the winding so as to assess what part of the winding passes the charging current.

Feedback will minimise effects of parasitic impedances on the amplifier

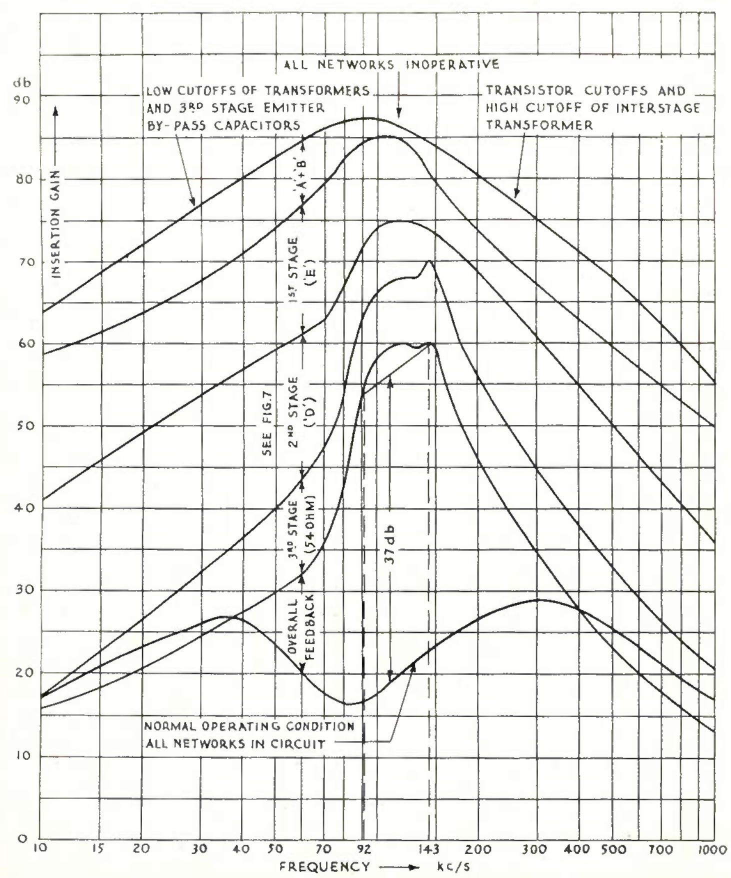


Fig. 9 — Insertion Gain with Step-by-step Addition of Networks (Measured).

impedance if they are inside the loop. The resistance and leakage inductance of the line winding remain unaltered. This leakage inductance can be made to vanish by choice of suitable coupling coefficients. Turns ratios of the substitute ideal transformer are not necessarily equal to those of the practical one. Where extreme return losses are desired, the network impedance R_n must be chosen according to the former.

The transmission properties of the "highside" hybrid coil are readily obtained from the foregoing. The shunt inductance now appears across the line terminals.

5.2 On Multiple-Unit Output Stages: The design of multiple-unit output stages must ensure that the desired load distribution is obtained, unless it relies on selection of closely matched units. Fixed base bias requires individual D.C. emitter resistances where even distribution of the D.C. load is essential. In parallel operation from a common driver stage, both the CB and CE connections require individual A.C. emitter resistances to ensure even distribution of the A.C. load. The same applies to push-pull class A stages which, through the action of the push-pull transformers, are effectively parallel stages. Without such precautions units may not reach the overload point simultaneously, and although overall overload power may be the same, large distortions will occur at lower levels.

Fig. 8 shows the single-ended output stage of Fig. 7 has been translated into a parallel push-pull class AB stage. Base bias is seen to be common to all four units, but each unit has its own bypassed emitter resistance of 270 ohms, so that the emitter current of each unit is stabilised individually. On the other hand, if each unit is given its own A.C. emitter resistor — the 54 ohm resistor of Fig. 7 — large output signals will be flat-topped, because during signal peaks half the transistors cut off. Therefore the emitter feedback voltage has been made proportional to the (algebraic) sum of the total emitter current of each push-pull pair by providing each pair, through an emitter coupling transformer, with a common emitter resistor of 27 ohms. Upon cut-off in one unit, the opposite unit will pass additional current, and the input signal peaks will be faithfully reproduced in the total output current. To see this in more detail, let us idealise by assuming that base currents are negligibly small; that the base-to-emitter voltage drop is zero during conduction; that no collector voltage cut-off occurs; and that we are dealing with a single push-pull stage. Of course, neither the signal voltage on the input transformer, nor the feedback voltage on the emitter coupling transformer can contain D.C. components. So long as always at least one side conducts, the zero voltage drop from base to emitter requires that the base voltage always equals the emitter voltage, so that both have equal D.C. and equal A.C. components. The base D.C. voltage being fixed, the emitter D.C. voltage must be equal for all signal levels, therefore the D.C. drain must be equal to the standing current. Equality of the A.C. components obviously requires the total output current — to which the emitter A.C. voltage is proportional — to be undistorted. However, if the signal level becomes too large, the average current through each side will exceed the standing current, the emitter D.C. voltage will rise and the stage will be completely cut-off for instantaneous signal voltages less than the difference of emitter and base D.C. voltages. Operation will be Class C with corresponding distortion. The limit of undistorted output current — corresponding to pure class B operation — is reached when the rectified average current for undistorted output is equal to the standing current.

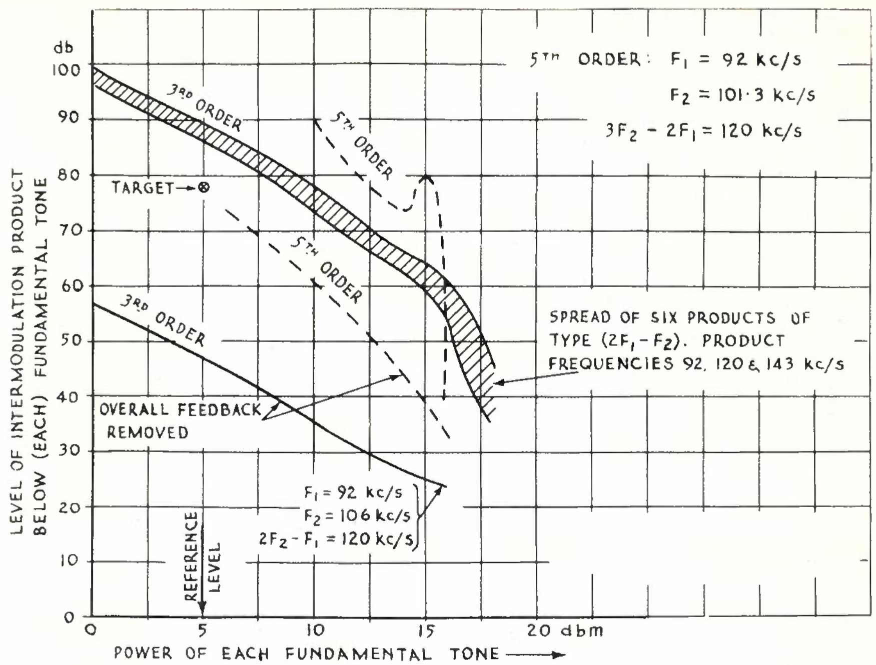


Fig. 10 — Two-Tone Intermodulation Performance.

Now in a class A stage, the peak current for undistorted output equals the standing current. Hence, the collector load impedances being suitably chosen and neglecting saturation resistance, a self-biasing class AB stage has a greater power handling capacity than a class A stage in the ratio of the peak to the average of the signal, if both stages have the same standing current and operate from the same voltage.

If the stage is designed for, and has to handle, very peaky signals, such as speech or multi-channel carrier telephony signals, the overload superiority is very considerable, and much greater than would be expected from a single-tone overload comparison. The class AB stage shares with a class A stage the advantages of even current drain, ready temperature stabilisation, inherently small even-order distortion, and insensitivity to short-circuits in the output — a complete short gives the same dissipation in the maximum signal as in the no-signal condition. The class AB stage shares with class B stages, at least to some extent, the good efficiency with large signals.

In a class B stage where there is no emitter D.C. bias, the principle of emitter feedback by means of an emitter coupling transformer can be used to minimise cross-over distortion, together with linearising the large signal performance.

6. CONCLUSION

The amplifier described in this report shows that the quality of performance and the degree of feedback generally expected from 12-channel open-wire amplifiers can be obtained with transistors.

The design of such amplifiers is somewhat more laborious than of comparable valve amplifiers, because more parameters have to be accounted for. It is substantially more complex only if interaction of stages is a factor. The cascaded common emitter amplifier which we found to be best suited for the purpose leads to small interaction and permits a convenient stage-by-stage design.

Hardly ever has a circuit for com-

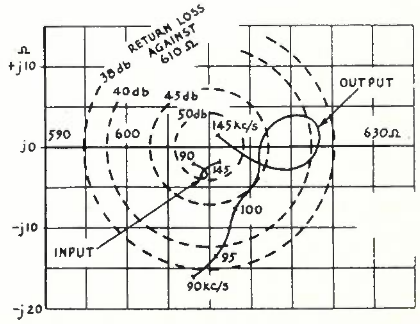


Fig. 11 — Impedance.

munication equipment gone from the paper design stage straight into production. Experiments to verify paper designs are an essential step in the development. To what extent experiments take part in the design procedure is a matter both of particular conditions and of the taste of the designer. A paper design probably facilitates an assessment of the effects of spread in the transistor parameters. The only experiments that were conducted in conjunction with the design of the first "forerunner" circuit described in Section 4 were impedance tests on a trial output transformer, made at frequencies not higher than 150 kc/s. After an error in the D.C. bias circuitry and a few wiring errors in the physical realisation had been cleared — to err is human — the circuit performed satisfactorily and did not require further amendments. Naturally, the experience gained was used in the later designs.

When considering the complexity of our circuit, Fig. 7, in which 9 inductors and 8 capacitors are used to "shape the gain", it should be remembered that some of these are used to obtain the desired frequency response of the external gain, and that part of the complexity is a consequence of the feedback being taken over three transformers; also that because of the high operating temperatures, silicon transistors are used which have inherently less gain than germanium transistors. As further improved transistors become available, and in particular if a high-frequency power transistor replaces the cumbersome push-pull output stage, simpler circuits will be possible.

7. ACKNOWLEDGMENTS

Acknowledgment is made to the Engineer-in-Chief of the Postmaster-General's Department for permission to publish this paper. The author also wishes to thank Mr. D. A. Gray of the P.M.G. Research Laboratories, for carefully studying the original manuscript and pointing out several errors therein.

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TECHNICAL NEWS ITEM

New Method of Jointing Plastic Conductors

Lineman Grade II N. Symons of Mosman, Sydney, has won an Improvements Board award of £1000 for an improved method of jointing plastic covered wires. The method is considered to be a major development by world standards and is protected by an application for world patent rights. It should allow full use to be made of the non-hygroscopic nature of the plastic insulation in communication cables.

When the Department first began using plastic cable, wire joints were made conventionally by stripping wires, twisting them together and then insulating the joint with a plastic sleeve. Individual wire joints were not sealed from moisture and no attempt was made to internally seal off the joint area from movement of moisture inside the sheath. Complete reliance was placed on the overall sheath seal and the whole procedure closely followed the jointing methods used for paper insulated lead sheathed cables even though the plastic cable with its polythene sheath and polythene wire insulation was fundamentally different.

Early experience with plastic cable in Australia as well as in England and America showed that, unlike paper-lead cable, almost all faults occurred at joints. It was obvious that if moisture could be prevented from reaching the wire joint itself, the performance of plastic cable, already better than that of paper-lead cable, would be even further improved. Two methods were adopted to achieve this. The first came from England, where water barriers were used at the ends of the plastic cables entering a joint. This is an internal seal which prevents water present inside the plastic cable from entering the joint area. Thus the English answer was to seal the cable both internally and externally.

The second method came from America, where a special sealing sleeve was used to make each wire joint waterproof. The special sleeve was made of plastic and filled with silicone grease.

For plastic to plastic joints the wire joint seal is favoured because with the water barrier approach faults can still occur if the sheath seal fails and experience with all kinds of sheath sealing techniques has been unsatisfactory both here and overseas.

Until the advent of the Symons hot twist method of wire jointing the only method available was the grease filled sleeve. This was not entirely satisfactory because —

(i) sleeve could be dislodged;
(ii) grease could be squeezed out;
(iii) soldering of wire joints was necessary because of possible deleterious effects of the grease on copper;
(iv) the sleeves were relatively expensive;
(v) different sizes had to be stocked for different conductor sizes.

Because of these shortcomings much effort was directed both here and overseas to develop a better sealed wire joint. The approach adopted mostly was to try to find a more effective sealing sleeve.

The Symons method produces an effective sealed wire joint without special sleeves of any kind using a technique which is easily learnt and which requires only a source of heat. The method is to twist the unstripped wires loosely together, apply heat to the twisted wires to soften the plastic and then twist the wires together using the normal crank twist. This is illustrated in Fig. 1. The wires move through the softened plastic and make a good elec-

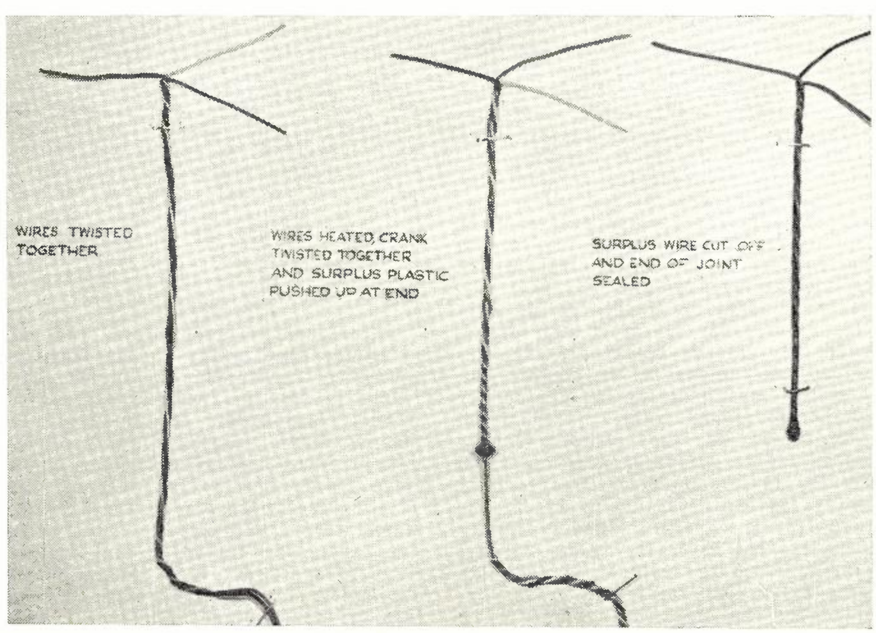

Fig. 1 — Multiple Joint made by Hot Twist Method.

trical joint whilst the plastic reseals itself around the wire twist. Surplus wire is then cut off but in doing so plastic is pushed back over the wire twist for about a half-inch to form a blob of plastic at the end of the joint. The exposed wire ends are then sealed by heating this blob of plastic causing it to seal over the end of the joint.

The advantages of this method are —

(i) a permanent waterproof seal is produced;
(ii) stripping of wires is unnecessary;
(iii) soldering is unnecessary;
(iv) no special sleeve is required;
(v) technique is easily learnt;
(vi) for polythene insulation any source of heat may be used. Standard acetylene torches are being used but effective joints can be made using a cigarette lighter. (For P.V.C. covered wires a bare flame is unsatisfactory and radiant heat is required.)

Tests in the laboratory and in the field have proved the effectiveness of this method and it has now been adopted as the standard method for all plastic to plastic joints in the Australian Post Office.

READERS' SURVEY

About 350 subscribers answered the following two questions when renewing their subscriptions last year:—

"(a) Which article in Vol. 11 of the Journal did you find most interesting?
(b) On what subject would you like to see an article printed?"

The results of this survey are published for the information of readers and comments are invited.

Most Interesting Article: The votes received for most interesting article in each issue of Volume 11 are given in Table 1.

An analysis of all the votes for all articles in Vol. 11 was made in terms of type of article, and this is compared with the actual content of the Volume in Table 2.

Requests for Articles: The requests for articles were grouped as far as possible into six main types and the most frequently requested subject in each group is shown in Table 3.

All requests were classified into types in order to arrive at a desirable balance of articles for future issues and Table 4 shows the results of this analysis.

Conclusions: This survey has indicated that more telephone equipment articles, particularly on crossbar equipment, are required and that general review type articles on latest developments are popular. It would appear that there have been too many articles on external plant topics and not enough on radio and television.

The Board of Editors has already taken action as a result of this survey, and we expect Volume 12 to adequately meet the needs of our readers.

TABLE 1: Most interesting article in each issue of Volume 11.

Issue	Article	Votes Received
1.	Television in Australia — V. Kenna	27
2.	The SE.50 Group Selector Circuit — F. Scott, L. Wright	21
3.	The Victoria-Tasmania Radio Telephone System (via Flinders Is.) — W. Beard	17
4.	Why the Australian Post Office will Eliminate Letters from Telephone Numbers — B. Marrows	8
5.	Features of the Crossbar Exchange at Templestowe, Victoria — L. Haig	39
6.	Principles of Crosstalk and Noise Suppression at Open Wire and Balanced Cable Carrier Stations — S. Dossing	26

TABLE 2: Comparison of votes for most interesting article compared with actual content of Volume 11.

Type of Article	Per cent. of votes by readers	Per cent. appearing in Vol. 11
Telephone Equipment	43	33
Long Line Equipment	22	22
Radio and Television	16	8
Lines	4	22
Telegraphs and Workshops	1	5
General	14	11

TABLE 4: Requests for articles classified into types.

Type of Article	Per cent. of requests received
Telephone Equipment	47
Long Line Equipment	12
Radio and Television	14
Lines	5
Telegraphs and Workshops	2
General	20

TABLE 3: Most frequently requested article of each type.

Type	Subject	Number of Requests
Telephone Equipment	Crossbar	60
Long Line Equipment	Coaxial Cable Systems and Equipment	15
Radio and Television	Broad Band Bearer Radio Links	14
Lines	Mechanical Aids	4
Telegraphs and Workshops	Tress	2
General	Transistors	25

EXPERIMENTAL SUBSCRIBER TRUNK DIALLING EQUIPMENT

*J. A. PRYOR**

INTRODUCTION

The planning of a Commonwealth-wide scheme for Subscriber Trunk Dialling (S.T.D.) has been in progress for several years and, early in 1958, it was decided that some experience of subscriber dialling over the trunk network was desirable. With this object in view, the Equipment Laboratory of the Telephone Equipment Section was requested to design equipment which would permit a number of official telephone services in the Central Administration to have access to the operator dialling network.

As an adjunct to this trial of S.T.D., it was decided to test a method of transmitting meter pulses on a two-wire junction during conversation. This facility permits multi-metering tariff determination equipment to be concentrated at a central point, for example, a main trunk centre, without recourse to three wire circuits (two wires for the speech channel and one wire for meter signals) to the terminal exchanges. The multi-metering equipment at the terminal exchanges is thereby simplified because it is not required to determine the appropriate tariff rate but only has to receive the meter signals from the central equipment. If the complex tariff determination equipment is not centralised, this equipment would have to be installed at all exchanges from which S.T.D. calls could orginate.

The equipment necessary for the S.T.D. trial was designed and assembled in the Laboratory during 1958 and the units were installed at two Melbourne exchanges — Russell Exchange (a city main exchange to which the Central Administration services are connected) and the automatic trunk room at City West Exchange (the trunk centre in Melbourne) — with the assistance of staff from the Victorian Administration. It was brought into service in mid-October, 1958, and has been in continuous use since that time.

REQUIREMENTS FOR THE EXPERIMENTAL EQUIPMENT.

The main purpose of the equipment is to provide S.T.D. facilities for a limited number of telephone services. Therefore, it was necessary to design equipment which would permit these services to have access to the operator trunk dialling network, while preventing other telephone services from obtaining access to the S.T.D. equipment. Ordinary services are barred access to the trunk network even when the correct access code is dialled by making a minor modification to the line circuits of the official services.

The S.T.D. equipment was designed so that it would accept connections from the modified line circuits but would reject ordinary line circuits.

The selection of an access code which ordinary subscribers would be unlikely to dial accidentally was another requirement. In Melbourne at that time, all levels from the first selectors except level "1" were in use for the trunking of the network. The last manual exchange in the inner city area — Central exchange with dialling code "0" — had been replaced by automatic equipment a short time earlier but the level was still being intercepted when the S.T.D. equipment was being prepared. Therefore, although level "0" was in reserve for the nation-wide S.T.D. system and ancillary services, it could not be used at that stage. In consequence, it was decided that level "1" would be used. To avoid the use of ranks of special group selectors to obtain an access code for the S.T.D. equipment, it was further decided to use a digit absorption relay set for the purpose. A suitable circuit had been developed by the New South Wales Administration for use in the Sydney network, and this relay set was adopted for the S.T.D. trial. The code for access to the S.T.D. equipment became "110" — of which the last two digits were absorbed by the relay set.

The further requirement that metering signals should be transmitted over a two-wire junction line was examined and it was decided that, for the trial, the metering signals would consist of pulses of low-voltage 50 c.p.s. A.C. connected cailho fashion, that is, as a superimposed signal, on the two-wire junction line. After rectification at the receiving end, the signals would operate relays to cause registrations on the meters of the telephone services.

As periodic time-zone metering was to be used for the trial, a range of periods was selected for the purpose. The periodic pulses were derived from an experimental tariff pulse generator which will be described in a later issue of this Journal.

GENERAL DESCRIPTION OF THE SYSTEM

A trunking diagram of the system is shown in Fig. 1.

When a call is made via the S.T.D. equipment, the digit "1" is dialled on the local first group selector at Russell exchange and a selective digit absorption relay set is seized. The digits "10" are then dialled to obtain access to the S.T.D. equipment. The intermediate repeater is seized, and the associated variable rate repeater and subscribers trunk selector are operated at City West Exchange. The next digit (the first digit of the trunk code) causes the subscribers trunk selector to search for an idle outlet in the transit group and the second trunk code digit causes a transit group trunk selector to search for an idle outlet in the group of trunk lines to the required trunk centre. When the 2VF trunk line relay set is seized, a discriminating signal causes the subscribers trunk selector to change from D.C. loop pulsing to a tone pulsing arrangement. Subsequent digits are transmitted as pulses of "X" tone (750 c.p.s.) to the distant trunk centre to operate the selector equipment there and establish the connection to the called service.

Simultaneously, the first two trunk code digits operate the code selector associated with the variable rate repeater and discrimination is effected for the digits which have been dialled. The discriminating arrangement prevents access to barred levels or, on working levels, provides a signal whereby the tariff selector in the variable rate repeater can select the tariff rate appropriate to the trunk code which has been dialled.

When an answer signal is received, an A.C. meter pulse is superimposed on the junction line by the variable rate repeater, and the intermediate repeater effects a registration on the meter of the calling service at Russell exchange. A "free" period of ten seconds follows before the periodic pulses from the tariff pulse generator cause further A.C. signals to be connected to the junction line.

The established connection is released by the calling telephone, and if a C.S.H. condition occurs, a forced release action is initiated by the variable rate repeater.

SUBSCRIBER'S LINE CIRCUIT

The services which were connected for the trial were direct lines with uniselected line circuits. To permit the S.T.D. equipment to differentiate between trial services and ordinary services, it was necessary to effect some change to the line circuit of the trial services.

In earlier multi-metering equipment which has been designed to permit multi-metering access on a limited scale, differentiation between barred and non-barred services was obtained by shunting the cut-off relay (K) of the barred line circuits with a non-inductive resistor. A detector circuit (which is described later in this article) was connected to the P wire by the multi-metering equipment, to recognise the modified circuit condition, and prevent access to the multi-metered trunk lines by the barred services. In such cases the line circuits which were barred access were mainly public telephones and they represented only a small proportion of the telephone services.

When the barred services are in the minority, this approach is feasible, but it was not possible to apply it in the case of the S.T.D. trial. Instead, the opposite action was taken. The resistance of the cut-off relay in the trial services was increased and the detector circuit in the repeater was modified so that it permits these services to make calls via the S.T.D. equipment, but ordinary services with the normal cut-off relay coils are denied access.

The affected part of the uniselector line circuit is shown in Fig. 2. The

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standard 1300 ohm coil of relay K has been replaced by a coil of 2600 ohms. The simpler measure of connecting a 1300 ohm resistor in series with the normal coil was not feasible because the operating current would be reduced and the relay would be unreliable in operation. However, it is necessary that the testing relay (H) in final selectors shall operate in series with the cut-off relay in the line circuit. To permit the operation of the testing relay a 2600 ohm non-inductive resistor is connected to the multiple P wire. When the uniselector is normal, the effective resistance of relay K, the uniselector magnet and the resistor is approximately 1400 ohms, that is, the normal condition for operation of the testing relay in the final selector. When a call is being originated by the trial service the uniselector is off normal, the resistor is not in parallel with the relay, and the unshunted resistance of relay K permits the service to make calls via the S.T.D. equipment.

In all other respects the operation of the line circuit of the trial service is unchanged from that of an ordinary service, and the trial service can also initiate calls via the local network as desired.

SELECTIVE DIGIT ABSORPTION RELAY SET

This relay set was used in the trial because it was necessary to use first digit "1" but, as it is not essential to the operation of the trunk dialling and multi-metering functions of the S.T.D. equipment, it has not been illustrated. The primary purpose of the relay set is to prevent the false traffic which is encountered on the first level from obtaining access to the S.T.D. equipment. To complete a connection from the line circuit to the intermediate repeater, it is necessary to dial the additional digits "10". When these digits are received, the digit absorption relay set switches the calling line to the repeater with no bridging apparatus on the negative and positive wires and the relay set is held by the guarding earth which is connected to the P wire by the intermediate repeater. If any other combination of digits is dialled, N.U. tone is transmitted to the calling service.

One digit absorption relay set is associated with each intermediate repeater and, as there are only a limited number of S.T.D. circuits, the relay set incorporates a forced release facility which ensures that false traffic cannot hold an S.T.D. circuit for longer than twenty seconds.

INTERMEDIATE REPEATER

The primary purpose of the intermediate repeater, Fig. 3, is to receive the A.C. metering signals which are superimposed on the speech pair by the variable rate repeater and to cause appropriate registrations on the exchange meter associated with the calling service. The repeater also incorporates circuit elements which::—

(i) repeat dial pulses to the variable rate repeater;
(ii) deny access to ordinary services;
(iii) detect a stop-on-busy condition and prevent false metering;
(iv) transmit N.U. tone to the calling service if a reversed circuit is encountered when a connection is being established;
(v) apply a guarding earth to the incoming P wire if the junction line to the trunk centre becomes faulty, and operate a supervisory alarm;
(vi) provide timing of the release unguard period and prevent seizure of the releasing circuit if another selector "tests-in" during the release unguard period; and
(vii) force the release of the connection if a release signal is received from the variable rate repeater.

The intermediate repeater uses a transformer bridge arrangement as this permits superimposed signals to be applied to the circuit with negligible coupling effects between the incoming and outgoing sides of the repeater. It also permits the use of the low inductance high speed relay for impulse repetition and this results in an improvement in pulsing performance in comparison with the usual capacitor bridge and high impedance 3000-type impulsing relay. For comparable pulse distortion, the maximum resistance limit of the dialling loop can be increased from 1000 ohms (capacitor bridge) to about 2000 ohms (transformer bridge). The transformer/relay arrangement has been adapted from Siemens-Edison-Swan practice and it employs their small transformer which occupies the space of one 3000-type relay.

Circuit Operation

When the selective digit absorption relay set connects the calling line to the repeater, relay A is operated by the loop of the calling telephone. Relay A operates relay B and loops the outgoing junction to seize the variable rate repeater at City West Exchange. Relay B operates relay BB and BX, and an earth is connected to the incoming P wire to guard the connection and hold the line circuit and selective digit absorption relay set. Relay G is disconnected from the outgoing junction by relay BB.

When the calling service dials, relay A pulses. Relay CD operates on the first release of relay A, and both relays B and CD hold during dialling. Relay CD connects a short-circuit loop for impulse repetition on the outgoing junction and operates relay CC. At the end of each digit, relay CD releases slowly and relay CC releases after a delay of 0.75 to 1.5 seconds. When the called service answers, the potentials of the outgoing junction are reversed by the variable rate repeater and relays F, FR and FRA operate in turn. The potentials on the incoming line are reversed for supervisory purposes and the meter signal transformer is introduced into the incoming line.

Meter signals are transmitted by the variable rate repeater as 50 c.p.s. A.C. pulses superimposed on the junction line. These pulses are rectified by MR2, and relay AC operates in unison with each pulse. Relay AC connects positive battery to the incoming P wire to operate the exchange meter of the calling service.

When the calling service replaces the receiver, relays A and B release in turn. The guarding earth is disconnected from the incoming P wire and relay BB releases. The release unguard period

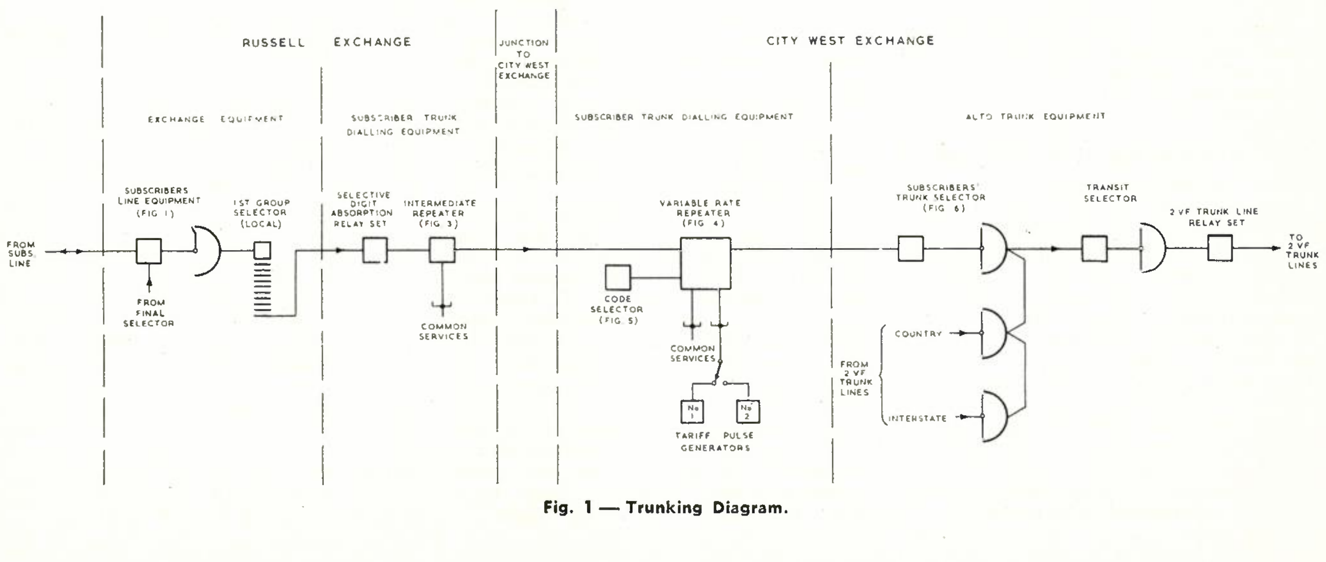

Fig. 1 — Trunking Diagram.

occurs during the release time of relay BB. When it releases, the guarding earth is reconnected to the incoming P wire.

Relay G is connected to the outgoing junction line and, when the line potentials return to normal, the relay operates. The guarding earth is disconnected from the incoming P wire and the repeater is available for another call.

Forced release from the C.S.H. condition is initiated by the variable rate repeater and a release signal is transmitted to the intermediate repeater by connecting positive battery to the positive wire of the junction. Relay F releases (if operated) and relay RL operates. Relay RL releases relay A. Relays B and BB release in turn and, when relay G operates, the repeater is available for another call.

If the repeater is releasing and another call is connected during the release unguard period, relay BB holds to the earth which has been connected to the incoming P wire by the intruding call. Relay RL operates because relay B has

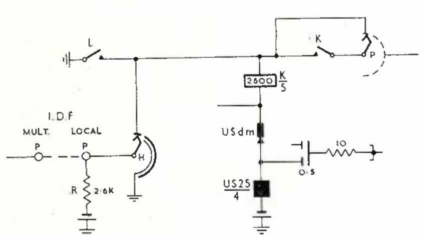

Fig. 2 — Part of Uniselector Line Circuit.

released, and prevents the intruding call from reoperating relay A. When the intruding circuit disconnects the earth on the P wire, the call will be released because the intermediate repeater has not connected a holding earth. Release of the repeater will be resumed and, when relay BB releases, it will be available for another call if relay G has operated.

If the junction becomes faulty, that is, open-circuit, short-circuit, reversed, or earthed on the negative leg, relay G will release and the delayed alarm circuit is operated. After a delay of 30 to 60 seconds, audible and visual alarms are given and the supervisory lamp on the intermediate repeater glows.

Safeguards against false metering are described later.

VARIABLE RATE REPEATER

The variable rate repeater, Fig 4, in association with the code selector, determines the tariff rate applicable to the trunk code which has been dialled and, when an answer signal is received, transmits the meter signals to the intermediate repeater in the form of 50 c.p.s. A.C. pulses superimposed on the junction pair.

The repeater also incorporates circuit elements which:—

(i) provide impulse repetition to the code selector and subscribers trunk selector;

(ii) detect a stop-on-busy condition and prevent false metering;

(iii) apply the first meter pulse within two seconds after the answer signal has been received and then provide a "free" period of ten seconds before connecting the appropriate tariff pulse lead to the metering relay;

(iv) superimpose 50 c.p.s. A.C. signals on the junction pair to transmit meter pulses to the intermediate repeater;

(v) maintain a guarding earth on the incoming P wire if the equipment fails to restore to normal, and operate a supervisory alarm;

(vi) transmit N.U. tone to the calling service if a barred level code is dialled;

(vii) initiate forced release of the connection if P.G. or C.S.H. conditions occur, and transmit a release signal to the intermediate repeater;

(viii) transmit N.U. tone to the calling service if a reversed circuit is encountered when a connection is being established.

The variable rate repeater uses a transformer bridge arrangement similar to that of the intermediate repeater.

Circuit Operation

Relay A operates when the junction is looped at the intermediate repeater. Relay A operates relays B (repeater) and AA (subscribers trunk selector). Relays BB, E, BC, and BX operate.

When the calling service dials the first trunk code digit, relay A pulses and relays AR (repeater) and AA (subscribers trunk selector) pulse in unison. The code selector steps and relay ON operates when the selector moves from the home position. Relay CD operates on the first release of A, and both relays B and CD hold during the digit. Relay CD operates relay CC. At the end of each digit, relay CD releases slowly and relay CC releases after a delay of 0.75 to 1.5 seconds. When the second trunk code digit is dialled, relays A and AR pulse, relays CD and CC operate, and the code selector steps again. On subsequent digits, relays CD and CC operate when each digit is dialled, but the code selector does not step.

When the called subscriber answers, earth is connected to the D lead by the subscribers trunk selector and relays FA, FR, and FS operate in turn. Contact units of relay FR reverse the potential on the junction pair for supervisory purposes and introduce the meter signal transformer into circuit.

Relay MC is connected to the one second pulse lead and, when the relay pulses at one-second intervals, it steps the tariff selector. On the second step, relay J operates and an AC signal is superimposed on the junction pair. After nine steps, the tariff selector self-drives until it detects a mark from the code selector on the MS4 bank and relay MC re-operates and cuts the drive. Relay J is connected to the tariff pulse lead and it is pulsed by the periodic time pulses from the tariff pulse generator (T.P.G.). Each time relay J operates, the meter signal is superimposed on the junction pair. Metering continues until the equipment is released.

When the calling service replaces the receiver, relays A, B and BC release in turn. The guarding earth is disconnected from the incoming P wire, relay BB releases and the release unguard period occurs during the release time of relay BB. Relay BX releases in turn. Meanwhile, when relay B released, relays FR and FS were released and the code selector and subscribers trunk selector commenced to restore to normal and, when relay BC released, the tariff selector followed suit. When all selectors reach their home position, relay ON releases. The guarding earth is disconnected from incoming P wire and relay A is re-connected to the negative wire of the junction to operate relay G in the intermediate repeater.

When a barred level is dialled, either on the first or second code digit, the code selector connects earth to the NU lead after the last impulse of the digit and relay FA operates during the slow release of relay CC. Relay FS operates and relay CC holds. N.U. tone is transmitted to the calling service and the subscribers trunk selector is released so that a trunk line will not be held unnecessarily. Further dialling is ineffective.

Forced release occurs under the following four conditions:—

(i) **No Digit Dialled (P.G. Condition):** When the repeater is seized initially, relay BC is operated and connects earth to the start lead of the three-minute delay circuit. When the S pulse is received, relay SP operates and locks. After a further three minutes, the Z pulse is received and relay SZ operates and locks. Relay SZ causes relay ON to operate, releases relays A and SP, and connects positive battery to the positive wire of the junction pair to transmit a release signal to the intermediate repeater. Relays B, BC, BB, and SZ, and ON release in turn, and the repeater is available for another call.

(ii) **Digits Dialled But No Answer Received:** As described in the previous paragraph, relay SP will operate when the S pulse is received. However, it will be released if a digit is dialled, that is, if relay CD operates subsequent to the operation of relay SP. After the last digit, relay SP will operate and lock when the S pulse is received. After the delay period, relay SZ operates and forced release will be effected as described above.

(iii) **Called Subscriber Held (C.S.H. Condition):** If the calling service holds the connection after the called subscriber has replaced the receiver, relay FA will release. This will close the circuit for relay SP, which was opened when the answer signal was received. Relay SP will operate and lock when an S pulse is received. After the delay period, relay SZ will operate and forced release will be effected as described above.

(iv) **Call Being Held on Barred Level:** In this case, relay FA (but not relay FR) will have operated. Relay SP will operate and lock when an S pulse is received. After the delay period, relay SZ will operate and forced release will be effected as described above.

If any of the associated selectors, that is, tariff selector, code selector or subscriber's trunk selector, fail to restore to normal when the repeater releases, relay ON will remain operated after relay BC releases. The supervisory alarm circuit will be operated and after 30 to 60 seconds audible and visual alarms are given and the supervisory lamp on the repeater glows. The negative wire is open-circuited so that relay G in the intermediate repeater cannot operate and the circuit is guarded.

Safeguards against false metering are described later.

CODE SELECTOR

The code selector, Fig. 5, is used in association with a variable rate repeater. It provides two-digit discrimination between working and barred levels, and determines the tariff rate applicable to calls via the working levels. For the purposes of the trial, interstate trunk codes are available as working levels but most of the Victorian intra-state trunk codes are barred. The connections shown in the figure are those which are used in the experimental S.T.D. equipment.

Circuit Operation

When the repeater is seized, earth is connected to the BX and CD Op. leads to operate relays BS and CDR respectively, and relay NU operates.

When the first trunk code digit is dialled, relay AR (repeater) pulses and the CS magnet is pulsed via the CS magnet lead. During impulsing, relay CDR holds via the CDR Hold lead but it releases at the end of the digit and the discriminating relay(s) DB, DL, DU and DUA, or DW and DWA operates via the CS5 bank. Relay HR operates and locks, and the uniselector self-drives to contact 12. Relay HR releases and relay CDR re-operates.

When the second trunk code digit is dialled, the CS magnet is pulsed again. Relay CDR holds during impulsing and, when it releases at the end of the digit, earth is connected to a tariff marking lead via contacts of the discriminating relay which was previously operated, and thence to the bank of the tariff selector in the variable rate repeater so that the tariff pulse lead appropriate to the trunk code can be selected.

Relay NU is held via other contact units of the discriminating relay. Subsequent digits do not affect the code selector.

When the repeater releases, earth is disconnected from the BS lead. Relay BS releases and, in turn, the discriminating relav(s) releases. Relay HR operates and earths the ON lead to indicate that the CS uniselector is off-normal. The uniselector self-drives to the home contact, relay HR releases and disconnects earth from the ON lead, and relay NU releases.

On a barred level call the circuit is seized and the first trunk code digit is received as described above. At the end of the digit, relay CDR releases and relay NU starts to release slowly.

If a barred level has been dialled, a discriminating relay will not be operated. Therefore, relay HR is not operated and, when relay NU releases, earth is connected to the NU lead to operate relay FA in the repeater.

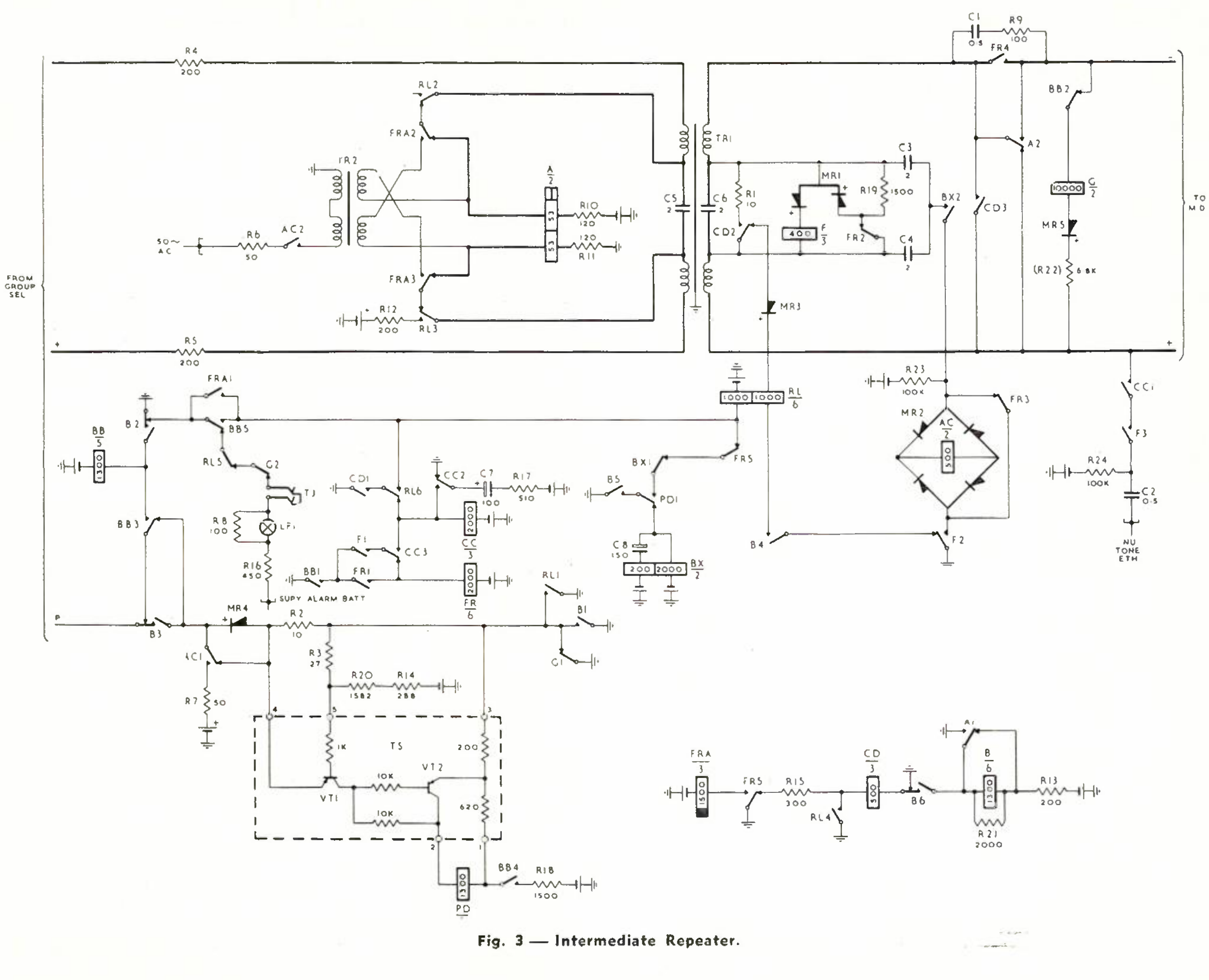

Fig. 3 — Intermediate Repeater.

If the first digit was a working group but the second digit is a barred level in that group, a discriminating relay will have operated after the first digit but contact units of that relay are not connected to the bank contact corresponding to the second digit. At the end of the second digit, relay CDR releases. As there is no holding circuit for relay NU via the CS5 bank, it releases and connects earth to the NU lead to operate relay FA in the repeater.

SUBSCRIBERS TRUNK SELECTOR

The subscribers trunk selector, Fig. 6, is provided so that the trial services have access to the 2VF trunk dialling network. The circuit has been designed so that it can be mounted in place of the Part I relay set of bothway 2VF Trunk Line circuit (Siemens Drawing XT9656). The Part II relay set is removed from the shelf.

The subscribers trunk selector is associated with the variable rate repeater and is controlled by the latter circuit. When the first trunk code digit is dialled, the selector searches for a free outlet in the required trunk transit group. When the second trunk code digit is dialled, impulses are repeated to the transit group trunk selector via a D.C. loop and, when the transit group trunk selector finds a free trunk line, a discriminatory signal causes the subscribers trunk selector to repeat all subsequent digits as VF pulses. During VF pulsing, a voltage limiter is connected across the line to suppress surges from the local side of the connection and, as a further safeguard, the connection is "split" during dialling.

Circuit Operation

Relays AA and BA in the subscriber's trunk selector operate when relay B (repeater) connects earth to the AA and BA leads, and relay BAR operates in turn. When the first trunk code digit is dialled, the DA magnet is pulsed via the DA lead and relay CDA operates via the CDA lead. When the digit switch moves off normal, earth is con-

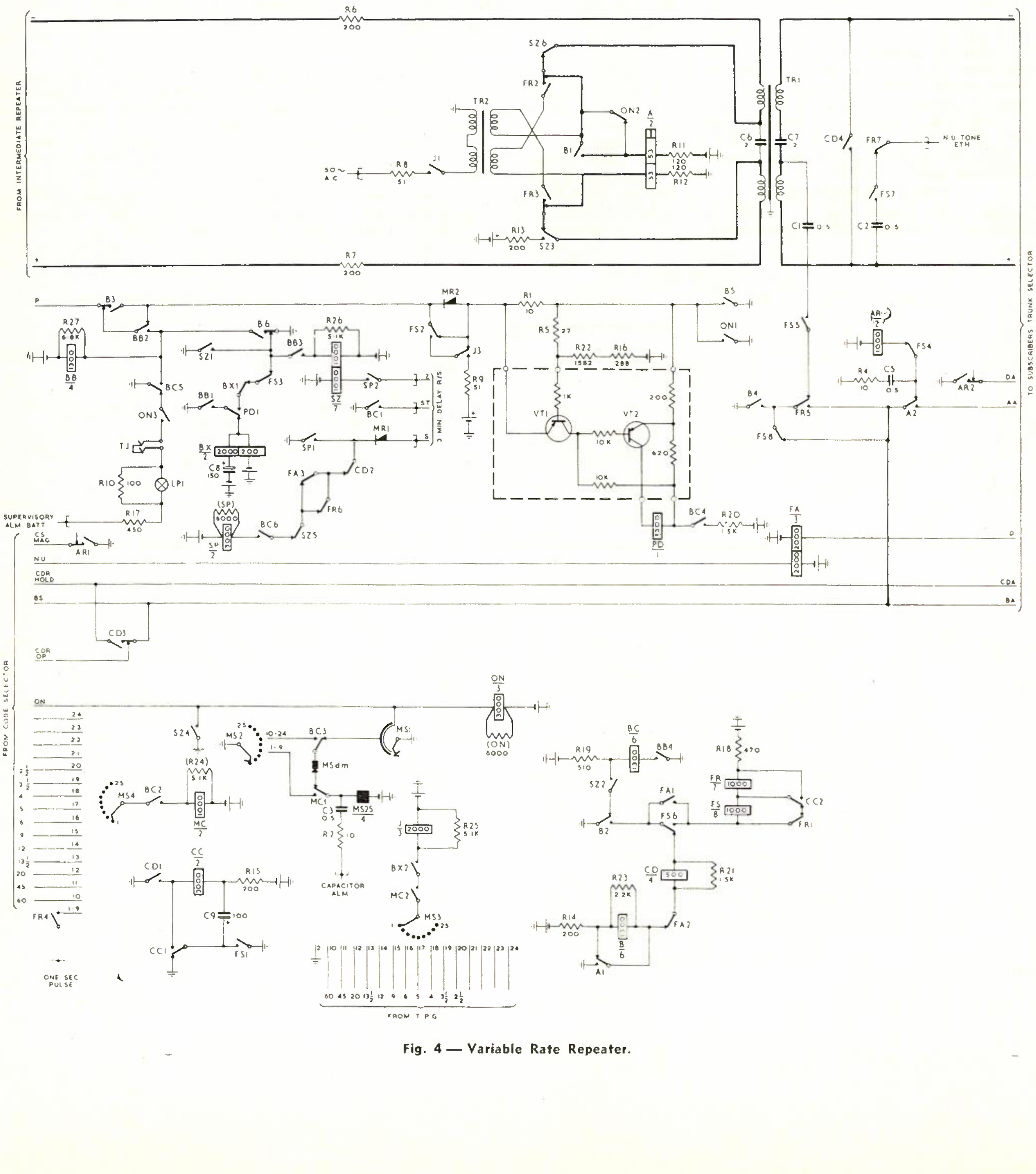

Fig. 4 — Variable Rate Repeater.

nected to the ON lead via the DA3 bank. Relay AA pulses but it has no effect at this stage. At the end of the digit, relay CDA releases and the motor uniselector searches for a free outlet in the marked group. When it moves off-normal, the ON springset connects earth to the ON lead. The motor uniselector drives until it reaches the required group which has been marked by means of connections between the DA4 bank of the digit switch and the M and M1 banks of the motor uniselector. When a free outlet is found, relay TA operates via earth, BA1, MR3, both windings of relay TA in series, DA4 wiper and bank, the connection to M or M1 bank, CDA5 normal, P wiper and bank, to 550 ohm resistance and negative battery in the transit group trunk selector. Relay TA disconnects the motor uniselector drive circuit and allows relay S to operate and lock. Relay S connects earth to the P wire to guard the connection and releases relay TA. Relay E is connected to the negative and positive wipers and operates in series with relay A in the transit group trunk selector.

When the second trunk code digit is dialled, relay AA pulses and relay CDA operates. Relay A in the transit group trunk selector is pulsed via a short-circuit loop by contact unit AA1, and relay E releases when it is short-circuited. At the end of the digit, relay CDA releases, relay E re-operates, and the 1500 ohm two-stage drop-back resistor is disconnected.

When the transit group trunk selector finds a free trunk line in the required group, relay VF in the subscribers trunk selector operates via the S lead in series with relay TT in the trunk line circuit, but relay TT does not operate due to the high resistance of relay VF. Relays VFR and X operate and lock to prepare the circuit for VF signalling. Relay E is released and relay VF releases when the operating path is disconnected at the trunk line relay set.

On subsequent digits, relay AA pulses and relay CDA operates. Relay DL operates, "splits" the connection, and connects a VF termination to each side. When relay AA pulses, "X" tone signals are transmitted to line during the break periods to operate the 2VF receiver at the distant trunk centre. At the end of each digit, relay CDA releases and relay DL releases after a delay of about three seconds. If digits are dialled with pauses of less than three seconds between digits, relay CDA will re-operate when each digit is dialled, relay DL will be held and the line will be "split" throughout dialling.

When the answer signal is received, relay D is operated by negative battery which is connected to the positive wire at the trunk line relay set. Earth is connected to the D lead, relay DD operates and locks, and relay VFR releases.

When the calling service replaces the receiver, the variable rate repeater commences to release and relays AA and BA in the subscribers trunk selector are released. Relays BAR, DD, X, and S release, and the motor uniselector and digit switch self-drive to their home positions and earth is disconnected from the ON lead.

If all outlets in the marked group are engaged, the motor uniselector will drive until it reaches contact 52 on the second half of the bank. Relay TA operates in series with resistor R7 via the M and P wipers, relay S operates, and busy tone is transmitted to the calling service.

If a receiver in the 2VF signalling system is subjected to a surge of sufficient amplitude, it will not respond to a true signal which follows immediately after the surge. The period required for recovery from a surge, that is, the blocking period, can be 70 milli-seconds. Therefore, when pulses in the local D.C. circuit are being converted to VF pulses, the surge voltages which are induced in the D.C. circuit must not be allowed to enter the VF signalling circuit.

With operator trunk dialling, this object can be achieved by "splitting" the connection when the operator's dial is moved off-normal preparatory to dialling a digit. The trunk line is thus isolated from the switchboard circuit and any surges induced in the local circuit during a digit will not interfere with the 2VF receivers.

With S.T.D., this procedure is not practicable because there is no indication that dialling is about to commence until the dial contacts break for the first impulse. A surge will develop in the local circuit and it will pass into the 2VF signalling system before splitting of the connection can be effected. The 2VF receiver at the distant end will be blocked and it will not respond to the early part of the first tone pulse. This pulse will, therefore, be clipped but subsequent pulses will not be affected because the connection will be split during the first break pulse.

Provision is made in the subscriber's trunk selector for:—

(i) the amplitude of the initial surge to be limited;

(ii) the line to be split soon after the commencement of each digit.

Surge suppression is obtained by means of 2/2A rectifier (MR1) connected across the line and the surge voltages which are developed in the local circuit are reduced in amplitude before passing to the trunk line circuit. This arrangement prevents blocking of the distant 2VF receiver and avoids clipping of the first impulse. The voltage limiter is disconnected when the answer signal is received.

When the outgoing trunk line is found, the subscribers trunk selector prepares to send VF pulses in place of D.C. pulses. Shortly after the first break pulse of the next digit is received, the line is split and both sides are connected to V.F. terminations for the duration of the digit. If the digits of the required number are dialled with interdigital pauses not exceeding three seconds, the line will be split until the complete number has been dialled. This arrangement has the advantage that, if the caller should speak while dialling a number, the voice frequencies cannot cause false operation of the distant 2VF receiver.

The release circuit of relay DL is arranged so that, if an answer signal should be received within three seconds of the end of the last digit, the capacitor is disconnected and relay DL will release quickly, that is, the speech path is connected immediately.

SAFEGUARDS AGAINST FALSE METERING

The experimental S.T.D. equipment incorporates several safeguards against false metering which could occur due to plant failure. Such safeguards are important in a multi-metering system, particularly where it is possible to make long-distance calls where the tariff pulse rates are high.

The circuits which are used in the trial equipment were originally developed by Mr. G. A. M. Hyde for earlier multi-metering equipment that has been installed in recent years at country centres near the capital cities in the Commonwealth.

Private Wire Detector Unit

Stop on Busy Condition: If a telephone service is engaged on a call, it is possible for another service to become cross-connected with the established conversation due to a fault in equipment, for example, dirty banks, wipers with low tension. If the cross-connection occurs between the line circuit and a repeater or final selector, the calling service and the intruding service will be held until both services replace their receivers.

If the cross-connection should occur between the line circuits and a multi-metering repeater, and metering was in progress, the meter pulses would be registered on the meters of both the calling service and the intruding service unless some preventive action is taken. Again, the connection will be held until both replace their receivers, and the meter pulses will be recorded on both meters until the connection is released by both services.

Although such cross-connections constitute only a very small percentage of the total calls attempted, they can occur due to unavoidable plant failure, and the private wire detector unit has been developed to prevent false metering in such cases.

By connecting a bridge network of resistors and a detecting circuit to the incoming P wire, it is possible to recognise the difference between a normal call and a cross-connected call. The detector unit is shown in Figs. 3 and 4. The operation is identical in both cases but the description will refer to Fig. 3.

The four "arms" of the bridge network consist of resistors R2, R3, R14 and R20, and the holding relays of the preceding equipment — the uniselector line circuit, group selector and selective digit absorption relay set. The detecting circuit consists of a transistor switch and a 3000-type relay. The switch was developed by Standard Telephones & Cables (A'sia) Pty. Ltd., Sydney, in conjunction with the Equipment Laboratory, and it is mounted in a case which has the same dimensions as a metal-cased, paper 2uF capacitor.

The bridge circuit has been re-drawn in schematic form (Fig. 7) for clarity of operation. The design is such that when the combined resistance of the holding relays is correct, the voltage across terminals 4 and 5 of the detector unit is approximately 0.2 volts. In this condition, transistor VT1 is switched on, transistor VT2 is biassed off, and there is no current in relay PD. When a cross-connection occurs, that is, in the stop-on-busy condition, the holding relays of the intruding equipment are also connected to the P wire, the overall resistance is reduced, and the voltage across terminals 4 and 5 falls to approximately 0.1 volt. Transistor VT1 is biassed off, transistor VT2 is switched on, and relay PD operates. The bridge network has been designed so that these voltage values will be obtained but, in selecting the resistance values, allowance has been made for a number of factors.

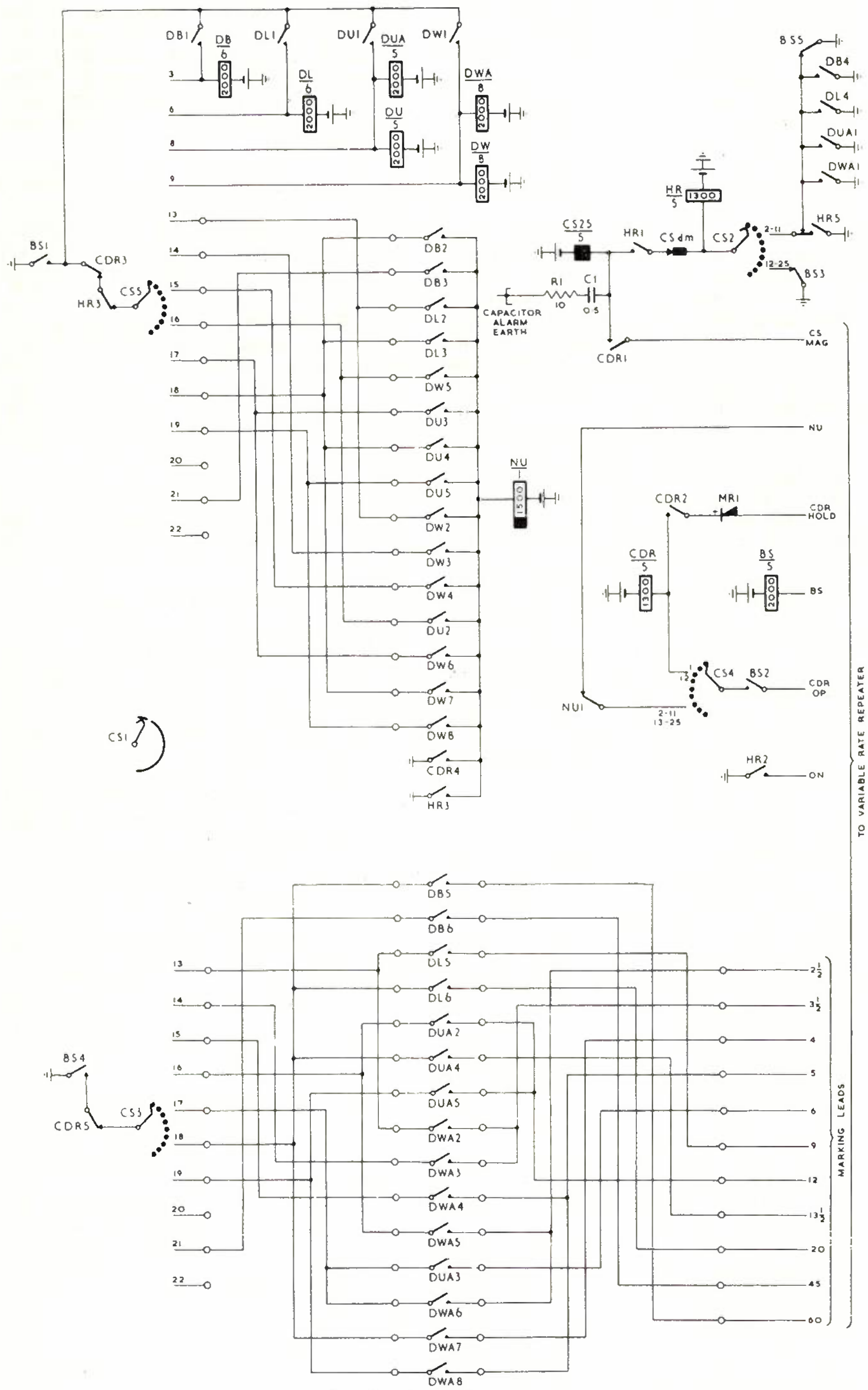

Fig. 5 — Code Selector.

Resistor R2 (10 ohms) is in series with the P wire and the resistance should be as low as possible to ensure that the guarding potential does not become unduly negative. However, the resistance must be sufficiently high to ensure that the voltage change from the normal to the stop-on-busy condition will operate the detecting circuit reliably in the worst conditions. The value chosen for resistor R2 is a compromise between the two requirements, and it is copper-wound so that a change in the resistance of the holding relays due to variation in ambient temperature will be compensated by a similar percentage change in resistor R2.

For economy, resistor R3 (27 ohms) is higher than resistor R2 so that the balance resistor will be of higher resistance than the holding relays and the current drain in the repeater circuit will be correspondingly lower.

With the values of resistance R2 and R3 fixed, the balance resistance is a function of the overall resistance of the holding relays which, in turn, is dependent on:—

(i) the trunking arrangement;
(ii) the type of equipment;
(iii) the manufacturing tolerances on resistance of components;
(iv) the heating effects of normal operation of the equipment.

(i) As the number of stages of selection which precede the repeater increases, the number of holding circuits in parallel on the P wire increases and the overall resistance decreases.

(ii) The resistance of holding circuits of various items of equipment are different, for example, the nominal resistance of the holding circuit of an SE50 type group selector is 2750 ohms whereas that of a 2000 type selector is 1550 ohms.

(iii) The usual manufacturing tolerance for resistance of wire-wound resistors and 3000 type relays is ± 5%.

(iv) The resistance of the relays and magnets increases during normal operation due to the heating effect of the current, for example, the 1550 ohm holding circuit of a 2000 type group selector can increase to 1700 ohms during prolonged operation.

Therefore, the balance resistor must be such that the detecting unit will not switch when the overall resistance of the holding relays on a normal call is at the minimum possible value; but it must permit the detector unit to switch when a maximum resistance holding circuit becomes cross-connected with a call in which the overall resistance is also at the maximum value.

As the number of selection stages increases, the minimum value of overall resistance for a normal call approaches the maximum value of overall resistance of a cross-connected call. Eventually a point is reached where there is insufficient difference between the two conditions to effect the required voltage change at the input to the detecting unit.

A single balance resistor cannot cover a trunking arrangement in which a rank of group selectors consists of different types of selectors which have markedly different resistances in the holding condition, for example, 2000 type and SE50 type, and it is necessary that all selectors in any one stage of selection should be of the same type.

The balance resistors in the trial equipment have been designed to cater for a trunking scheme which includes the modified line switch, one group selector, and a selective digit absorption relay set. The first selectors at Russell exchange are 2000 type and resistor R20 only is used, that is, resistor R14 is strapped. Provision was made for SE50 type group selectors by including resistor R14. If the group selectors were SE50 type, resistors R20 and R14 would be connected in series.

Finally, to ensure that variation in the bridge resistors do not unduly affect the circuit limits, resistors R2, R3, R14 and R20 have a tolerance of $\pm$ 1%.

Reverting to the circuit operation, when relay PD operates, it releases relay BX, which has a release time of 1.5 to 3.0 seconds. If an answer signal has been received before the cross-connection occurs, that is, if relay FR is operated, contact unit BX2 disconnects the circuit of relay AC. This prevents metering pulses being connected to the P wire while the stop-on-busy condition persists. If the call has not been answered, relay SZ operates when relay BX releases, and forced release action is initiated.

Relay BX is incorporated in the circuit to cover a condition which occurs when a 200 outlet group selector switches to a free outlet and the alternative outlet is engaged on a multi-metered call. A holding relay of the group selector remains connected to the P wire of the alternative outlet until relay B in the selector releases. This simulates a stop-on-busy condition and relay PD in the repeater will be operated until relay B releases. Relay BX, being slow to release, covers this period. The circuit has been designed so that capacitor C8 will become practically fully charged in approximately 20 milliseconds when relay PD is released. Relay BX will thus hold even if the above condition should occur in rapid succession, for example, on an early choice where the outlet is open circuit

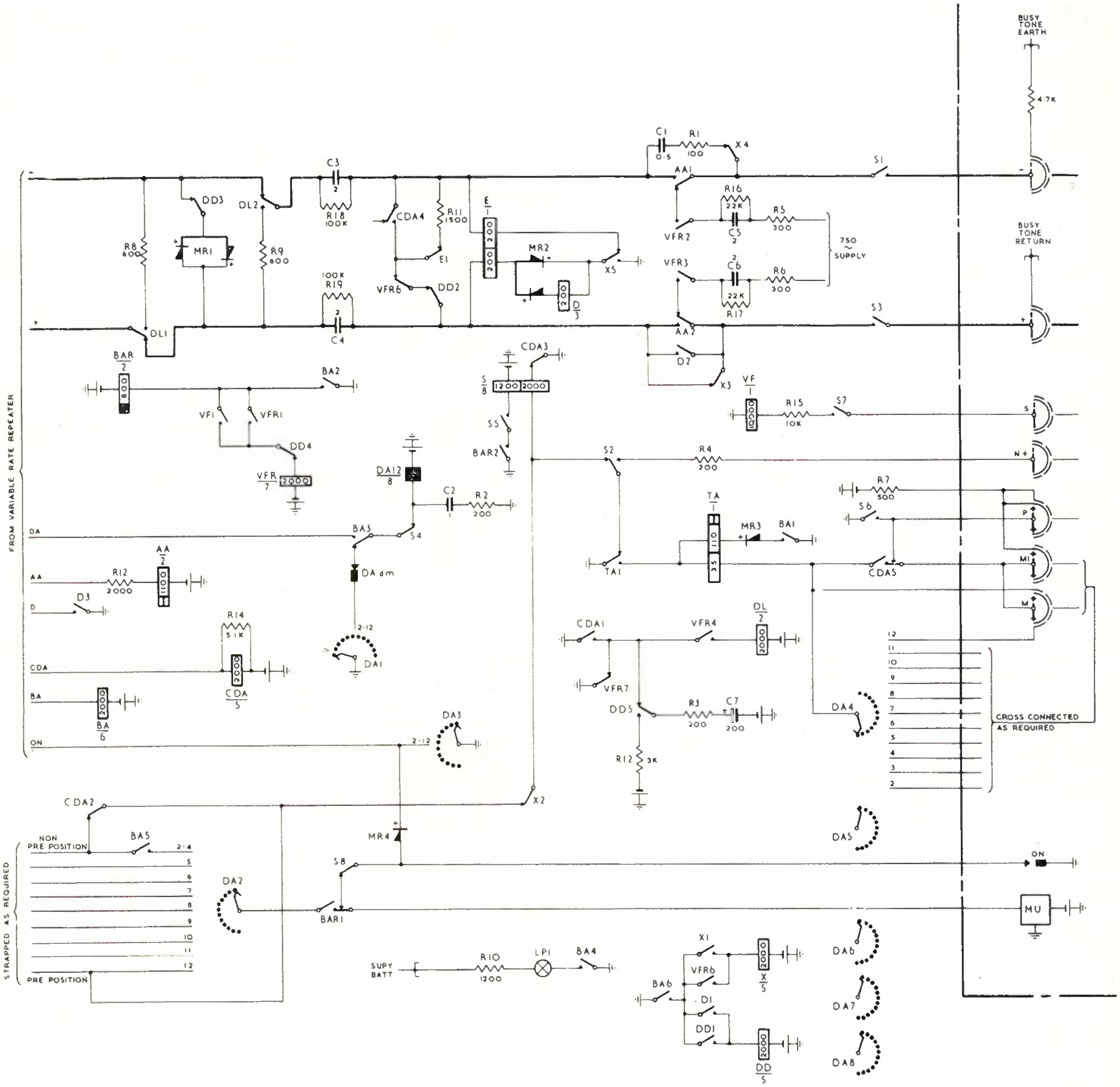

Fig. 6— Subscribers Trunk Selector.

and causes a "drop-out" or "step-off-open trunk" condition.

Discrimination Against Barred Services: The detector unit is also used to prevent barred services from obtaining access to the S.T.D. equipment. The balance resistors are such that a call which is established from a uniselector line circuit with a 2600 ohm K relay will not cause the detector to operate. However, a 1300 ohm K relay will cause operation of relay PD and prevent the barred service from obtaining access.

Detection of Reversed Lines

If a reversed line, for example, reversed jumper, is encountered when a unit fee call is being established, the standard auto-auto repeater responds to the reversed potentials as it would to an answer signal, but one meter pulse only is recorded on the calling subscriber's meter throughout the call.

However, if multi-metering equipment responded to the false answer signal in similar fashion, periodic meter pulses would be recorded after the "free" period elapsed. These pulses would continue until the calling subscriber replaced the receiver — even though the caller took an appreciable time to answer or no answer was received — and a considerable number of pulses could be recorded without conversation actually taking place.

To guard against false metering of this nature, relay CC is included in the repeater circuits. It is a slow release relay due to the action of the associated capacitor, and does not release until 0.75 to 1.5 seconds after a digit has been dialled. If a reversed line is encountered at the end of the digit, relay F will operate while relay CC is slowly releasing. Relay CC locks and prevents the operation of relay FR, and N.U. tone is transmitted to the calling service. Meter pulses will not be registered because relay FR has not operated.

There is the possibility that, if a subscriber answers a call within 1.5 seconds, the caller can receive N.U. tone, but the probability is low — particularly when the random connection of the silent period in the ringing current cycle is considered. The release time of relay CC cannot be less than 750 milliseconds for it must cover the release time of the end-of-digit relay (CD) and the searching time of a group selector over a level and the overall time for these actions can be about 600 milliseconds.

Reversed Junction Guard

The junction guard feature which was described earlier is a further aid in preventing false metering. If the junction pair should become reversed, the intermediate repeater is immediately guarded and cannot be taken into use.

METERING ON TWO-WIRE JUNCTIONS

For the purposes of the experimental S.T.D. equipment, the pulses are transmitted as signals of 50 c.p.s. A.C. superimposed on the junction pair. The input voltage to the metering transformer is 50 volts R.M.S., which is derived from a step-down transformer connected to the commercial A.C. supply. The meter signals are almost inaudible during silent periods and have no effect on conversation.

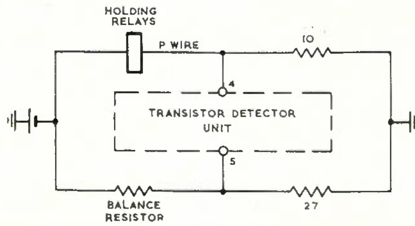

Fig. 7 — Private Wire Detector Unit Schematic Circuit.

The superimposed A.C. signals have been used for the trial as it is possible that such signals will be used to operate check meters at subscribers' premises. When S.T.D. is introduced on a wide scale, such meters will be required at a small number of establishments such as hotels where immediate knowledge of trunk call charges must be available so that the call can be debited correctly.

As the battery feed to the subscriber's telephone cannot be interrupted with the present form of instrument, it is necessary to used superimposed signals. The use of A.C. signals on the junctions allows the multi-metering repeaters to transmit meter signals on both junctions and subscribers' lines, depending on whether the calls have been received from another exchange or a local service.

DEMONSTRATION MODEL

A demonstration model of the experimental S.T.D. equipment was prepared by the Laboratory for a display which was held in Hobart, Tasmania, in 1959, in conjunction with the Annual Conference of the Institution of Engineers, Australia. The display was also open to the public for several days.

The equipment was installed in the display hall by staff of the Tasmanian Administration with the supervision of a Laboratory officer, Mr. D. Thomas, who also demonstrated the model. A large trunking diagram with lamps at appropriate places illustrated the progress of a call and another lamp gave visual indication of the meter pulses. Provision was made for simulation of unstandard conditions so that the safeguards against false metering could be demonstrated.

CONCLUSION

The trial equipment has been in service for over twelve months and it has provided a useful guide to the conditions which subscribers will meet when S.T.D. facilities are extended to the trunk dialling network and to the reliability of a method of transmitting meter pulses.

However, it does not represent the final thoughts on the design of S.T.D. facilities for a nation-wide system. Equipment which will be required for the latter purpose will be more flexible to meet the requirements of the wider application.

ACKNOWLEDGMENTS

The S.T.D. equipment which is described in the article is based largely on a previous S.T.D. design by Mr. G. A. M. Hyde. Detailed design of the trial equipment was originally undertaken by Mr. D. C. Sowden, and supervision of testing and installation was performed by Messrs D. K. Dunn and L. W. Watkins. Acknowledgment is accorded to the efforts of the officers of the Victorian Administration who arranged for connection of the trial equipment to plant in situ. The author gratefully acknowledges the assistance of the Laboratory staff in the preparation of the article.

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SWITCHBOARD ATTACHMENTS FOR BLIND TELEPHONISTS AND AUDIBLE CORD SUPERVISION

N. A. CAMERON and M. W. FARMER, B.Sc.

INTRODUCTION

Following a representation by the Royal Victorian Institute for the Blind to the Director-General, attachments have been designed and installed which enable blind telephonists to operate efficiently P.M.B.X. (1) and P.A.B.X. lamp signalling switchboards. This article gives details of the components designed in the Research Section of Central Office and also outlines the methods used by Melbourne staff to apply these devices while maintaining satisfactory installation and maintenance practices.

Existing Facilities — An article in the Australian Post Office Magazine (2)

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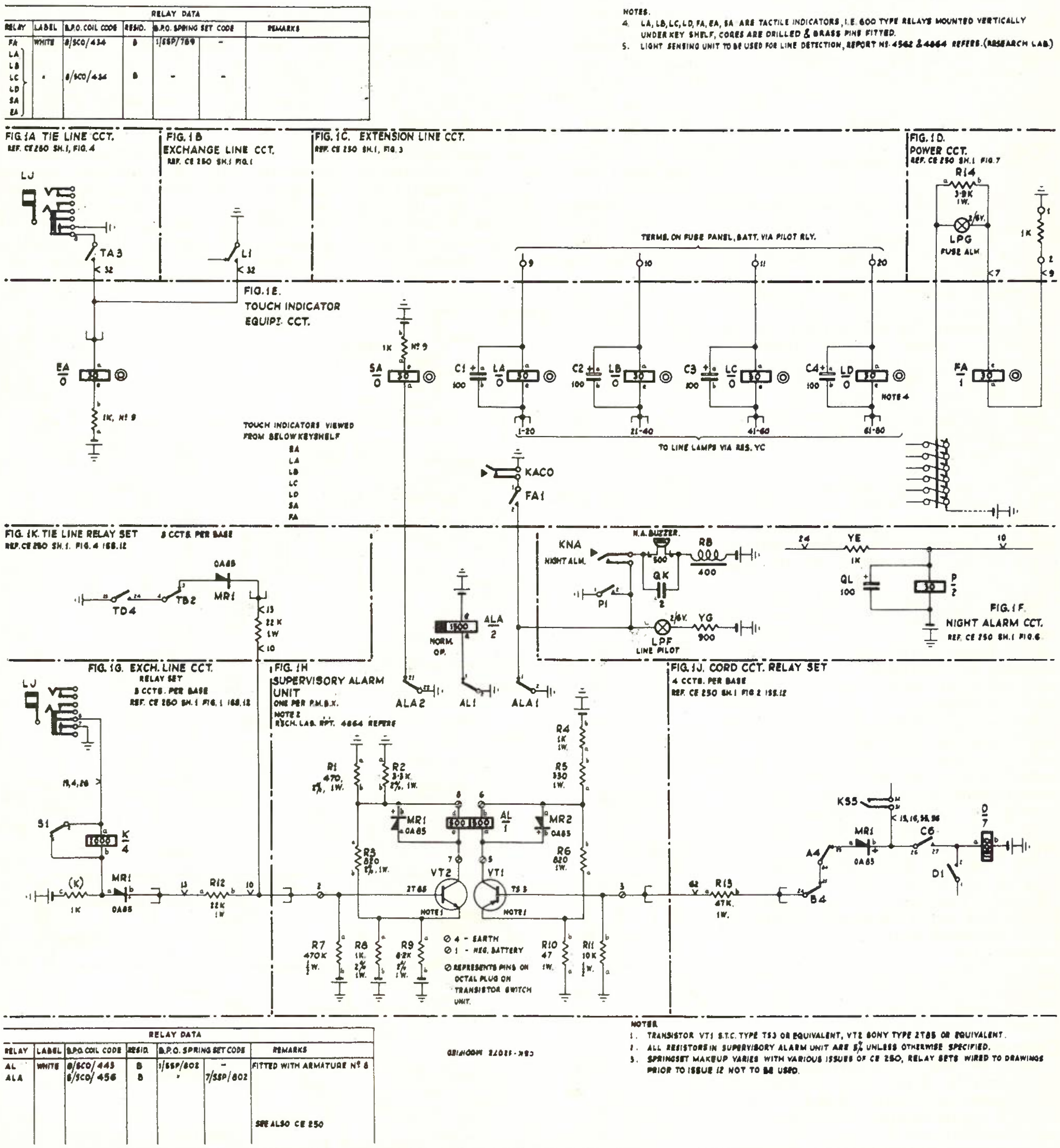

Fig. 1 — Circuit showing modifications to provide Blind Operators' Facilities, including Audible Cord Supervision.

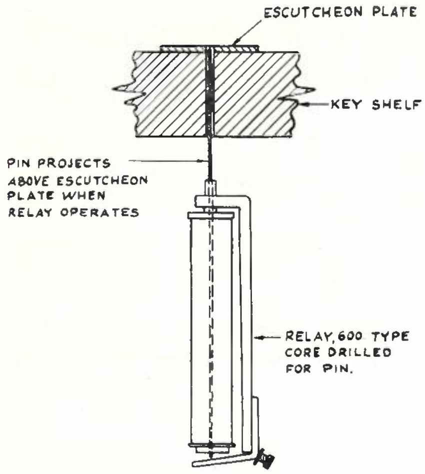

Fig. 2 — Principle of Touch Indicator.

describes what the Department has done in the past to assist blind telephonists. Most of these telephonists work on eyeball indicator type switchboards and very little modification to standard equipment is needed beyond removing the glass covers which are normally over the indicators. Some partially blind telephonists work on lamp signalling switchboards but up to the present these have been P.A.B.X. boards for two main reasons:—

(i) They have a minimum of lamps compared with P.M.B.X. switchboards, and
(ii) The circuitry is more adaptable to giving an audible alarm on supervisory signals.

Drawings TD667 and WD1229 are typical examples of the lamp signalling P.A.B.X. boards modified by the Department. Metal studs and a valve oscillator are provided, and when the telephonist bridges an appropriate stud on the keyshelf to a metal sensing bar, the oscillator produces a tone in the headset. A metal stud is provided for each incoming line and others for supervisory purposes.

Siemens and Halske have developed a small solenoid type indicator (called a tactile indicator) to fit into the lamp socket in place of the ordinary switchboard lamp (3). This enables the telephonist to locate the calling line by touch. Field trials in Melbourne with these indicators have resulted in some failures due to clogging by dust, and bending of the nylon extension of the solenoid. Their use has not been extended. The Bell Telephone Laboratories have produced a transistorised photoelectric audio device consisting of a photo-transistor mounted on the finger which, on passing a glowing lamp, produces a tone from a transistor oscillator in the telephonist's headset (4).

Proposed Facilities — It was decided to aim at a set of attachments which would enable totally blind telephonists to operate efficiently P.M.B.X. and P.A.B.X. switchboards. This approach simultaneously covered the case of the partially blind telephonist. The P.M.B.X. represents the more difficult problem, and the following comments refer to that equipment. Reference to P.A.B.X. application is contained towards the end of the article. In designing for a totally blind operator, it is necessary to provide:—

(a) an **audible alarm** for every condition requiring the attention of the operator;
(b) a means of **sectionalising** the switchboard to facilitate speedy location of the point requiring attention, and
(c) a means of **locating** glowing lamps.

The methods used to achieve the facilities required are described in the following paragraphs.

AUDIBLE ALARMS

The lamp signalling P.M.B.X. to Drawing CE-250 gives an audible alarm on all conditions except cord supervision and some fuse alarms. In the case of cord circuit supervision a study of the circuits concerned shows that there is no readily applicable single modification to give a call-terminated alarm for both extension-extension and exchange-extension calls. However, by using transistors to detect voltage changes that occur during the termination of a call, a satisfactory solution has been obtained (5).

Cord Circuit Modification. Modifications to the cord circuit relay base facilitate an alarm to be given on the termination of extension to extension calls. Reference to Fig. 1J shows how the voltage occurring at the junction between KS5 and contact C6 is fed through an added germanium diode, MR1, and added contacts A4 and B4, to a commoning point, U-jack 62, in the relay base. The other three circuits in the base have similar modifications brought out to U-jack 62. The voltage variations appearing at U-jack 62 are coupled through an added 47K ohm resistor to the transistor switch (Fig. 1H).

At the initiation of an extension-extension call, the KS5-C6 junction achieves a negative potential sufficient to operate a PNP transistor switch. However, the operation of A4 and B4 contacts will prevent this happening until both extensions have hung up. The

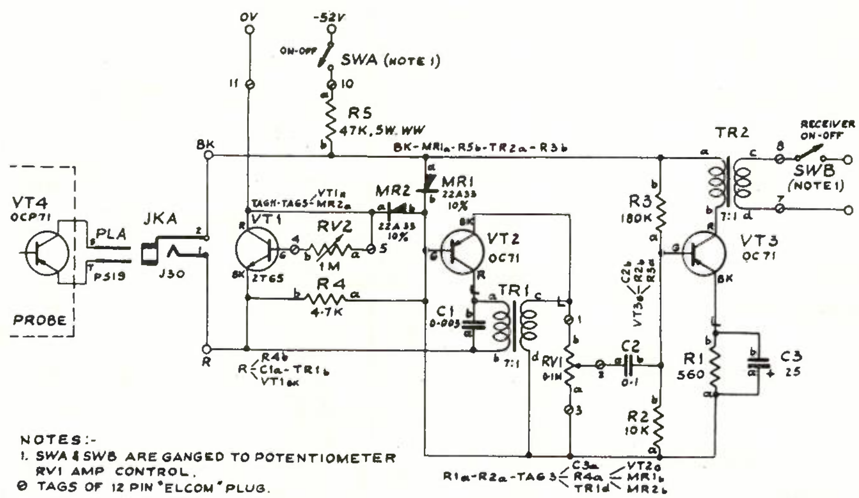

Fig. 3 — Circuit of Probe Unit.

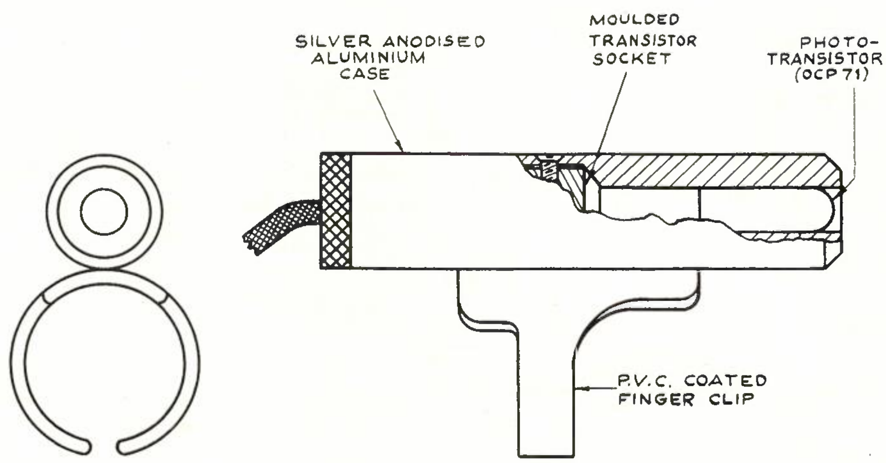

Fig. 4 — Transistorised Probe.

four germanium point contact diodes prevent interaction between cord circuits, and the 47K ohm resistor limits the take-off current to approximately 1 mA. When the cord circuit is idle, or during an exchange-extension call, the take-off point is either floating or at ground potential; neither of these conditions will operate the transistor switch.

Exchange Line Circuit Modification—To provide an alarm on the termination of exchange-extension calls, modifications are made to the exchange line circuit relay base and these are shown in Fig. 1G. The pick off point is the junction of a 1K ohm resistor and the winding of relay K. This point is at a potential of -50 volts before a call and approximately -33 volts during a call. However, at the termination of the call, relay S releases and contact S1 puts earth on the pick-off point. This change of potential is fed, as in the cord circuit modification, through an added isolating diode, MR1, and an added 22K ohm resistor to an NPN transistor switch.

Tie Line Circuit Modification—Alarm conditions on tie line calls are obtained in a similar manner to that described for the exchange line circuit, but the addition of a springset to each of the relays TB and TD is necessary (Fig. 1K).

Supervisory Unit—The circuit of the supervisory unit is shown in Fig. 1H. It consists of the NPN and PNP transistor amplifier switches previously mentioned, operating a double wound relay, AL. Contact AL1 releases a normally operated slow release relay ALA, contact ALA1 of which operates the night alarm buzzer. Contact ALA2 operates a special touch indicator SA, to be described later. Each transistor switch has been designed to be thermally stable whether "on" or "off" Type OA85 diodes have been placed across the relay winding to absorb any transient voltage peaks which might cause transistor damage. Relay ALA has been made slow release because on an outgoing extension-exchange call, contact S1 (Fig. 1G) releases briefly during dialling impulses, and this could cause unnecessary operation of the supervisory alarm.

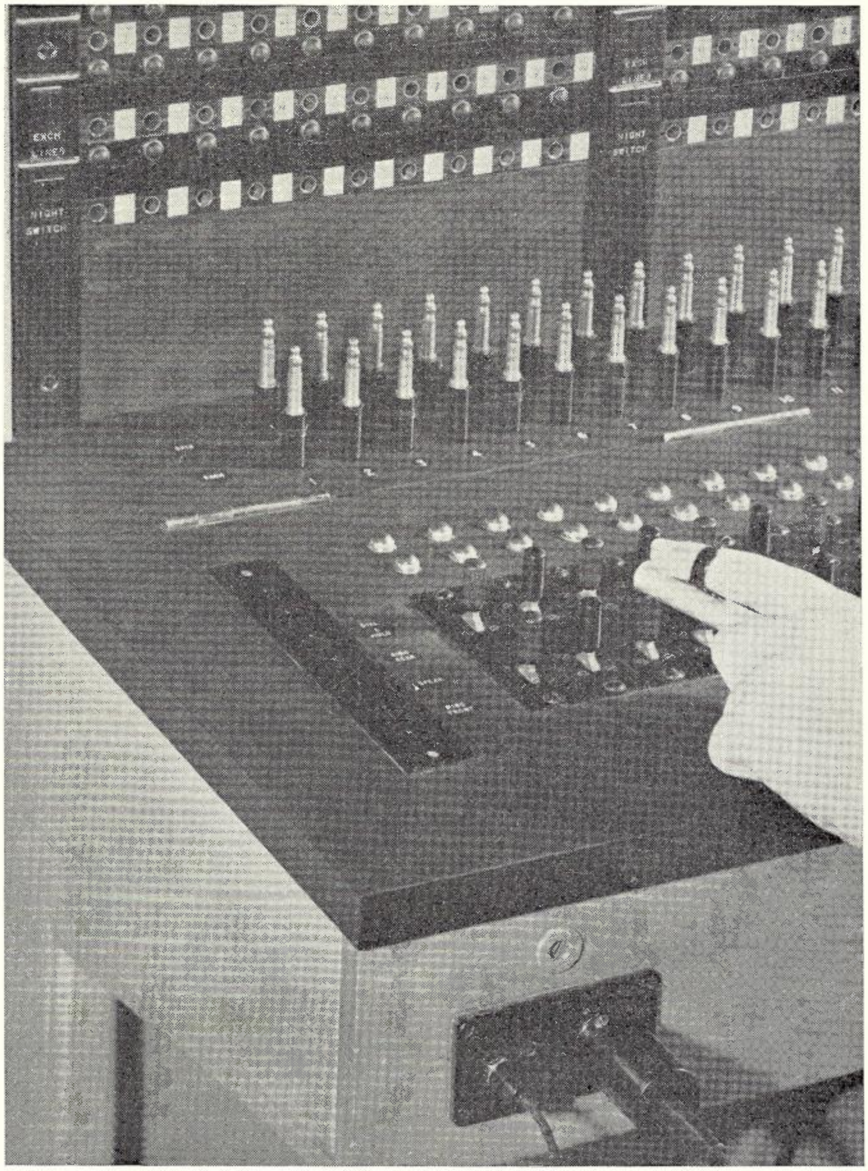

Fig. 5 — Face of Switchboard showing additions for Blind Operators. Touch Indicators (left of Key Shelf); Probe on Finger; Probe Plug and Jack (adjacent to concentric plug); Smooth and Ribbed Key Handles and Plug Covers.

SECTIONALISATION OF SWITCHBOARD

As well as receiving an audibie signal, it is necessary for the operator to be made aware of what portion of the board requires attention. For practical purposes the attention required for a P.M.B.X. can be divided into seven groups:—

(i) Extension lamps 1 - 20
(ii) Extension lamps 21 - 40
(iii) Extension lamps 41 - 60
(iv) Extension lamps 61 - 80
(v) Exchange and Tie Lines
(vi) Supervision
(vii) Fuse Alarm

For each of the above groups, a solenoid type touch indicator has been provided, consisting of a modified 600 type relay mounted to one side of the keyshelf. Fig. 2 shows physical details of the indicator and Fig. 1E the circuitry involved. For example, consider a call originated by extension 45. The extension line circuit operates as normal and contact P1 of the pilot relay (Fig. 1F) brings in the audible alarm. Touch indicator LC (Fig. 1E) operates in series with the extension circuit and the pilot relay, and the indicator pin associated with extension 41-60 projects above the surface of the keyshelf. On hearing the audible alarm, the operator feels for the projecting pin and then proceeds to search over the extension field in the 41-60 area. Similar procedures apply in the other six groups listed.

In the case of some fuse alarms, the standard P.M.B.X. circuit provides a visual alarm but no audible alarm. To provide an audible alarm for all fuse failures, a tactile indicator, FA in Fig. 1E, has been placed in series with the fuse alarm lamp. FA has one make springset which, when operated, brings in the audible alarm. An alarm cut-off key, KACO, has been provided for those fuse alarms which do not operate the pilot relay; it is impracticable to provide a similar facility for those fuse alarms which do operate the pilot relay.

LOCATION OF GLOWING LAMPS

Having received an audible alarm, and determined from the touch indicator the section of the board requiring attention, the blind operator must then locate the glowing lamp. The first unit developed for this purpose consisted of a photo-voltaic probe using a silicon solar cell, controlling a transistor oscillator coupled to the telephonist's headset (6). When the probe was brought near the glowing lamp a tone was heard in the headset, the volume of the tone being an indication of the closeness of the probe to the lamp. Because of the size of the silicon cell it was necessary to use a probe (a spun aluminium cone) of 1½ inches diameter and about 3 inches long. The size of the probe was a definite handicap and a further disadvantage was the cell's marked bias towards light through red lamp caps. When photo-transistors became commercially available in this country, a new probe was designed 2¼ inches long and ½ inch in diameter (7). This enables the probe to be mounted on the operator's finger.

Light Sensing Probe and Oscillator Unit—Fig. 3 gives the schematic circuit of the unit designed to locate the glowing lamp. The probe contains a PNP photo transistor type OCP71 working in conjunction with an NPN transistor type 2T65 as a thermal balance transistor. The two emitter leads are coupled together through a common 4.7K ohm load resistor. The thermal currents of the two transistors will be in opposition through the load, and there will be no voltage developed across the load until the photo-transistor is illuminated. A one megohm potentiometer is connected in the base circuit of the NPN transistor as a sensitivity control. This will permit adjustment to compensate for ambient illumination on the photo transistor, and the slight mismatches between the characteristics and ambient temperatures of the two transistors. This control requires only infrequent manipulation.

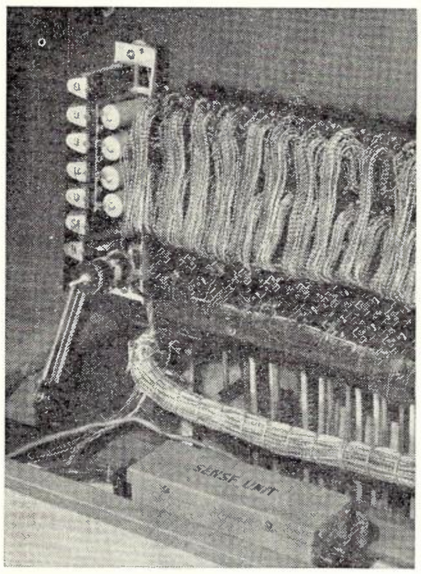

Fig. 6 — Keyself raised to show addition of seven Touch Indicator relays, and the Sense (or Probe) Unit.

The voltage drop across the 4.7K ohm load resistor is used to power a transistor oscillator with a frequency of approximately 1 Kc/s. The oscillator will operate with a voltage supply of about 0.05 volts or higher and its output is coupled to the telephonist's receiver circuit via a volume control and transistor audio amplifier. Two 3.3 volt zener reference diodes provide supply voltages from the switchboard battery.

APPLICATION IN THE FIELD

This article so far has dealt mainly with the circuit devices which have been developed to meet the special needs of blind operators. The methods used to equip a P.M.B.X. switchboard with the special aids are now described. The items of equipment to be added to the switchboard are summarised as follows:

(i) Probe, Probe Unit, Probe Controls;
(ii) Supervisory Unit, Resistors and Diodes;
(iii) Touch Indicators;
(iv) Relays AL, ALA;
(v) Other Aids.

Probe, Probe Unit, Probe Controls: The probe containing the light-sensitive photo-transistor was designed and manufactured in conjunction with Melbourne Postal Workshops. The probe was required to satisfy the following conditions:—

(a) Light weight;
(b) Minimum size;
(c) Durable;
(d) Protective to the transistor;
(e) Comfortable to the user;
(f) Adaptable for different sized fingers;
(g) Cord connection to facilitate maintenance and minimise possibility of heat damage to the transistor from a soldering iron.

A design was evolved which satisfactorily embodied these requirements. Fig. 4 gives details of the probe. Fig. 5 shows the probe connected to a switchboard. Normally, the probe is attached to a finger of the left hand.

The probe or sense unit circuit, as in

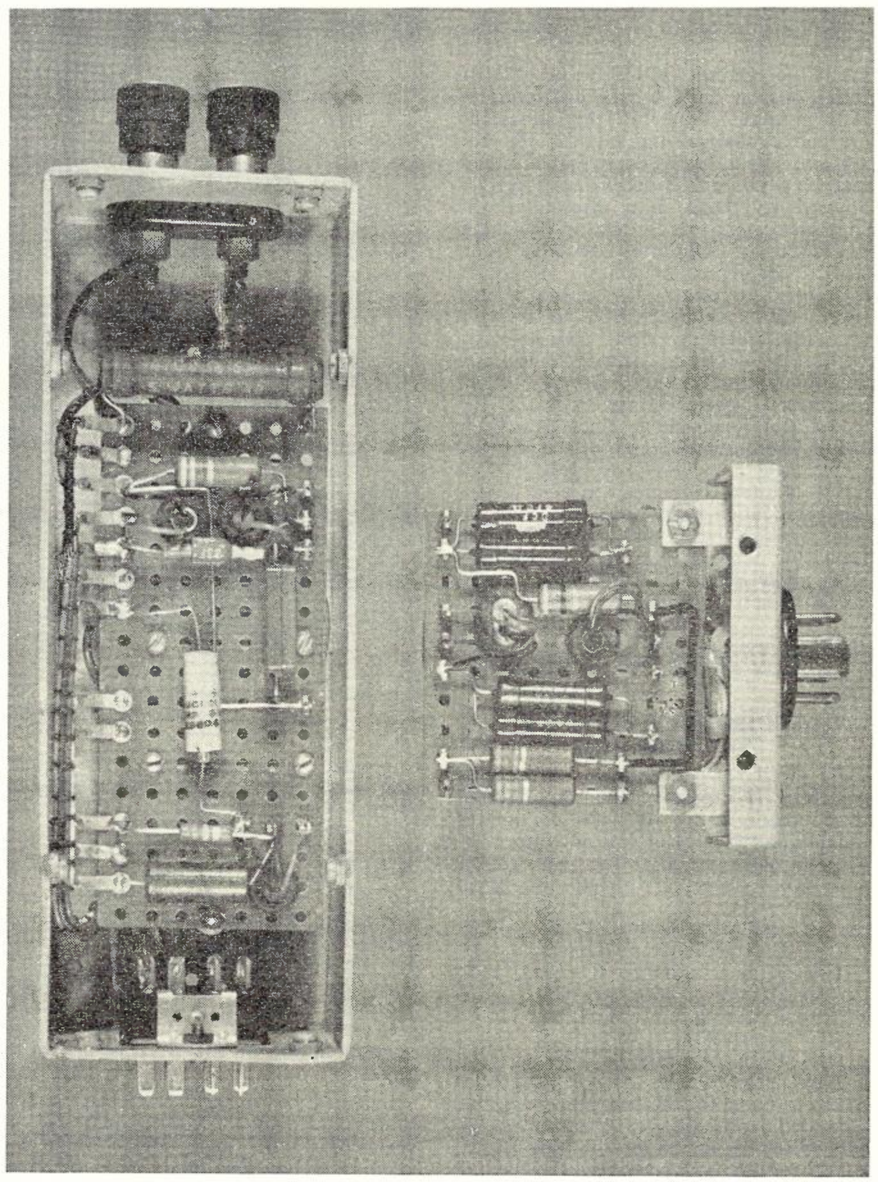

Fig. 7 — Covers removed from — (left) Probe Unit; (right) Supervisory Unit.

Fig. 3, was built into a metal box 7 x 2½ x 2 inches. The unit was mounted under the keyshelf, Fig. 6, and a wire form provided to interconnect the probe jack, the probe unit, and the probe controls, and also to provide battery for the probe unit. An ELCOM type plug was built into the unit and the associated socket was fixed in the switchboard under the keyshelf. It is the intention that maintenance on the unit will not be carried out on site, but that a replacement unit will be inserted and the family unit removed for repair. This approach is based upon three factors:—

(a) The units will be relatively few and normal maintenance staff will not be familiar with them;

(b) The components in the unit are not yet in common use in sub-station plant and any repairs would be best carried out in some location where suitably trained staff is available;

(c) Notwithstanding (a) and (b), a faulty unit must be replaced forthwith, as any delay would seriously affect the satisfactory operation of the switchboard, and also cause distress to the operator.

Fig. 7 shows the probe unit with cover removed.

The probe controls are illustrated in Fig. 8, which shows the volume and sensitivity controls positioned above the jack field. The operator needs to make only occasional adjustments to these controls. The volume control includes an on-off switch for the probe unit.

Supervisory Unit, Resistors and Diodes: The supervisory unit (or transistor switch), circuit as in Fig. 1H, was built into a metal box 3 x 3 x 1½ inches. Relay AL was not included. An octal plug was built into the unit and the associated socket was mounted in a relay mounting type F151/5AV. The relay mounting was required also for the AL and ALA relays. A wire form was provided to make the necessary connections to the components mounted on the relay mounting. Fig. 9 shows the relay mounting with the cover removed. Fig. 10 shows the location of the added relay mounting in the switchboard. As with the probe unit, it is the intention to replace any faulty supervisory unit and carry out any necessary repairs away from the location of the switchboard. The resistors and diodes added to the exchange line, cord, and tie line circuits were simply and satisfactorily mounted in the appropriate relay bases.

Touch Indicators: Fig. 2 shows the make-up of the touch indicators in use in Melbourne. Figs. 5 and 6 illustrate the indicators fitted to a switchboard. The black perspex escutcheon plate around the indicator pins, Fig. 5, was provided to facilitate location of the indicator area by the operator, and also to provide some protection for the pins. The functions allocated to the particular indicators were so chosen as to follow the layout of the appropriate components on the switchboard. Referring to Fig. 5, and commencing nearest the operator, the indicators have the following functions respectively: Exchange and Tie Lines, Extensions 1-20, Extensions 21-40, Extensions 41-60, Extensions 61-80, Cord Supervision, Fuse Alarm.

Relay AL, ALA: There was no suitable ready location in the switchboard for these relays. Therefore, a relay mounting type F151/5AV was installed in the switchboard and the relays and supervisory unit installed in it. See Figs. 9 and 10.

Other Aids: Outlined above the the major aids to facilitate blind telephonists satisfactorily operating switchboards. However, there are some other aids which, while relatively minor, nevertheless constitute a real benefit to the blind operator. These are discussed briefly now.

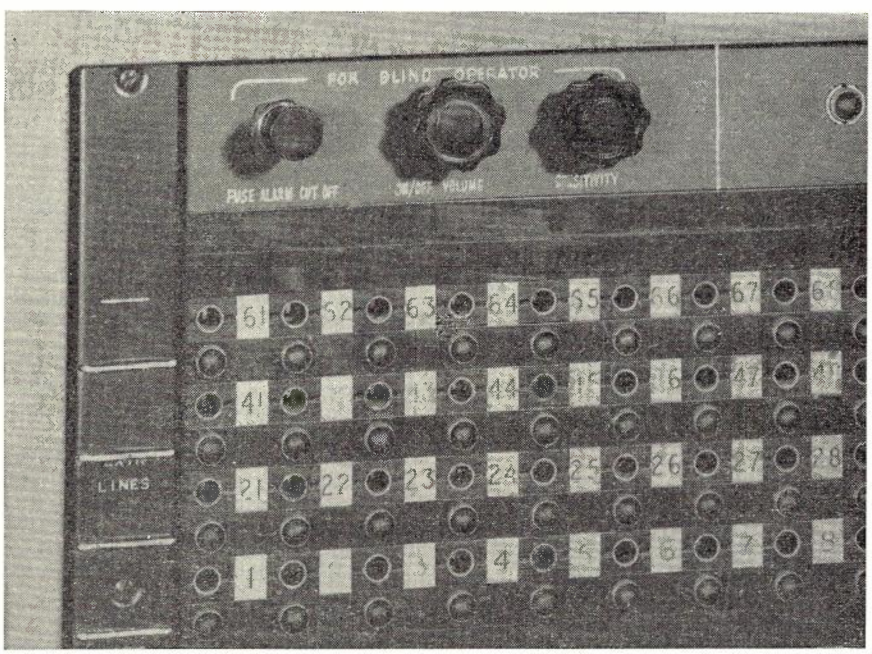

Fig. 8 — Controls for Blind Operator added above jack field. Note ribbed stile strips to facilitate location of row.

Key handles are provided alternately smooth and serrated, Fig. 5, and associated with the appropriate plug covers which are provided alternately smooth and ringed. The ringing is done with a plastic sleeve, as shown in Fig. 5. In the knowledge that a smooth key handle is associated with a smooth plug cover, and that a serrated key handle is

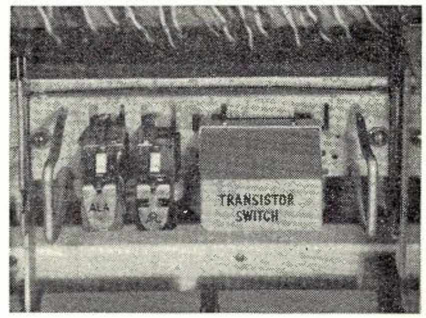

Fig. 9 — Relay Mounting added in Switchboard. Note plug-in Transistor Switch on Supervisory Unit.

associated with a ringed plug cover, the blind operator is greatly assisted in speedy operation of the switchboard.

To assist the blind operator in speedy location of the required lamp and jack, the jack field was sectionalised by ribbing the stile strips in line with each row of jacks (and lamps), as illustrated in Fig. 8. If necessary, the jack field can be further divided by the addition of a central vertical notched strip.

It is worthy of note that some lamp caps have poor translucent properties. Such lamp caps should be replaced on switchboards fitted with the probe sensing device.

Braille number rings are standard items for addition to dials. In practice, blind operators generally do not require this aid.

As detailed above, on a switchboard operated by a blind telephonist, the night alarm buzzer must be switched on at all times. The tone of the standard buzzer, while satisfactory for its intended purpose, could be considered unsatisfactory for constant use. Provision of a more pleasantly toned buzzer is being investigated for use on switchboards modified for blind telephonists.

Layout of Components: The components as described and illustrated have been so located on the switchboard as to reduce to a minimum the time taken to establish any connection.

In Melbourne, a subscriber who employs a blind telephonist has agreed to his modified switchboard being used for the training of blind telephonists. In this way, optimum operating procedures will be taught to all blind telephonists. Cases can arise where other physical disabilities could require a lay out different to that described.

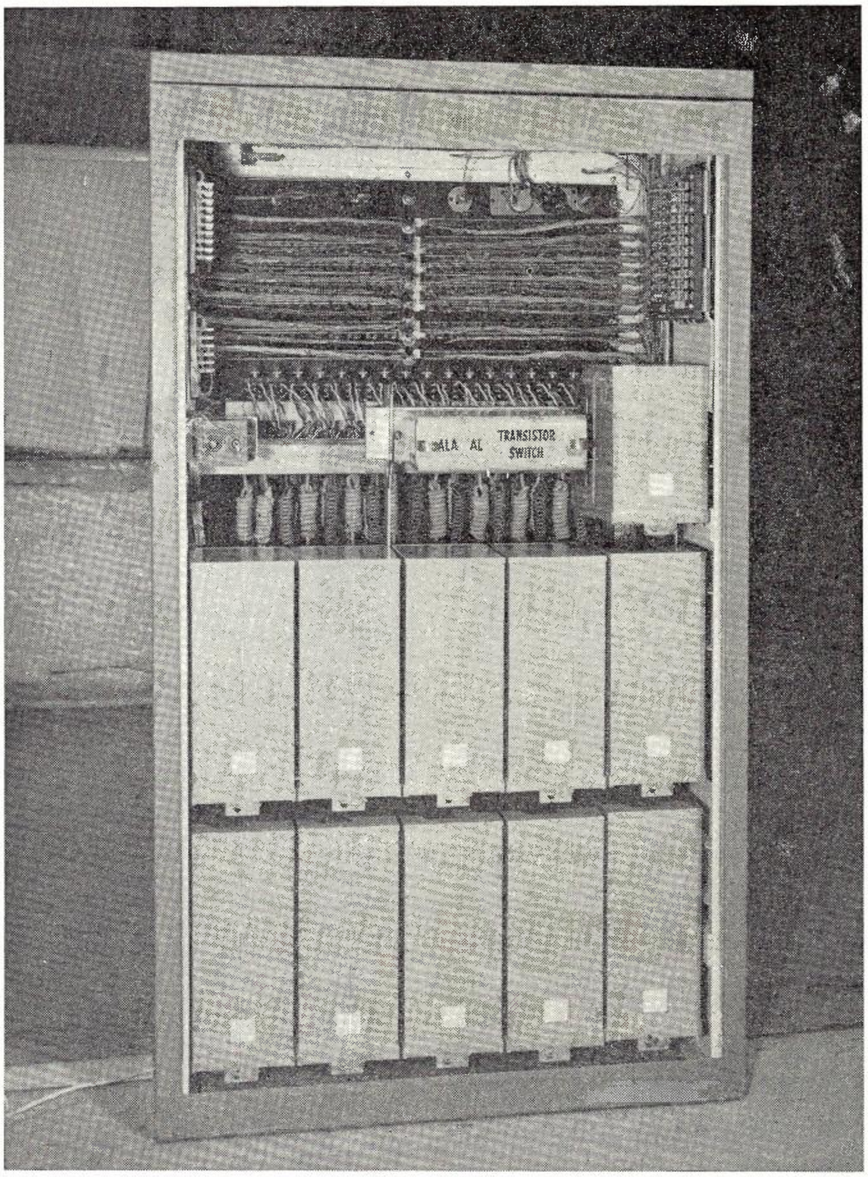

Fig. 10 — Rear of Switchboard showing added relay mounting for relays and Transistor Switch (or Supervisory Unit).

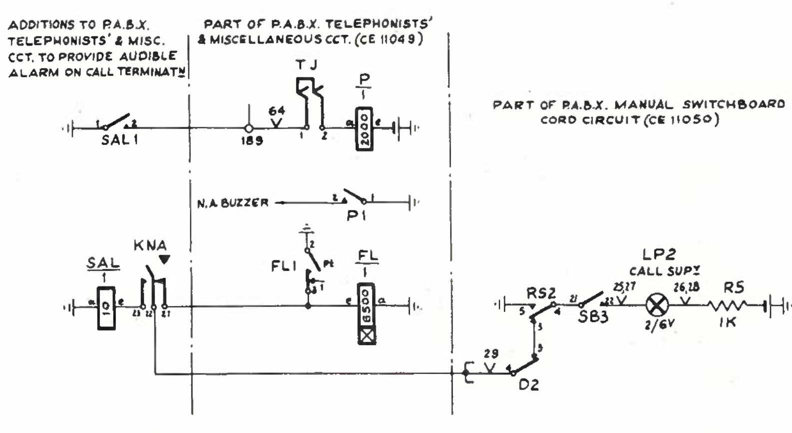

Fig. 11 — Circuit of P.A.B.X. Modified for Audible Cord Supervision.

Application to P.A.B.X. Switchboards: It is not intended to discuss at length the modification of P.A.B.X. switchboards. The principles as described above are applicable. However, in this case, the modification to achieve audible supervision can be made electromechanically by the addition of a relay. Fig. 11 illustrates this point (8).

Sighted Operators: The additions and modifications to the switchboard do not affect standard operation by a sighted operator.

Other Applications: For some time requests have been received from subscribers to have audible cord supervision included in their lamp signalling P.M.B.X's. A typical example is that of a hotel which, while having to keep the switchboard open 24 hours a day, cannot economically employ an operator during all that time. To date, these requests have had to be rejected due to the technical difficulties involved. Now, however, the supervisory units as described above could be used to provide the desired facility.

CONCLUSION

The Department has fulfilled a real community service in this matter and has earned the deep gratitude of the individuals and organisations concerned.

The detailed circuit and mechanical drawings are indexed on drawings VX4548/1 and V.B.2279/1.

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ACKNOWLEDGMENT

The authors wish to acknowledge the contribution made by Mr. F. W. Wier.

REDUCTION OF NOISE GENERATED BY ENGINES INSTALLED IN TELEPHONE EXCHANGES — PART II

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THE DESIGN OF VENTILATING DUCTS

It has been pointed out in the previous part of this series (Vol. 12, No. 2) that it is often not possible to insulate a source of noise except by recourse to artificial ventilation, whereby the necessary air supply to the engine room is maintained and an adequate transmission loss is preserved. In this section, the design of ventilating ducts is discussed. The design of the necessary duct work is undertaken in the course of the solution of the noise problem after this has been formulated, and enables the work to proceed in an orderly and systematic fashion.

Ducts used for ventilating purposes may be constructed from a variety of materials, for example, masonry, timber, synthetic board or sheet metal. They may be integral with the building construction (a chimney is in this category) or installed subsequent to it, a practice often adopted in air-conditioning plant installation. Consideration of the design of ventilating ducts may be made under two headings dealing with the ventilating and the acoustical aspects.

Mechanical Design

The purpose of a ventilating duct is to transfer air from one point to another at a given rate. Ideally, the conditions of flow of an incompressible fluid, in which the effects of friction and turbulence can be neglected, are expressed by the Bernouilli equation, namely:—

$$\frac{V^2_1}{2g} + \frac{P_1}{p_1} + z_1 = \frac{V^2_2}{2g} + \frac{P_2}{p_2} + z_2,$$

where V = velocity in feet per second,
g = acceleration due to gravity,
P = pressure in pounds per sq. ft.,
p = density of the fluid in pounds per cu. ft.,
z = elevation above datum in ft.

For air, the differences ($z_1 - z_2$) and ($p_1 - p_2$) are usually negligible. The pressure difference ($P_1 - P_2$) is the head required to maintain equilibrium conditions of flow in a pipe of uniform diameter. Taking friction into account, the pressure loss in a straight, circular pipe is expressed by the equation:

$$h = f . \frac{L}{d} . \frac{V^2}{2g},$$

where h = loss of head due to friction in feet,
L = length of pipe in feet,
d = effective diameter of pipe in feet,
V = velocity in feet per second,
g = acceleration due to gravity (32.2 ft./sec.^{2}),
f = frictional coefficient.

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A practical chart has been published by the A.S.H.V.E. (14) from which has been derived a graph appropriate to our purpose, Fig. 13. The scales of this graph refer to the roughness of the duct surfaces.

Scale A: Roughness 0.006″. An average value for unlined ducts of riveted sheet iron or dressed timber.

Scale B: Roughness 0.04″. Medium rough surface, for example, concrete or plaster or asbestos cement sheet.

Scale C: Roughness 0.1″. Very rough surface, for example, spiral riveted duct, sprayed plaster or perforated metal.

Scale D: Roughness 0.2″. Extremely rough surface, for example, roughcast cement or plaster, expanded metal, "nofines" concrete, sprayed asbestos fibre, etc.

The graph refers to circular ducts. To find the circular duct equivalent to a rectangular duct of sides a x b the following formula may be used:

$$d = 1.265 . \left[\frac{(ab)^3}{a + b} \right]^{1/5}$$

or; log d = 1/5
[3(log a + log b) − log (a + b)] + 0.1021

Alternatively the graph of rectangular equivalents (Fig. 14) may be used. This graph is most useful for the determination of the equivalent rectangular duct.

In accordance with hydraulic practice the losses caused by elbows may be expressed as equivalent lengths of straight duct. For square elbows the equivalent length of duct is 80 d where d is the diameter of the equivalent circular duct. The equivalent length falls to from 10 to 20 d when the elbow is made on a curve (at the duct centre line) of radius not less than the diameter of the equivalent circular duct. For design purposes, a value of 20W (where W is the width of the duct at the bend) should be used in estimating the loss of a curved elbow of rectangular cross section irrespective of the radius of curvature, provided this is not less than W.

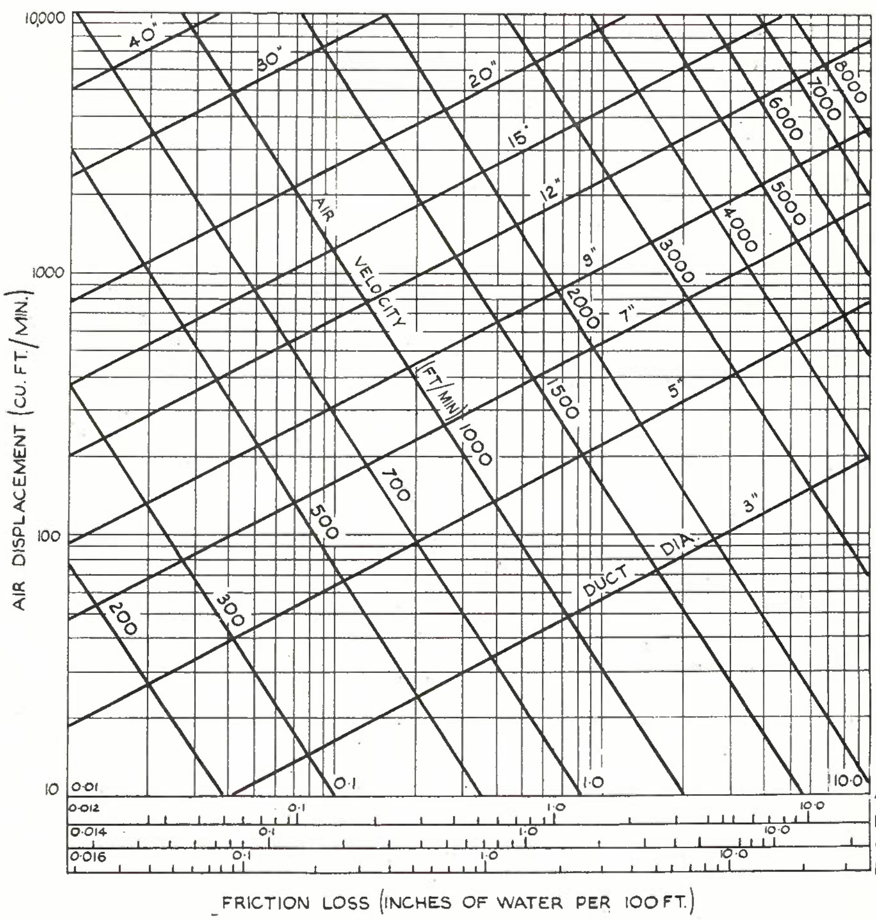

Fig. 13 — Loss versus Displacement of Round Ducts.

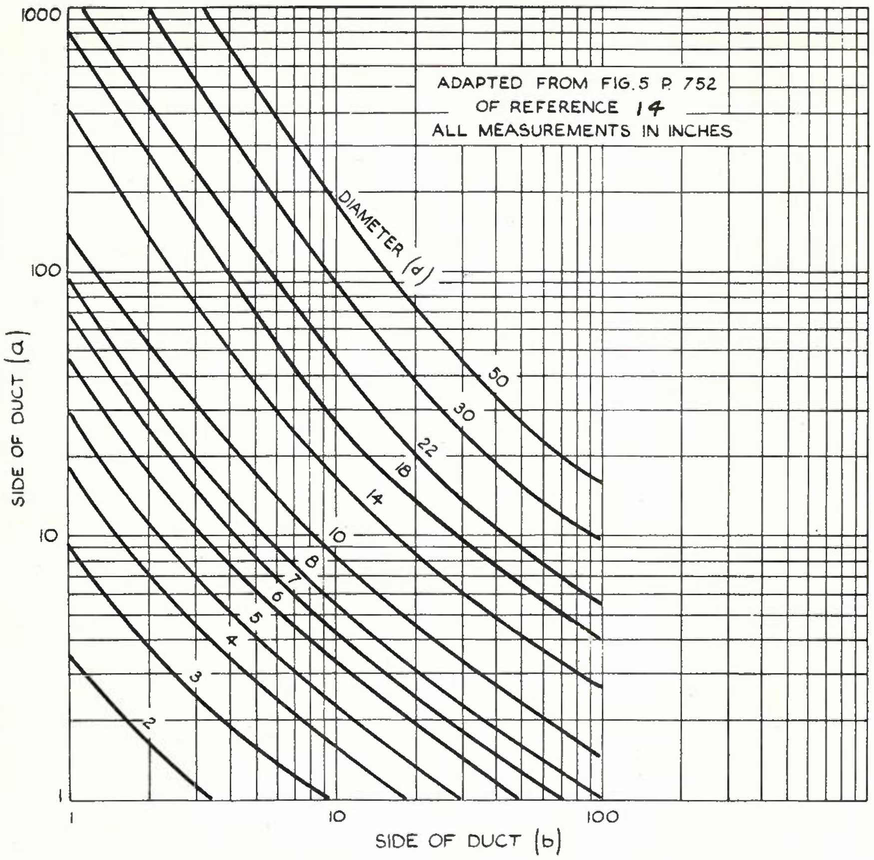

Fig. 14 — Rectangular Equivalents of Round Ducts.

The permissible air velocity in the duct is governed by—

(a) the pressure head available;
(b) the liability of the duct walls to be set into vibration;
(c) the resistance of the duct to erosion; and
(d) the permissible noise at the inlet or outlet grills.

The static pressure head (a) is determined from the fan characteristic and loss curve of the duct. Factor (b) is unlikely to be important if the walls are required to have good sound attenuation.

With regard to erosion (c), cement and plaster linings will resist erosion by air at velocities up to 5000 ft. per min. Sound absorbent protected by perforated metal will withstand a similar high velocity but sprayed asbestos or rockwool in batts or blankets or loose, retained behind wire mesh or expanded metal, should not be exposed to air at a higher velocity than 2000 ft./min.

Operating characteristics for axial flow fans of the type used in water-cooled internal combustion vehicular engines are not readily available, and in the absence of experimentally determined curves a 45° regression of percent discharge volume on percent static pressure may be adopted or a curve similar to Fig. 15 may be used. It is then necessary to know either the blocked static pressure, the maximum discharge volume or some intermediate point on the curve. The latter may be determined by measurements made on any duct connected to the same type of fan.

The noise caused by air-flow at the inlet or outlet grills is considered in the next section.

Acoustical Design

We commence the acoustical design with a knowledge of the two basic requirements of total attenuation, N db, to be provided by the duct and the cross-sectional area, A sq. ft., necessary to allow the required airflow to be maintained by the fan.

The sound intensity in the duct is attenuated by dissipation of energy at the walls and by reflection at discontinuities. If the walls of the duct are rigid and impervious, there is a negligible loss due to friction at the wall of the air particles in the boundary layer. However, if the walls are covered with absorptive material an additional loss due to viscous motion of the air in and out of the pores of the material will occur.

Since the loss is proportional to the perimeter of the lining and the total energy flux of the sound waves is proportional to the cross-sectional area of the duct, the attenuation per unit length is proportional to the ratio of the perimeter and cross-sectional area.

In Part I, it was shown that, in a room containing a sound source, the total reverberant energy removed from the room each second is—

$$W_R = DV\alpha N = DV\alpha.\frac{cS}{4V},$$

where D is the reverberant energy density, V the volume and S the surface area of the room, α the mean absorption coefficient and c the velocity of sound. A unit length of duct may be considered as a room without end walls and containing only reverberant energy. The absorbent area of this length of duct is numerically P, the perimeter, and the volume A is the cross-sectional area. If W_R is the energy loss per unit length, the energy density decreases by an amount

$$\Delta D = \frac{W_R}{V} = \frac{W_R}{A}.$$

The attenuation may therefore be expressed as—

$$\frac{\Delta D}{D} = \frac{\alpha c}{4} . \frac{P}{A} .$$

In practice, the loss is not completely in accordance with this expression since the propagation is not random.

An empirical formula for straight ducts was established by Sabine as follows:—

$$\text{Loss} = 12.6\ \alpha^{1.4}\ P/A \text{ db per ft.}$$

where α is the chamber absorption coefficient, that is, the coefficient determined by a reverberation chamber method which is the one usually quoted for any material; P is the internal perimeter of the duct and lining, and A is the open area of the duct. The last two values must be expressed in inches and square inches respectively.

It has been found that the Sabine equation is approximately correct only under restricted conditions, that is to say, when the ratio of the cross-section dimensions of the rectangular duct is no more than 2:1 and the absorption coefficient is not greater than 0.4. However, experience in the use of the Sabine formula in the design of ducts for two Melbourne automatic exchanges indicates that good results are obtainable with its use, when an optimum solution is not required. A nomogram for the solution of the formula is given in Fig. 16. It should be noted that in the Sabine formula, if the duct area A and

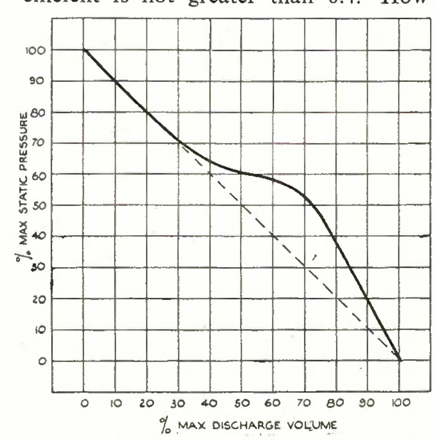

Fig. 15 — Operating Characteristic of Axial Flow Fan.

absorption coefficient α are held constant, the total attenuation of a duct is proportional to the product of the perimeter and length of lining. Hence the total area of lining used to obtain a given attenuation with a duct of prescribed area is not dependent on the duct proportions.

Example: A duct of 200 sq. inches cross-sectional area and 30 db attenuation is required using an absorbent lining of 0.35 absorption coefficient. From Fig. 16, a duct of 60 inches perimeter provides an attenuation of 0.9 db per foot and the length of duct required is therefore 33 ft. The total area of lining is 165 sq. ft. The duct in this case has dimensions 10 inches x 20 inches.

A duct of dimensions 14 inches x 14 inches would have the same area as previously but a perimeter of only 56 inches. The attenuation per foot is only 0.8 db and the length of duct required is 36 ft. giving a total area of lining the same as before.

Duct Wall Transmission

The required transmission loss of the walls of the duct is most simply determined by considering it to be equal to the attenuation provided by the duct at any point. Thus the inlet end of either the inlet or the exhaust duct is considered to be a point of zero attenuation and wall transmission loss requirements are small.

The minimum wall transmission loss along the duct length is then determined by the attenuation from the inlet. However, if the duct passes through a noise barrier, such as the engine room wall, the transmission loss of the duct walls at this point must not be less than the total duct attenuation. Usually, the duct is constructed throughout with the one gauge of sheet metal and the losses, calculated by the mass law, of various gauges and given in Table 7, may be used as a guide.

Table 7

Gauge	Weight (lbs./sq.ft.)	Transmission loss (db)
16	2.55	28
18	2.02	27
20	1.59	26
22	1.27	25
24	1.01	24
26	0.19	23

In order to realise the transmission loss determined by the mass law it is necessary to take precautions to avoid panel resonances in the duct walls and leakage of sound at joints. The walls should be prevented from vibrating by "cross-breaking" the panels or by adding stiffeners. The addition of a heavy coating of mastic to the outside of the duct is an advantage where the absorbent lining has no intimate contact with the walls such as is obtained with sprayed or cemented absorbent. Seams in sections of ducts should be closely riveted, welded or otherwise closed, or sealed with a mastic. Joints between sections should be drawn up to rubber insertion or sealed with a mastic.

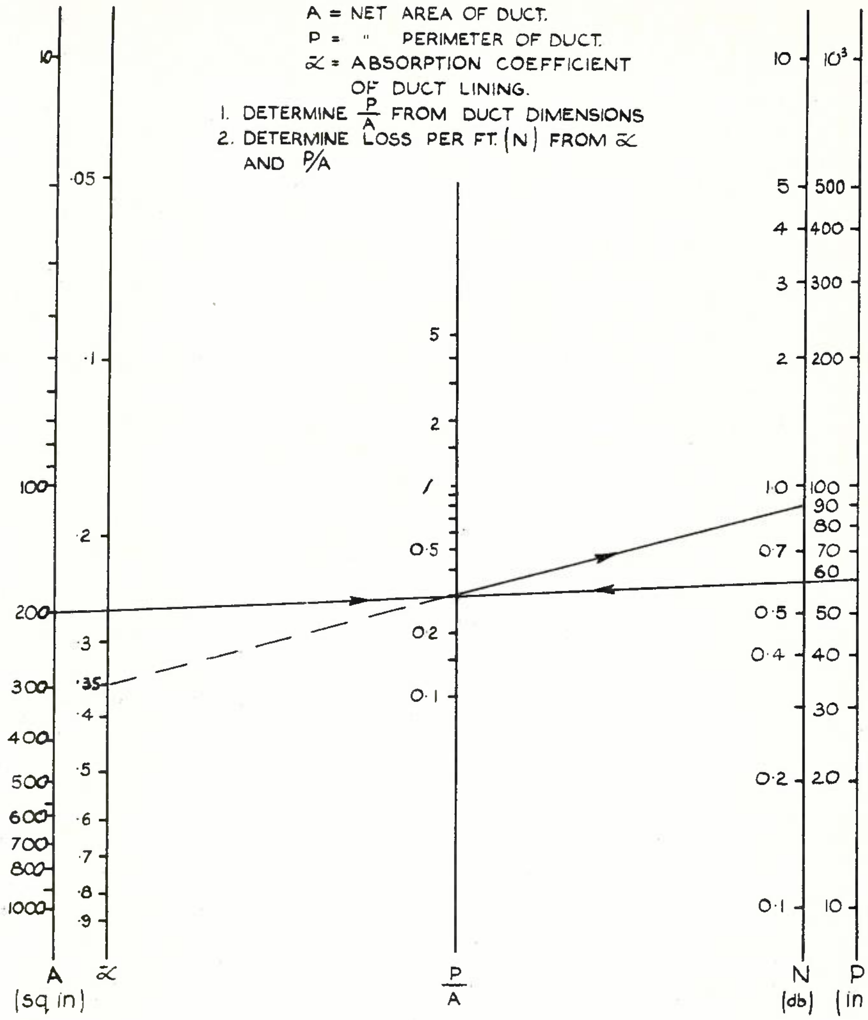

Fig. 16 — Attenuation of Sound-Absorbing Ducts.

It should be noted that the type of duct construction usual in ventilating work is not adequate in the high noise levels occurring in engine rooms.

Plenum Chambers

When a duct of sufficient length to provide the requisite attenuation of sound cannot be installed for engineering or other reasons, recourse may be made to the use of a large absorption chamber or plenum. An advantage obtainable with a plenum is that the absorptive lining may be designed to have a maximum coefficient at a relatively low frequency. In general this would require the use of a very thick lining or of air-space behind the lining. It should be noted that whenever an air-space is used as an integral part of an absorptive lining in a duct or plenum it should be interrupted by barriers across the line of flow of power if shunting of the acoustic attenuation of the duct is to be avoided.

The design procedure to be followed for a plenum is based on the concept of the room constant. The attenuation of the plenum in decibels is given by:—

$$N = 10 \log \frac{P_{IN}}{P_{OUT}}$$

where P stands for the power entering and leaving the chamber.

Then,

$$\frac{P_{OUT}}{P_{IN}} = S\left(\frac{\cos \Theta}{2\pi d^2} + \frac{1}{R}\right)$$

In this equation S is the area of the inlet or outlet, d the slant distance from inlet to outlet, Θ the inclination of d from the inlet axis, and R the room constant is equal to α $(1 - \alpha)$ where α is the absorption coefficient of the lining. The geometry of the plenum is made plain in Fig. 17, which illustrates a baffle arrangement whereby very high attenuation is achieved by ensuring that only reflection paths exist between the inlet and outlet. In the equation given above the direct component of the energy transferred through the chamber is $(S \cos \Theta)/(2\pi d^2)$. With this component absent the energy transfer is dependent directly on the area of the port and inversely on the room constant,

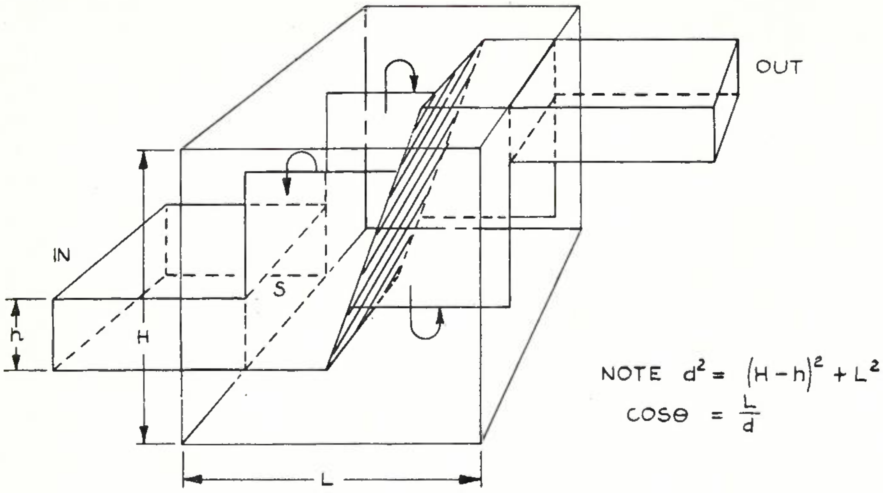

Fig. 17 — Plenum Chamber with Baffle.

The attenuation is, then, simply

$$N = 10 \log \frac{R}{S} = 10 \log \frac{\alpha}{(1-\alpha)S},$$

which for fixed S may be made as large as desired by increasing α. For large S the attenuation decreases; the plenum in fact degenerates into a mere continuation of the duct.

It will be noted that the volume of the chamber does not enter into the calculation of attenuation. In practice, the chamber will resonate in its normal modes so that standing waves are set up in it. The larger the chamber the lower will be the frequencies of these modes. There is increased absorption of energy at these frequencies which therefore increases the attenuation by some 5 to 10 db.

Splitters

A splitter is a panel of sound absorptive material inserted in a duct along the axis of energy flow. The splitter is not impervious to sound as in the sectionalised duct. A section through an attenuator consisting of splitters is shown in Fig. 18 (a); an improved arrangement which provides better attenuation at high frequencies is shown in Fig. 18 (b).

Experiments have been made (15) with the simple splitter of Fig. 18 (a) in which the cross-section is divided equally between the air-channels and the absorbent. The curves of Fig. 19 were produced as a result. The absorbent used in these experiments was rock wool of 6 to 9 lb. per cu. ft. density. The attenuation, in db per ft. length of silencer, depends upon the frequency of the noise component, the thickness of the splitter and the width of the air channel. For constant splitter thickness and frequency, the attenuation increases as the air channels are reduced in width. With other conditions fixed, the attenuation also increases with thickness of splitter up to a point where the thickness is about a quarter wave length. Lastly, for fixed air channel and splitter dimensions, the attenuation increases with frequency up to a point where the wave length is less than about two-thirds of the air channel width and the sound begins to beam down the air channels without striking the sound absorbent surfaces. The advantages of the improved splitter design of Fig. 18 (b) can be appreciated as providing a more uniform attenuation over the frequency range and also extending that range to a higher frequency. A similar result may be achieved with a simple type of splitter if it is constructed in two stages, one for low-frequency and the other for high-frequency components. In this case added advantage may be gained by using a high density absorbent for the low frequency section and an absorbent of half that density for the high frequency section.

ACOUSTICAL MEASUREMENTS

Two kinds of acoustical measurement may be recognised. The first is properly a part of the science of psycho-acoustics and involves the subjective reaction or response of a listener or of a team of listeners. The second is wholly physical and usually (with notable exceptions) involves the transduction of acoustic into electric energy. In the physical analysis of noise the latter principle is always adopted since the techniques of the electrical engineer may then be utilised.

The Subjective Measurement of Loudness

Simple quantitative judgments of the loudness of a sound are readily made by a person and are expressed in a scale such as "deafening", "very loud", "loud", "soft", etc. Attempts have been made by psychologists to establish a more precise scale and the sone-scale is a result of this work. Initially a reference point must be fixed and this was chosen as the loudness of a pure tone of 1000 c/s frequency of intensity 40 db above a normal listener's threshold of hearing. The loudness of this tone is one sone. The scale of loudness was then determined by asking subjects to change the loudness of the 1000 c/s tone by factors of 2, 10, 0.5 and 0.1. From the data thus obtained a relationship between loudness and loudness level (discussed in the next section) may be determined. One such relationship is given by Fletcher (2) but a more recent determination has been made at the National Physical Laboratory, London, with the result shown in Fig. 21.

It is well known that a narrow band of noise filtered out of a wide band noise such as white, random noise is heard with a sensation of pitch. From experiments made with bands of noise it has been suggested that a band whose width does not exceed about 600 mels* sounds as loud as a pure tone of the same intensity located at the mean frequency of the band. Means are therefore available for dealing with noise as well as pure tones.

Loudness level. An alternative way of evaluating the loudness of a sound is to compare it with a standard sound which can be varied in intensity in a defined way. The result of subjective experiments of this type are the "equal-loudness contours" for the pure tones and narrow bands of noise. The standard sound is a tone of 1000 c/s frequency and the reference pressure level of this tone is taken as 0.0002 microbar. The unit of loudness level is the phon and the loudness of a sound in phons is equal to the sound pressure level of the equivalent 1000 c/s tone in decibels above the reference level. It should be noted that although experiments have been made only with pure tones and narrow bands of noise in order to determine the relationship between sound pressure level and loudness level, any sound however complex may have its loudness level estimated in phons by comparing it sub-

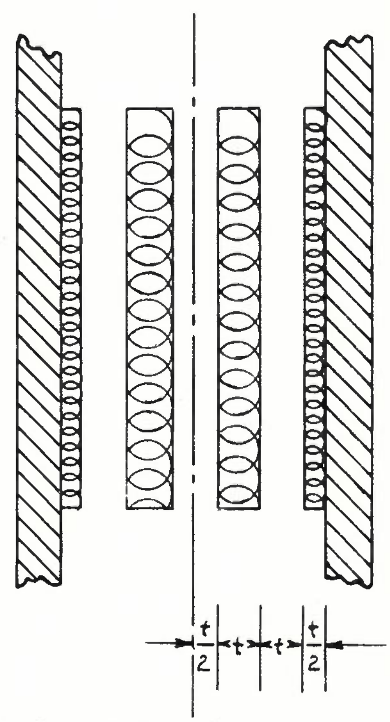

Fig. 18 (a) — Section of Duct with Absorptive Splitters.

*The mel is a measure of the pitch of a sound (a subjective response), the reference being taken as the pitch of a 1000 c/s tone at a pressure level 40 db above 0.002 micro bar. The pitch of pure tones varies predominantly with frequency and only slightly with sound pressure level. A 600 mel band occupies the frequency ranges 40-550, 620-1500 and 4200-8500 c/s, for example, and is roughly approximated by an octave band.

jectively with the 1000 c/s reference tone. The inclusiveness of the definition of loudness level has resulted in a number of attempts to construct a so-called phon-meter and in some confusion as to the nature of the information provided by such meters. This subject is discussed further with respect to the sound level meter.

The earliest equal-loudness contours, those of Fletcher and Munson in 1933 and Churcher and King in 1937, showed

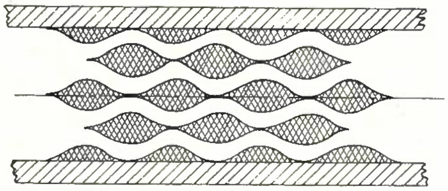

Fig. 18 (b) — Improved Splitters.

considerable discrepancies, especially at the lower frequencies. The recent determination by Robinson and Dadson is probably the most accurate.

However, the equal-loudness contours for narrow bands of noise determined by Pollack (17) agree best with the Churcher and King data (18) and for the present purpose the latter will be used, Fig. 20. This has the additional advantage that similar use has been made of them in the U.S.A.

Sound Pressure Level (S.P.L.)

The determination of sound pressure level is at present the basis for the objective measurement of sound since the measurement of other absolute quantities such as particle or volume velocity, sound intensity and sound energy density is difficult or complex.

The sound pressure level is defined as the ratio in decibels of the measured effective sound pressure to a reference effective sound pressure according to the equation

$$SPL = 20 \log \frac{p}{p_{ref}} \text{ db.}$$

The reference pressure may be either 1 microbar (1 dyne/cm^2) or 0.0002 microbar. The two reference levels differ by 74 db.

The S.P.L. of a pure tone may be measured with a sound pressure meter or with any calibrated pressure microphone amplifier and r.m.s.–reading voltmeter combination. The American sound level meter (19) when used in the "C" or "Flat" weighting position is a suitable instrument.

The concept of pressure level may be extended to include sounds having a continuous distribution of energy with respect to frequency. The reading of a sound pressure meter will then be dependent on its frequency bandwidth, assuming a uniform response throughout the band. The increase in the S.P.L. due to bandwidth is $C = 10 \log \Delta f$ (db) where Δf is the increase in bandwidth. The pressure level, L, of a uniform band of noise within the frequency range $f_a - f_b$ may be expressed in terms of its spectrum level S as follows:

$$L = S + C \text{ db.}$$

Sound Level. The term "sound level" has become restricted through usage to mean the reading in decibels obtained when a sound level meter is used in a sound field. The meter reading corresponds to a value of sound pressure integrated over the audible frequency range with a specified frequency weighting and integration time. Specifications for sound level meters have been drawn up in several countries and are in good agreement.

The sound level meter in its initial conception was intended to indicate the loudness level (in phons) of a complex sound. It has been known, in various places, as a "phon-meter", an "objective noise meter", a "sound meter", and a "sonometer". These alternatives have a number of defects; for example, the phon-meter does not measure phons; the objective noise meter does not measure electrical noise as does another type of noise meter; the sonometer does not measure sones. An analogy exists between the sound level meter and the psophometer — each was intended to simulate quantitatively a subjective response to noise and each gives a measure of the noise in an arbitrary manner. The Western Electric noise measuring set, in fact, is designed to function as both types of instrument.

Difficulties in attaching meaning to sound level measurements arise as a result of the limited approximation that the sound level meter makes to the hearing process. In the first place, three arbitrary weightings are used corresponding to the 40 and 70 phon contours of the Fletcher-Munson equal-loudness determination for pure tones, and a "flat" weighting for sounds whose loudness level would be expected to be above 100 phons. Gross errors may result when the spectrum of the sound being measured is very uneven, while doubt on the part of the operator as to what weighting to use may also be experienced. In the second place, although the meter is designed to have a dynamic characteristic similar to that of the ear, the mechanism of hearing is extremely complex and cannot so easily be simulated. Sound level meter readings of industrial noises may therefore differ considerably from subjective assessments of the loudness level of such noises.

Band Pressure Level. The measurement of the sound pressure level in a band of frequencies has been mentioned. It is now customary to perform an analysis of wide band noise by means of sets of filters. The filters are usually octave or half or third octave in bandwidth, that is, the ratio of upper to lower cut-off frequency is 2, 1.41 or 1.26 respectively. The band pressure level in any frequency band may be reduced to an equivalent spectrum level as described previously. For convenience, the correction factor, C, is given in Table 8 for the commonly used octave filters. Also given are the mean frequencies of the bands. In plotting noise spectra, it is usual to plot the spectrum level at the geometric mean frequency

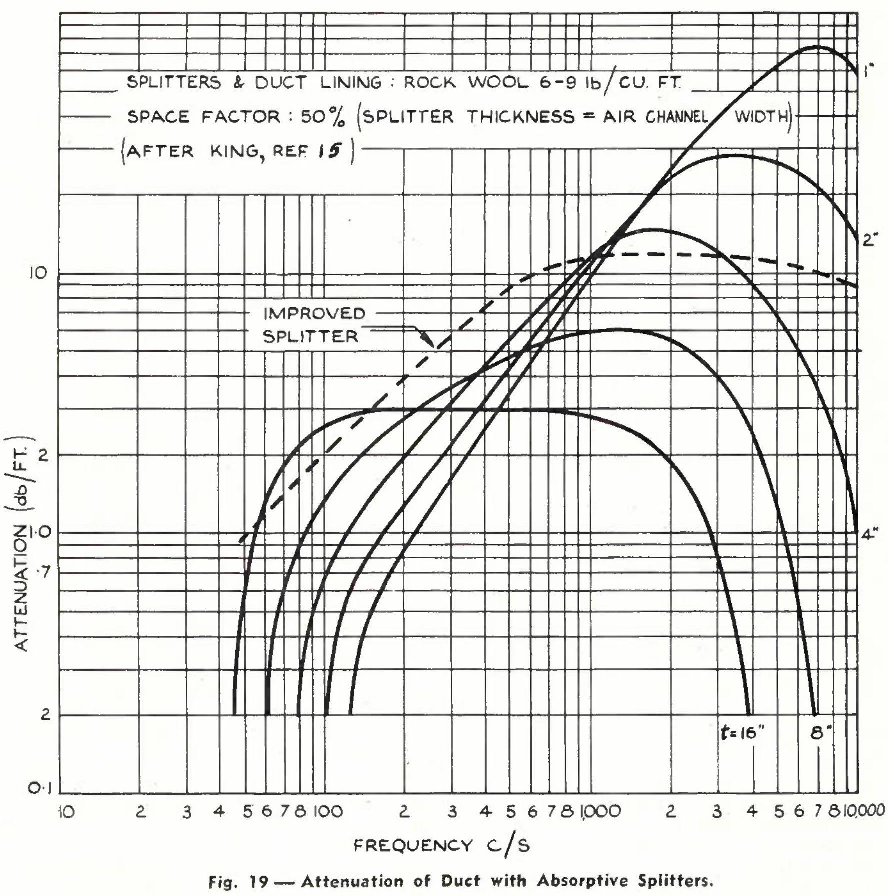

Fig. 19 — Attenuation of Duct with Absorptive Splitters.

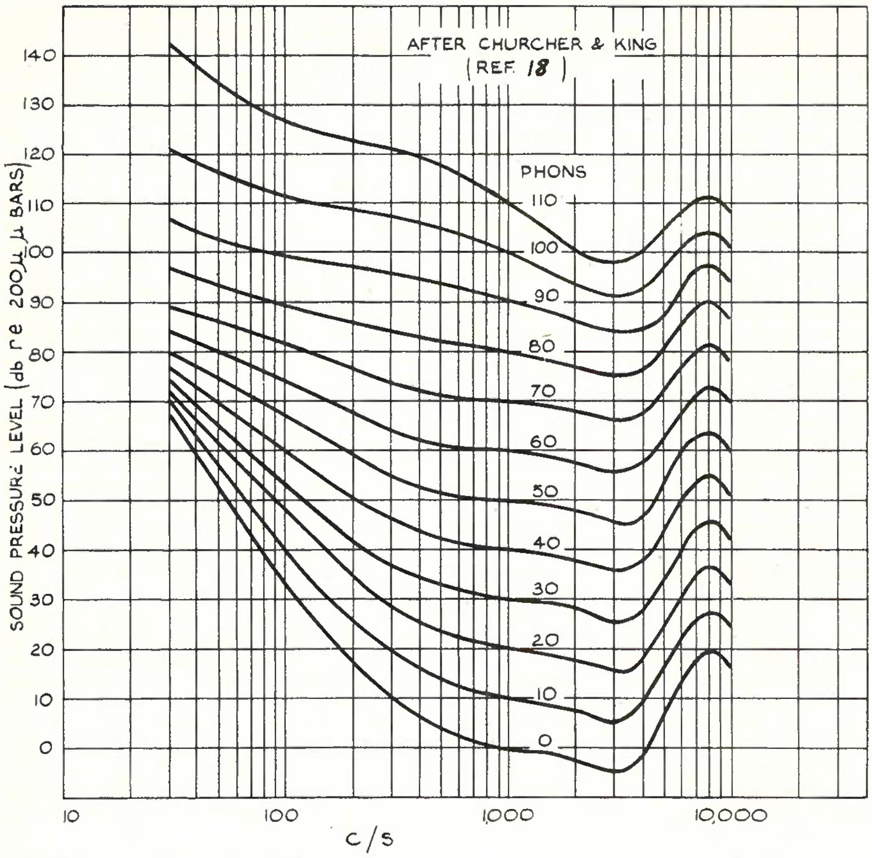

Fig. 20 — Equal-Loudness Contours, Free-Field.

of each band and draw a smooth curve through the resulting series of points.

Table 8
Octave Band Pass Filters

Fc_1 Lower (c/s)	Upper Fc_2 (c/s)	ΔF (cy.)	$C = 10 \log \Delta F$ (db)	$F_M = \sqrt{Fc_1 Fc_2}$ (c/s)
37.5	75	37.5	15.8	53
75	150	75	18.8	106
150	300	150	21.8	212
300	600	300	24.8	424
600	1200	600	27.8	848
1200	2400	1200	30.8	1696
2400	4800	2400	33.8	3392
4800	9600	4800	36.8	6784

For precise work, the filters used to determine band pressure levels should have their characteristics corrected for departure from the ideal nominal filter (see Ref. 20, Section 12.4).

The band pressure level in a band of width Δf_1 may be converted to a band pressure level for a band of width Δf_2 by adding algebraically the bandwidth correction factors provided that only continuous spectrum noise is being considered, the mean frequency remains the same and the two bandwidths are not too different having regard for the uniformity of the noise spectrum.

Acoustic Power Level (P.W.L.). The acoustic power level of a source is defined as: —

$$PWL = 10 \log \frac{W}{W_{ref}} \text{ db,}$$

where W is the acoustic power radiated and W_{ref} is a reference acoustic power which is usually taken as 10^{-13} watt. Thus, a source radiating one acoustic watt has a PWL of 130 db.

The measurement of the P.W.L. is a derived one based on the determination of sound pressure levels in the far-field of the source. The S.P.L. at a point in the far-field is completely determinable if the P.W.L. and directivity of the source and the characteristics of the environment are known. Conversely, the P.W.L. can be found if the S.P.L. distribution in the field and the characteristics of the environment are known.

A full discussion of this topic may be found in the literature (for example, Ref. 7, p. 367). In the present context, it is difficult to estimate P.W.L. of a diesel-engine operating in a small enclosure from measurements of the S.P.L. since the far-field requirement cannot be met. However it has been proposed that the P.W.L. of diesel alternator sets be sought from manufacturers in conjunction with their tenders for the supply of such machines and, since the determination of P.W.L. under suitable conditions is basically simple, this is to be recommended.

The Loudness Chart

The chart given in Fig. 22 offers a ready means for determining the loudness or speech interference of a noise for which an octave band analysis is available. The band pressure level within each octave band is plotted on the appropriate ordinate of the chart, interpolation between the contours of constant band pressure level being used where necessary. The band loudness in sones of each component may then be read off by means of the inner left-hand scale. The sum of the loudest component and half the remainder is then marked on the outer left-hand scale giving the "total loudness in sones" of the noise. The two right-hand scales are used to determine the loudness level of the components or the speech interference level of the noise. The total loudness level of the noise is obtained from the total loudness in sones by referring to the loudness level scale at the corresponding ordinate.

Example: A noise with the following analysis in octave bands is found by means of Fig. 22 to have a total loudness of 17 sones, a loudness level of 85 phons and a speech interference level of 53 db.

The steps taken in the determination are as indicated on next page.

The chart forms a permanent record of the noise analysis and the loudness, loudness level and S.I.L. determinations; additional copies are available from the Acoustics Division of the P.M.G. Research Laboratories, Melbourne.

Commonsense in Acoustic Measurement

At present the only method of obtaining complete data of noise is to use a sound pressure meter and wide-band (octave or third-octave) analyser supplemented as necessary with a narrow band analyser for pure tone components. By the use of such data a frequency spectrum of the noise, including any singularly loud components, may be calculated.

On occasion, it may be necessary to measure the reverberation time of enclosures or the transmission loss of walls, etc. As such measurements are outside the scope of this article and may be carried out only with the aid of specialised apparatus, the reader is referred to the appropriate literature (4, 20). However, it is often sufficient to estimate the room constant by consider-

Band	37.5	75	150	300	600	1.2	2.4	4.8	Total
Pressure level (db)	42	55	62	75	61	52	45	35	—
Loudness (sones)	0	0.6	2.8	9.5	4.4	2.8	2.1	0.3	$9.5 + \frac{13.0}{2} = 17.0$
Loudness level (phons)	—	34	56	72	61	54	50	21	85
S.I.L. (db)	—	—	—	—	61	52	45	—	53

ing the nature and area of the absorbing surfaces of an enclosure and the transmission loss of walls, etc., by using the mass law (Fig. 3).

The sound level meter may be used in conjunction with a set of filters as depicted in Fig. 29. Here, the meter is being used as an amplifier and indicating instrument, with the filter interposed, modern developments in sound level meters permitting them to be used in this way. Alternatively, a separate indicating meter may be used, in which case the meter characteristics should be properly chosen to obtain an r.m.s. integration over the passband.

A tape-recorder is a very useful instrument in noise control; it offers the advantages of enabling the noise analyses to be carried out in the laboratory and of preserving a record so that comparisons of noise before and after the application of control measures may be made. However, it is essential that the record-reproduce frequency response of the machine be known and, if sound pressure levels are required, the sensitivity of the recorder must be known. A reference tone of known intensity should therefore be recorded on the machine before the actual recording is made. A portable oscillator and sound source of the type shown in Fig. 27 may be used for this purpose.

When measurements are made out-of-doors, a windscreen may be necessary to avoid wind noise at the microphone. A suitable screen is shown in Fig. 28; in an emergency, a handkerchief or double layer of muslin or gauze may be draped over the microphone.

For qualitative work or for comparing noises known to have similar spectra the sound level meter (of which miniature versions exist) alone is very useful. It has been used, for example, to obtain statistics of noise levels in exchanges. Sufficient information is usually provided by manufacturers in the handbooks supplied with these instruments to enable the user to beware of any pitfalls.

ANNOYANCE

The attributes of noise which cause annoyance are those which involve the subjective, mental or emotional, reactions of the individuals exposed to it. The response which a person makes to noise may be extremely varied and is dependent on his temperament, physical and mental condition, habituation to noise, etc. It is possible to make generalisations only about comparatively large groups of people and even then only of those matters for which reliable statistics have been obtained by the use of adequate samples.

Attributes of Annoyance

Observation shows that certain features of noise are important in determining the response. The most discernable of these features are the following.

(a) **Unexpectedness.** Startle or fright reactions are elicited by the element of unexpectedness. The peak amplitude, time of onset and (steady-state) frequency spectrum of the noise are controlling factors.

(b) **Interference with Audition.** Listening to speech or music from either a "live" or electrical source against a background of noise may be most difficult. The annoyance caused by static in radio reception has been investigated, and it is well known that intelligible crosstalk on a telephone channel may be very annoying.

(c) **Inappropriateness.** Annoyance is increased if the interfering noise is objectionable for aesthetic or similar reasons. Noises disregarded during the day, such as industrial noises, may be very irritating at night. Noise that exhibits strongly the characteristic inappropriateness is almost always "man-made" and elicits the response that "something should be done about it" according to the subject's estimate as to the ease with which remedial action may be carried out.

(d) **Directionality.** It has been found that noise is more annoying if the source cannot be located or if the direction of arrival is not readily apparent. This is particularly so indoors.

(f) **Frequency Pattern.** For noises of equal loudness, as judged by subjective methods, those containing higher frequencies are more annoying. Conversely, reducing the high frequencies has been found in a number of instances to remove the unpleasantness previously associated with the noise.

(g) **Intermittency.** An intermittent noise resists habituation more than a continuous sound. Almost invariably, irregular mechanical noises are initially displeasing.

(h) **Fear.** In the animal world, the ear along with the eye and the nose is one of the basic organs, the use of which is an important element in the preservation of life. In man, there may be other conditioned reflexes as well, resulting from auditory stimulation. The most spectacular of these is, perhaps, the response of an individual to the whistle of a bomb or shell.

Evaluation of Annoyance

The most important work on the evaluation of annoyance caused by noise has been carried out in U.S.A. No objective analyses have been made in Australia. The American work has been concerned with annoyance in offices and in residential areas.

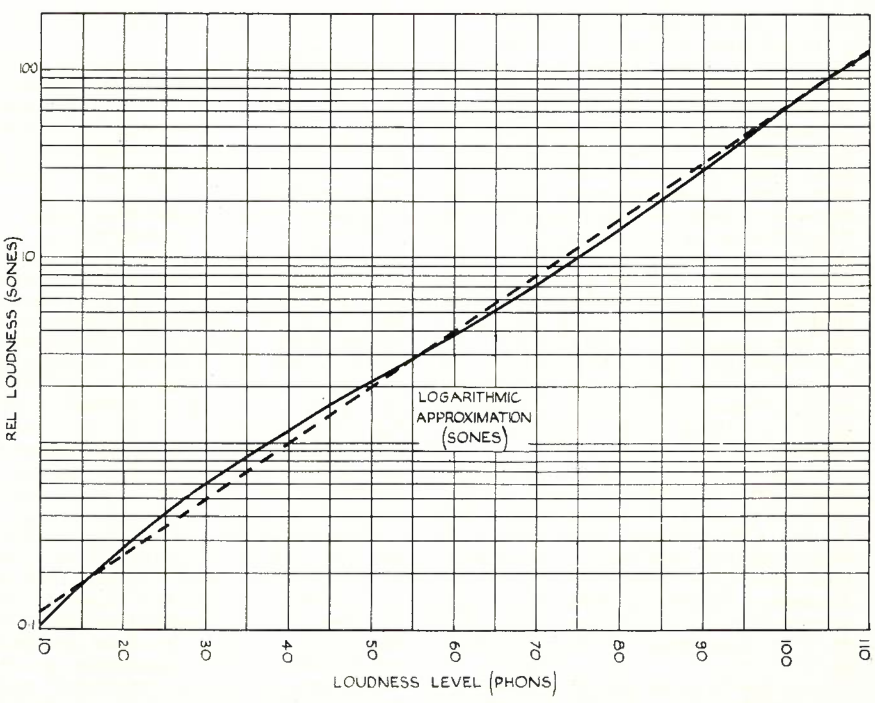

Fig. 21 — Loudness versus Loudness Level.

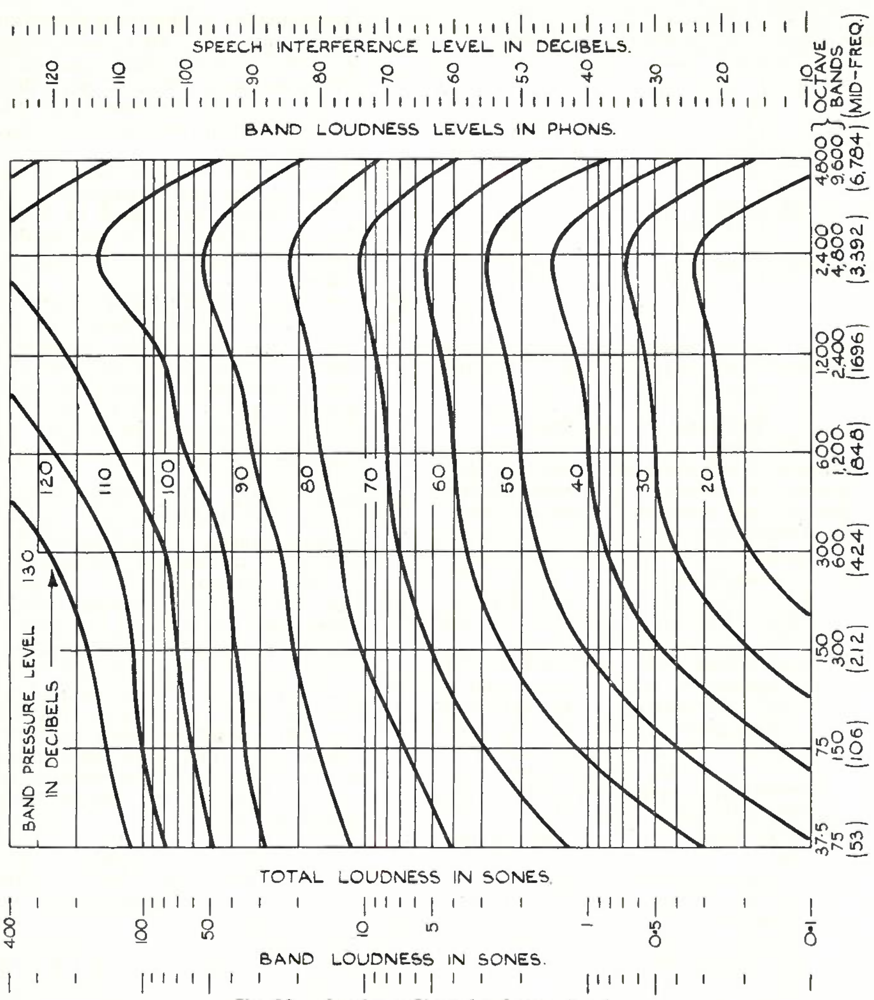

Fig. 22 — Loudness Chart for Octave Bands.

Annoyance in Offices

Although, as a result of a number of experiments, there appears to be no serious disadvantage due to the presence of noise attached to the accomplishment of manual or intellectual tasks, there is a tradition that the less exacting manual tasks may be performed with undimished proficiency in the presence of noise whereas intellectual tasks need quietness. The influence of this traditional view is difficult to escape when dealing with naive subjects. A large degree of noise suppression is therefore regarded as desirable on general grounds by office workers engaged on isolated tasks.

In addition, an office-worker will become more conscious of the adverse effects of noise when he has occasion to speak to others either directly or by telephone.

A survey of opinions and recommendations for office-quieting criteria have been made by Beranek (21) and Kryter (22). From these data it is concluded that the speech interference level should not exceed 40 db and the loudness level of the noise should not exceed 62 phons if ease of conversation at 8 ft. separation or if comfortable telephone conversation are to be maintained.

Situations in which criteria for other than office space are needed are unlikely to arise in acoustic design associated with telephone exchanges. However, data for other types of occupation have been published (23) and may be used with confidence.

Annoyance in Residential Areas

Criteria for the evaluation of the annoyance caused by noise in residential areas have been developed by Rosenblith and Stevens in U.S.A. (7*) on the basis of known instances of annoyance which were thoroughly investigated. These criteria are, however, tentative in their application to other types of noise and groups of people of different sociological characteristics. Their use in Australia should therefore be experimental and with a view towards their modification to accord better with local requirements.

The basis of the method of estimating annoyance is the determination of a set of curves which define the level rank of a noise when measured in octave bands. The level rank is modified according to the extent to which the noise possesses other characteristics depending on its short term frequency spectrum, peak factor, repetitiveness and its relationship to other noise to which it is added. Annoyance is defined as the response of a group of people, ranging from compulsion to take legal action to a mild disquietude, the existence of which is revealed only by questioning the subjects.

Level Rank. The region of aural stimulus between normal threshold of hearing and the damage level of the ear may be divided into sub-regions which correspond in level ranking to the degrees of annoyance which a noise continuous in both time and frequency and of variable amplitude can be expected to cause. Such a region is shown in Fig. 23 in which the spectral nature of the noise is assumed to be defined by its octave-band analysis. There are thirteen sub-regions ranked in order from A to M. Noise falling below sub-region A will in general be inaudible while that falling in the region above M will be damaging to the hearing mechanism and certain to elicit strong reactions from the normal subject.

This figure may be used as a working chart for the determination of the probable degree of annoyance of a given noise, the measured or calculated octave-band spectrum being plotted directly on the chart. Additional copies are available from the Acoustics Division, P.M.G. Research Laboratories, Melbourne.

Spectrum Character. If the noise under consideration contains pure-tones or single-frequency components it is likely to be more annoying than a continuous noise that does not have these features. The level rank of a noise is therefore taken as one higher than the rank reached by the loudest single-frequency components in any octave band given in Fig. 23.

The presence of pure-tones in an already existing noise is readily determined by listening to it. If however, calculations of annoyance are being carried out in advance of installation of the noise-generating equipment or of noise - abatement measures, another method of evaluating the significance of pure-tone components is needed.

By definition, a pure-tone becomes inaudible when its sound pressure level is equal to the spectrum level of a continuous noise occupying the related critical band. The critical bandwidths at the centre frequencies of the standardised octave bands are given in Table 9. It can be assumed, without serious error, that the critical bandwidth is approximately constant within a given octave band.

Table 9
Critical Bandwidths

Freq. c/s	Critical Bandwidth 10 logΔf (db)	Δf (c/s)	Correction to octave band level
53	18	70	0
106	18	60	0
212	15.5	37	6
424	15.5	36	9
848	16	44	10
1696	18	62	13
3392	20.5	112	13
6784	23	190	14

*Summarised on pp. 421-423 of this reference.

may be other conditioned reflexes,

In the estimation of sound reduction, particular note should be taken of pure-tone components. The level of the critical bands about these components may then be determined directly from the noise spectrum if this is in one-cycle bandwidths or by the use of the corrections (which must be subtracted) in Table 9 if the spectrum is in octave bandwidths. Pure-tone components significantly greater than the critical band levels, that is, by 3 to 6 db, should be considered as having increased annoyance and be dealt with as outlined above.

Peak Factor. The peak factor of a waveform may be defined as the ratio of the instantaneous, maximum value of the wave to some short-term parameter of the wave as a whole, such as the average or r.m.s. value. It is difficult to define the point at which a noise ceases to be recognised as impulsive; the noise of an automobile back-firing is undoubtedly impulsive but the noise generated by an internal combustion engine with an open or faulty exhaust might be a border-line case. It should also be borne in mind that after treatment a noise may change in character. For example, an unshielded internal combustion engine generates a noise of a generalised broad band character but, after treatment, most of the sound energy may be suppressed, leaving the engine exhaust noise as the most apparent. The noise may then appear to be impulsive although of relatively low peak factor. If a noise is judged to be impulsive the level rank is increased by one.

Repetitive Character. By repetitive character is meant the repetition rate of sounds of comparatively short duration. The community reaction with regard to the repetitive value of a noise is closely associated with the fear or startle aspects, and the field data required to establish a firm criterion is not as yet available. The figures of Table 10 are therefore tentative and are based on an exposure time of about 20 seconds. The correction number given in the table is to be applied to the level rank indicated in Fig. 23. For noise of greatly different duration from 20 seconds, judgment is necessary in the selection of an appropriate correction number.

Table 10

Exposure of 20 seconds duration	Level Rank corrections
1 per minute (taken as continuous)	0
10 to 60 per hour	−1
1 to 10 per hour	−2
4 to 20 per day	−3
1 to 4 per day	−4
1 per day or less	−5

Background Noise. The existence of background noise is of importance for two reasons. First, the intruding noise may be partially masked (in frequency or time or both) by the background noise. Second, the latter provides a reference level against which a subjective assessment of the intruding noise may be measured. It is desirable therefore to have detailed knowledge of the background noise, in particular, the octave-band levels. These may be plotted on the chart of Fig. 25 and a correction number for the noise obtained.

The correction number is the zone in which the maximum band pressure level lies. In some instances, as discussed below, the correction number is one more than that found by this procedure.

Due regard should be paid to the probability of other sources of noise being introduced into the locality at some future time. If this were to occur, the noise under present consideration would then be regarded as part of the background. The background level is therefore likely to increase at a progressively greater rate as more sources of noise begin to operate in the locality. If consideration of future development of the locality indicates that other sources of noise will be introduced it is advisable to increase the correction number by one unit. It should not be assumed that such noise sources will provide masking of the existing noise.

If an octave band analysis of the background noise in a neighbourhood cannot be obtained, the criteria for the neighbourhood noise factor given in Fig. 24 may be used.

Time of Day. The level ranking of noise given in Fig. 23 is based on night-time operation, when noise is likely to be more disturbing psychologically than if the source is operating only during the day. In the latter case a correction of −1 may be made in the level rank. This correction is made independently of the change in sound pressure levels due to altered meteorological conditions, which must be included in the estimation of the probable band source pressure levels prior to using the annoyance criteria.

Adjustment to Exposure. If the introduced noise is similar in character to the existing noise, then the neighbour-

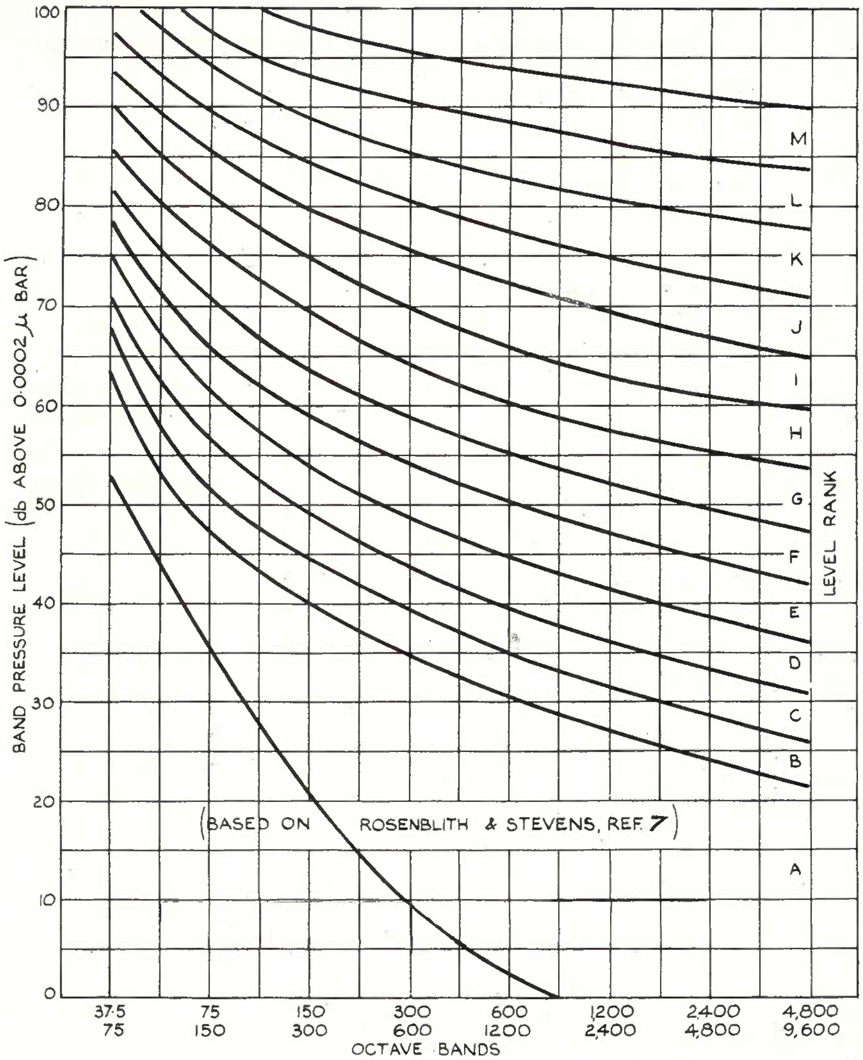

Fig. 23 — Level Rank for Residential Noise.

FOR	CONTINUOUS NOISE SPECTRUM :	USE CORRECTION NUMBER
	PURE TONE COMPONENTS :	ADD 1 TO CORRECTION NUMBER
	IMPULSIVE NOISE :	ADD 1 " " "
	DAY TIME OPERATION ONLY :	SUBTRACT 1 FROM " "
	SOME PREVIOUS ADJUSTMENT TO SAME NOISE :	SUBTRACT 1 " " "

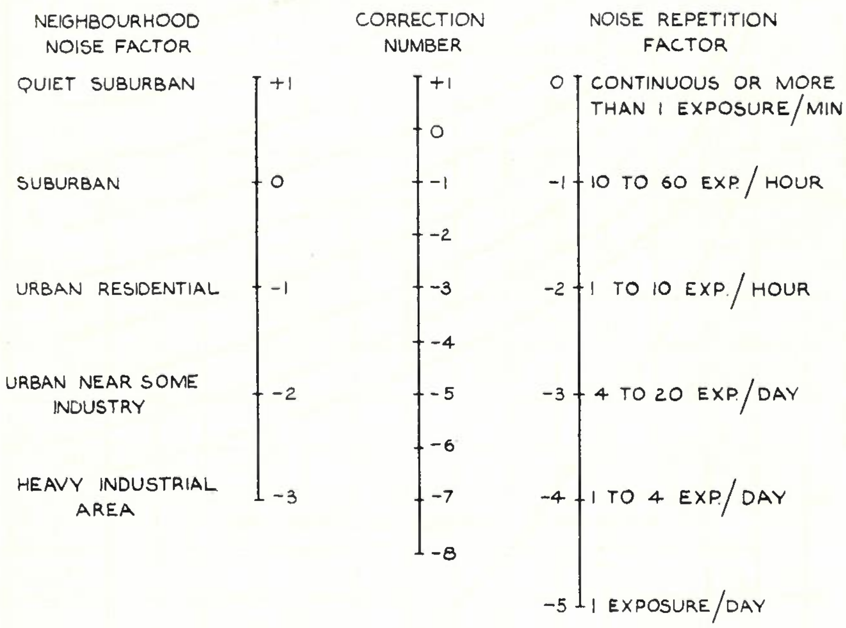

Fig. 24 — Adjustment of Level Rank for Varied Conditions.

hood community will have already made some adjustment to noise of this character and the initial reaction to the new source will be less strong than otherwise. A correction of −1 may therefore be allowed on these grounds. It is appropriate to mention here that the reaction of the community will be influenced by the standing of the originator of the noise in the esteem of the residents, and, as mentioned earlier, much can be done to enhance this in residential areas by the exchange staff ascertaining the times during the day when noise would be least objectionable.

Summary

The factors which together provide a correction number for use with the level rank chart (Fig. 23) are summarised in Fig. 24, which includes a nomogram for summarising the neighbourhood noise and noise repetition factors. Interpolation between whole numbers is possible with the nomogram and should be resorted to whenever possible since the additive process may cause errors also to add.

A positive correction number means that the ranking of the noise is increased, that is, from A to B, etc., and a negative correction, conversely, reduces the noise ranking. The result, after inclusion of the correction number, is the noise rating.

The noise rating corresponds to the level rank if

(1) the noise has a continuous spectrum,
(2) it is not impulsive,
(3) it is continuous in time,
(4) it is present at night-time,
(5) the environment is an average suburb, and
(6) the community is unused to noise of this nature.

Annoyance Ranking

The response of the community to noise may be rank ordered and, as indicated in Fig. 26, ranges from no annoyance to a strong degree of annoyance expressed in vigorous legal action. As in all rank ordering, it is not implied that there is a quantitative scale of annoyance, nor is it possible to interpolate between items ordered in this way.

Over the central part of the curve, a change of one unit in the noise rating causes a change, on the average, of annoyance expressed in mild complaints and in strong complaints. At the extremes of the curve a change of two or three units is required to establish a definite change in responses. At the upper end of the curve, legal action is unlikely to be taken unless the noise rating is 1 or greater. This means people are loath to take legal action with regard to noise since this is usually of a private nature taken on the grounds of damage to property or health. However, by legislative enactment, such as municipal by-laws, this obstacle may be alleviated and legal proceedings may become more common.

The range of expected response is indicative of the likelihood of stronger or milder response than the average to

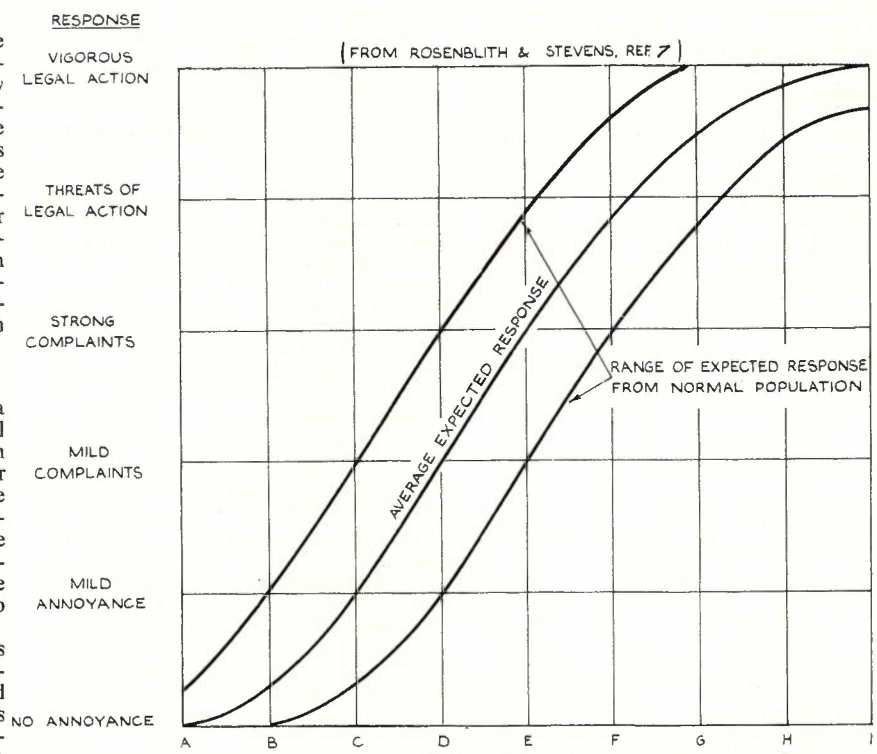

Fig. 26 — Predicted Response to Estimated Noise Rating.

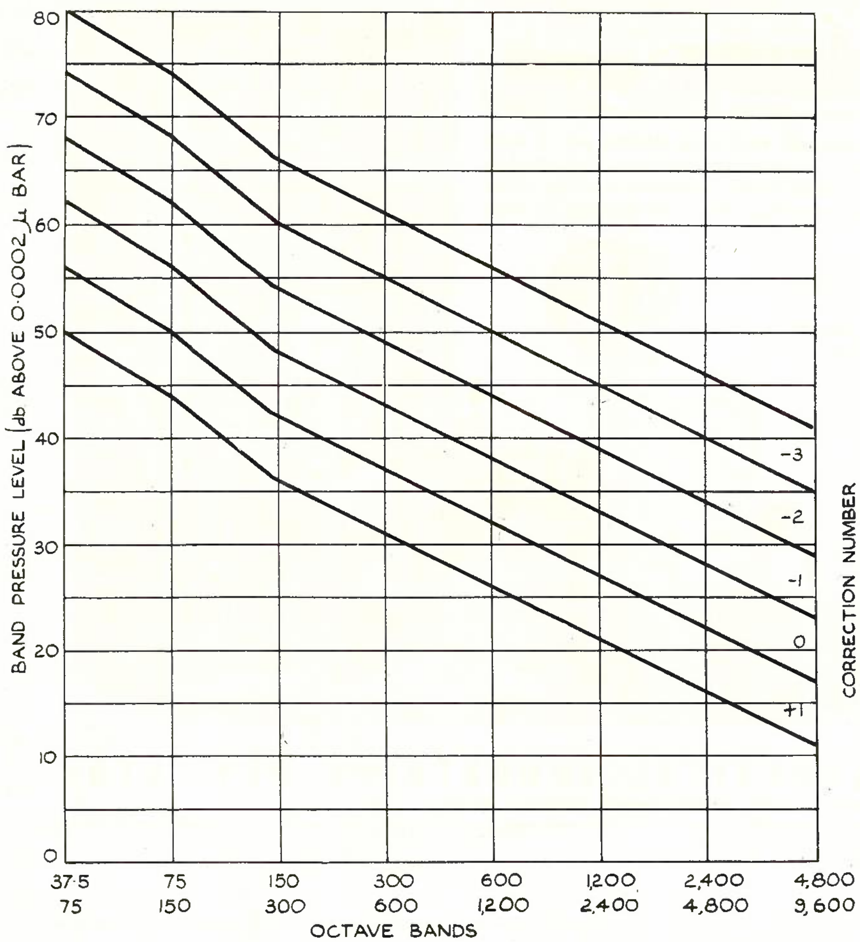

Fig. 25 — Correction Numbers for Neighbourhood Noise.

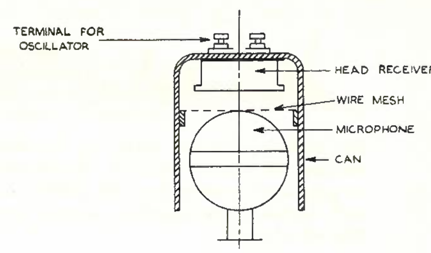

Fig. 27 — Microphone Calibrator.

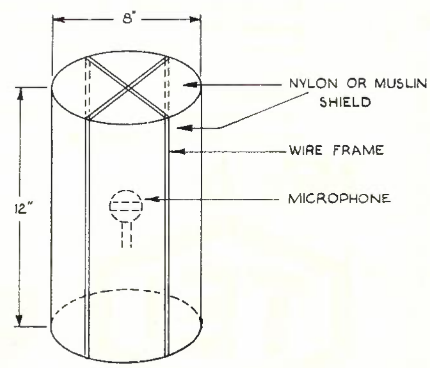

Fig. 28 — Microphone Wind-Shield.

a noise of given rating and is taken from plus one to minus one standard deviation. Thus 15% of the population may give a stronger response than is indicated by the upper range and 15% may give a weaker response than that indicated by the lower range.

CONTINUATION

The next part of this series will be the concluding one and will deal with the systematic design of noise suppression installations, using as examples those recently completed at Elwood and South Melbourne exchanges, Victoria, and Narrabeen, New South Wales.

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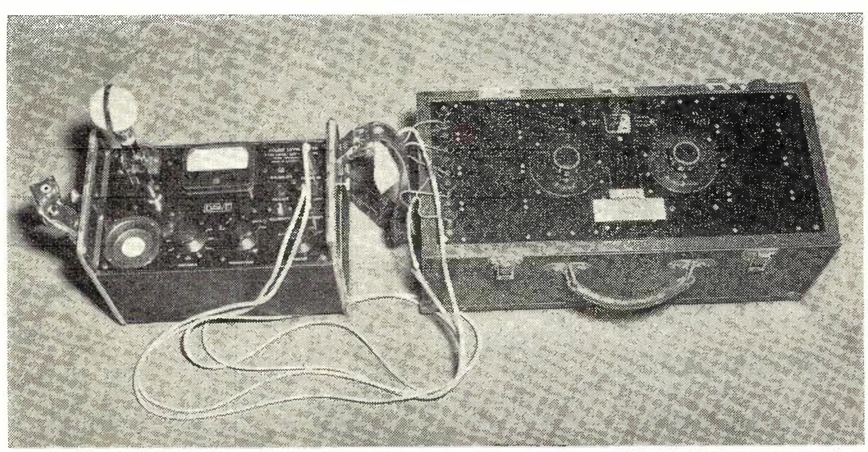

Fig. 29 — Sound Level Meter and Octave Filter.

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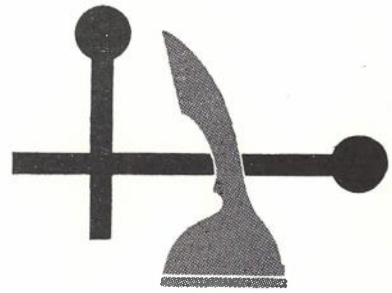

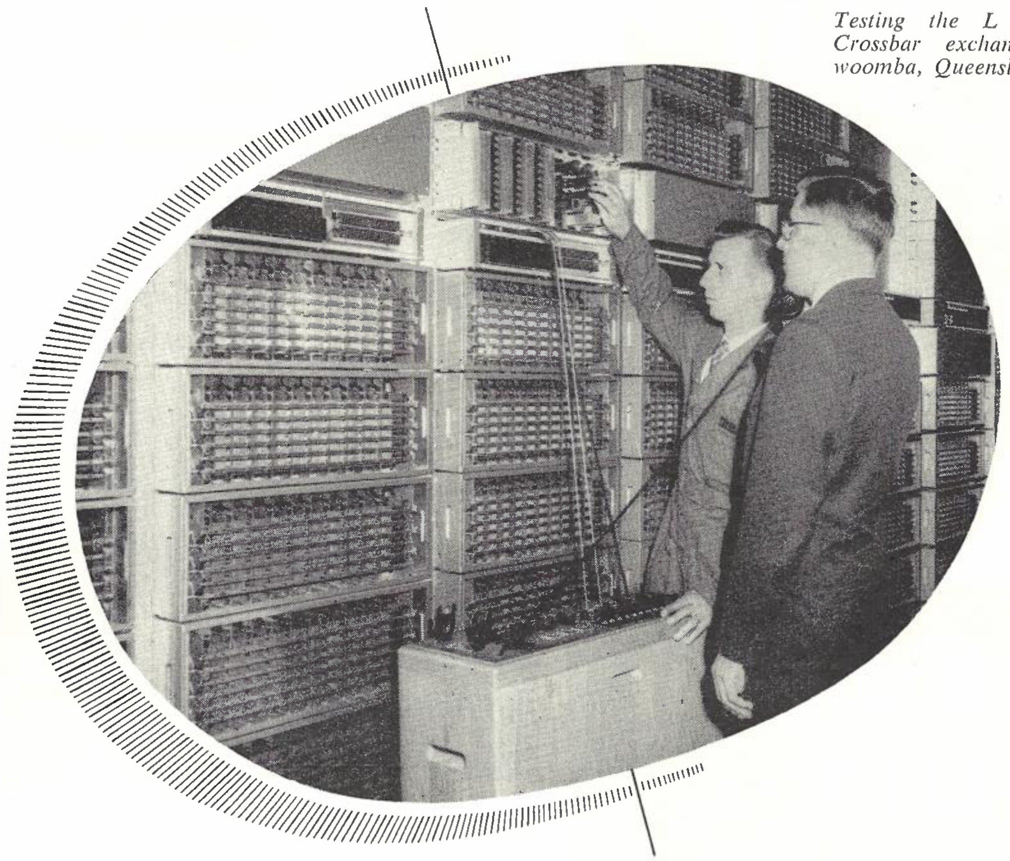

Testing the L M Ericsson Crossbar exchange at Toowoomba, Queensland.