# SUPPLEMENT

## TO THE POST OFFICE ELECTRICAL ENGINEERS' JOURNAL

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## Contents



## CITY AND GUILDS OF LONDON INSTITUTE EXAMINATIONS, 1967-68

# QUESTIONS AND ANSWERS

**Answers are occasionally omitted or reference is made to earlier Supplements in which questions of substantially the same form, together with the answers, have been published. Some answers contain more detail than would be expected from candidates under examination conditions.**

### TELEPHONY C, 1968

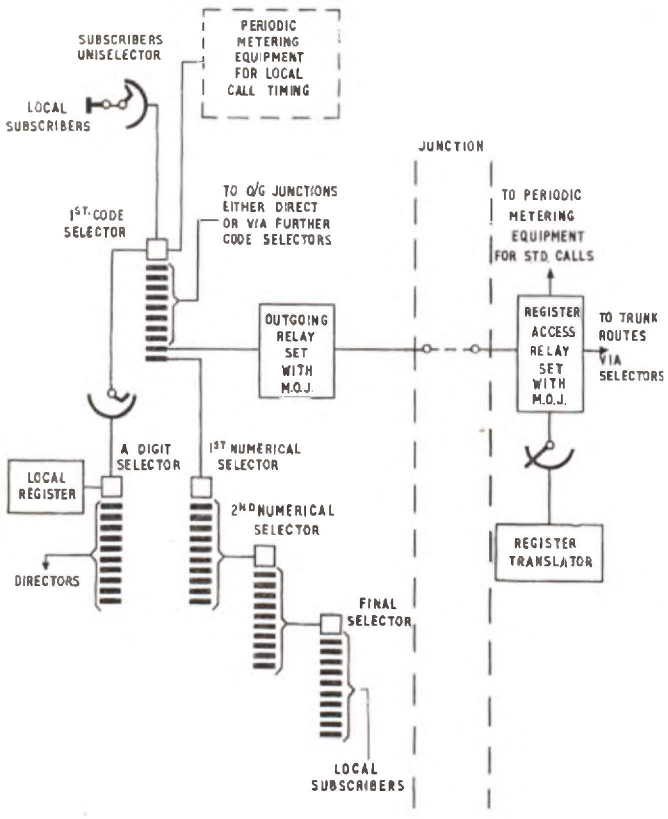
**Students were expected to answer any six questions**

Q. 1. *Outline the arrangements made for controlling the meter of a subscriber on a director exchange when making*

(*a*) *a call to a subscriber on the same exchange,*
(*b*) *an S.T.D. call.*

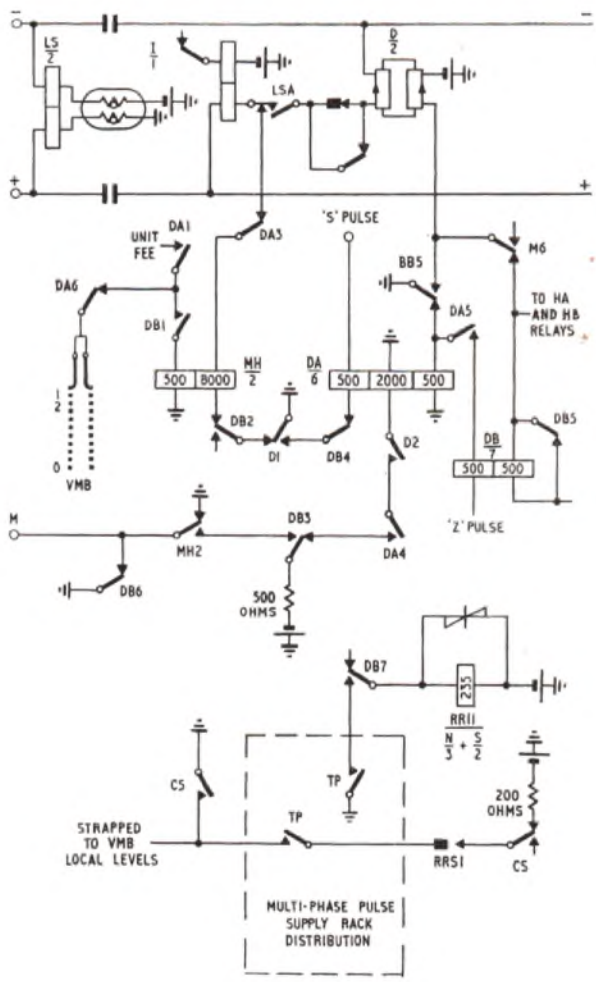
*Describe carefully where the meter pulses are applied to the connexion and the path by which the pulses operate the subscriber's meter.*

A. 1. Sketch (*a*) illustrates the trunking of a call from subscriber to




(*a*)

subscriber on a director exchange and also an s.t.d. call out of that exchange. On a local dialled call, the metering is initiated by the answer signal, brought about by the operation of the ring-trip relay F in the final selector, when the called subscriber answers. This signal initiates a reversal of potentials on the negative and positive wires at the input to the final selector, which is passed back through the second and first numerical selectors to operate the D relay in the first code selector. Sketch (*b*) shows the relevant circuitry.




(*b*)

Relay D, operating, prepares relay DA to operate on the first S-pulse via contact D1, and also to hold during the S-pulse and the Z-pulse, via contact D2.

On the first S-pulse, relay DA operates and prepares relay MH for operation on the unit-fee pulse via contact DA1. Contact DA4 holds relay DA between the S-pulse and the Z-pulse and contact DA5 prepares the operate circuit for relay DB.

When the Z-pulse occurs, relay DB operates and relay DA holds in series with it.

Relay DB, in operating, causes relay MH to operate via contact DB1 to the unit-fee pulse. Contact DB3 prepares the M-wire for the connexion of the meter-pulse and disconnects the hold coil of relay DA, which now holds in series with relay DB to the Z-pulse. Contact DB4 disconnects the operate circuit of relay DA and contact DB5 short circuits the coil of relay DB, enabling it to hold in series with relay HA or HB (not shown). Contact DB7 provides a path to allow the TP earth pulses to operate the ratchet relay RR.

When relay MH operates, a negative-battery meter-pulse is connected to the M-wire via contact MH2 and thence via the M-arc of the uniselector to operate the meter. The meter pulse is transmitted during the period of the Z-pulse.

On termination of the Z-pulse, relay DA releases and contact DA1 disconnects the operate circuit of relay MH. Contact DA6 prepares the operating circuit of relay MH for subsequent metering.

The method of subsequent metering operations depends upon whether the call is a dialled local call or a dialled trunk call.

*Dialled local call*

On levels of the first code-selector to be used for local calls, the appropriate terminal of the vertical marking bank (VMB) is strapped to the TP contact, as shown in sketch (*b*). After the first metering cycle and the operation of relay MH to the first unit-fee meter pulse, relay DA is normal and relay DB is operated; hence each operation of contact TP pulses the ratchet relay via contact DB7. After the tenth TP pulse, the S-springs of relay RR operates when contact TP releases.

Contact RRS1 prepares the operate circuit for relay MH. On the next TP pulse, relay MH operates and contact MH2 connects a negative battery pulse to the M-wire via contact DB3 to operate the subscriber's meter. Simultaneously the ratchet-relay magnet operates.

When contact TP releases, relay MH releases, terminating the meter pulse on the M-wire. The ratchet relay releases to step to the next contact. The S-springs of relay RR releases and contact RRS1 prevents the re-operation of relay MH, and therefore pulsing of the subscriber's meter, until the ratchet relay has counted the next local metering period.

*S.t.d. call*

On levels of the first code selector to be used for routing s.t.d. calls to the s.t.d. register-translator centre, the appropriate terminal of the VMB is connected to the negative wiper of the appropriate bank. Meter pulses received from the register-translator access relay-set operate the MH relay via the VMB, contact DA6 released and contact DB1 operated. Metering-over-junction (M.O.J.) relay-sets are where the centre is remote from the local exchange. Each meter pulse from the s.t.d. equipment operates relay MH and hence the subscriber's meter.

*Q.* 2. *Draw a trunking diagram for a U.A.X. having a capacity for 800 lines and typical junction routes.*

*Explain how a call is set up from a subscriber on the exchange to*

(*a*) *the local parent manual-board,*
(*b*) *a subscriber on an adjacent U.A.X.*

*Describe the junction signalling arrangements used when both calls to the operator and dial calls are passed over a single group of junctions.*

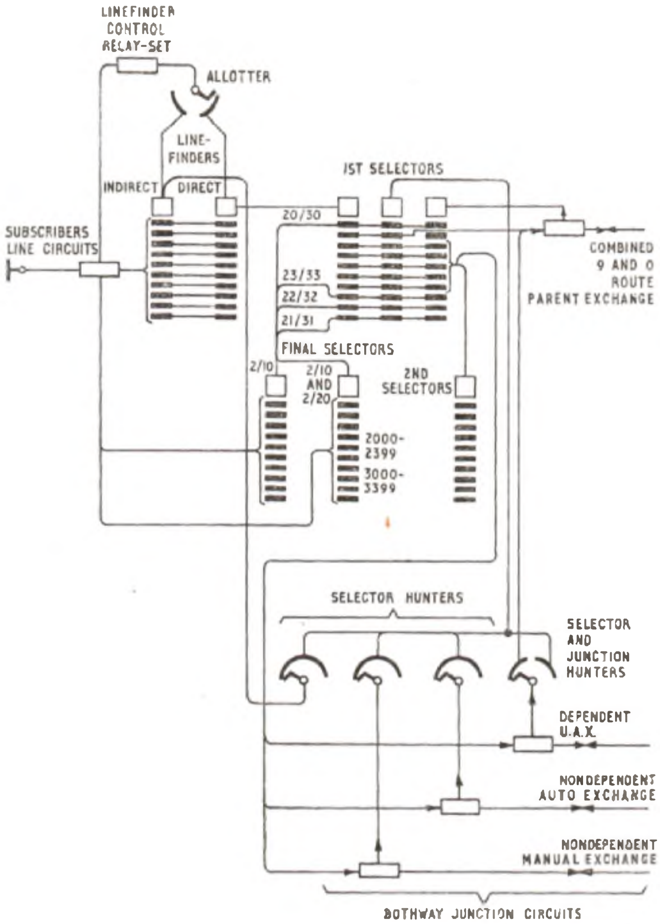
A. 2. The sketch shows a simple trunking diagram for a U.A.X. No. 14 equipped for a capacity of 800 subscribers' lines.

(*a*) *Call to the local parent manual-board*

The calling subscriber's telephone loop initiates a start signal to the linefinder control relay-set, and a free linefinder extends the call to the associated first selector. The junctions giving access to the parent exchange are terminated on level 9 of the first selectors. To obtain the parent manual-board, the digit 0 is dialled and when the first selector wipers reach level 9 the selector vertical magnet is disconnected, by the operation of NP springs. The tenth dial pulse operates a discriminating relay to indicate that the call is to a manual-board. The first selector hunts for a free junction on level 9 and switches the call to a junction relay-set. An appropriate signal is extended by the U.A.X. junction relay-set to the parent exchange junction relay-set, which switches the call to the manual-board. Ring tone is now returned to the caller. When the operator inserts a plug into the answering jack, ring tone is disconnected and the speech-path established.

(*b*) *Call to an adjacent U.A.X.*

The sketch shows that access to an adjacent U.A.X. may be obtained via a first selector by dialling a single routing digit or alternatively, via first and second selectors by dialling two routing digits. The code digit(s) required is determined by individual local requirements. Having dialled the routing digit(s), the caller is connected to a junction relay-set which extends a loop over the junction to seize the distant junction relay-set; the associated selector hunter of which extends the call to an incoming selector. The remaining digits are repeated to the distant U.A.X. by the originating U.A.X. junction relay-set. These digits are also repeated to the associated route-discriminating equipment (or route-restricting equipment where the distant U.A.X. accepts terminal traffic only). When the called subscriber answers, ring tone is disconnected and the junction line potentials are reversed at the distant U.A.X. to establish the speech transmission path and also to initiate a meter-pulse to the caller's meter by the junction relay-set at the originating U.A.X.




Where the call is from an ordinary subscriber to the parent automatic equipment, a loop is extended to the U.A.X. junction relay-set by the preceding group selector after the prefix 9 has been dialled. The U.A.X. junction relay-set extends a loop over the junction to the parent exchange equipment to seize the associated incoming selector in readiness for the next dialled digit. For a coin-box subscriber originating the same type of call, the circuit operation is similar, except that additionally a 150-ohm negative battery is extended from the group selector, via the M-wire, after switching to the U.A.X. junction relay-set has taken place. This signal ensures that codes dialled which are not permitted to coin-box subscribers will be barred by the route discriminating equipment associated with the U.A.X. junction relay-set.

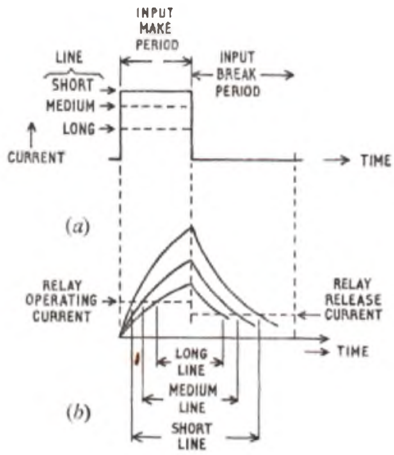
When digit 0 is dialled for a call to the parent manual-board the group selector is stepped to level 9 only, the last dialled pulse operating a discriminating-relay which results, in the case of an ordinary call, in the connexion of a 2,000-ohm negative battery to the M-wire before switching to a free junction outlet takes place. On switching to a free outlet, a discriminating relay is operated in the U.A.X. junction relay-set which disconnects the normal loop calling condition and connects a negative battery to the junction positive line wire. As a result the parent exchange equipment is seized and the call is switched to the parent manual-board, where the circuit of an ordinary answering lamp is completed. For a call to the parent manual-board by a coin-box, seizure of the U.A.X. junction relay-set is the same as before except that the 2,000-ohm negative battery applied to the M-wire is replaced by a 150-ohm negative battery to operate an additional discriminating-relay in the U.A.X. junction relay-set. The calling signal to the parent exchange is now an earth connected to the junction positive line wire which lights a coin-box answering lamp at the parent manual-board.

*Q.* 3. *What factors cause distortion of loop-disconnect d.c. pulses when they are used for setting up dialled calls over a long amplified junction provided on audio cable?*

*Sketch and describe the circuit arrangements which could be used to give satisfactory d.c. pulsing and signalling over such a junction. State the approximate length of line permissible if 20 lb/mile cable is used.*

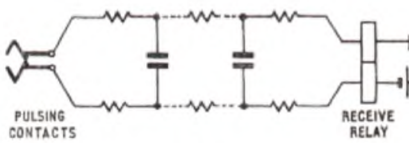
A. 3. The characteristics of loop-disconnect pulses, when transmitted over a long amplified junction via audio cable, are affected by the primary constants of the derived signalling circuit. The phantom circuit would be used for d.c. pulsing, the effective loop-resistance being half, and the line capacitance twice, that of a 2-wire circuit of the same length. In addition, the capacitance will be increased by capacitors necessary in the 2-wire/4-wire terminations for converting from 2-wire to phantom pulsing.

If the signalling circuit were purely resistive, i.e. with a shunt capacitance, series inductance and leakance of zero, then if a purely resistive detecting device were used at the receiving end, no distortion would be produced. Increased resistance would reduce the magnitude of the current pulses (sketch (*a*)).



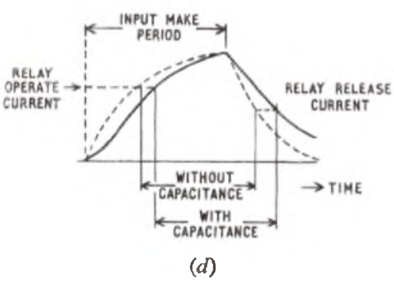

With an inductive receiving relay, however, the growth and decay curves of the received current are as shown in sketch (*b*). An increase in resistance will reduce the make period of the receive relay. The effective distortion will be determined by the values of the operate and release currents for the receive relay. For a particular length of line the operate and release delays can be equal, giving no distortion.

With line capacitance present, distributed as in sketch (*c*), the line




(*c*)

will be charged to battery voltage when the pulsing contacts are open. On closing the contacts, the line will discharge through the pulsing contacts, slightly increasing the delay in the build-up of the relay operate current (sketch (*d*)). On opening the contacts, the relay




(*d*)

current will continue to charge the line, delaying the release of the relay. The effect is to increase the make period of the relay due to capacitance.

The growth and decay effects are further modified by the line inductance and leakance, but under practical conditions the effects of resistance and capacitance are predominant. As the length of line is increased, the line capacitance effect will eventually result in failure to release of the receive-relay during pulsing.

A typical value of limiting line length for loop-disconnect pulsing over the phantom circuit of a 4-wire amplified circuit is 40 miles of 20 lb/mile cable. For amplified junctions of greater length than this, a form of double-current signalling described in A.5, Telephony C, 1960, *Supplement*, Vol. 54, p. 25, July 1961, is used. Such a system gives satisfactory operation over the phantom of up to 100 miles of 20 lb/mile cable.

*Q. 4. Describe, with sketches of the circuit elements concerned, the operation of a 1st Code Selector from seizure to the beginning of vertical stepping. Include a description of the circuit operation when the subscriber commences dialling before receipt of dial tone.*

*Q. 5. What factors determine the choice of signalling frequency to be used for a 1-V.F. in-band signalling system?*

*Give a circuit sketch of a typical 1-V.F. receiver and describe the operation.*

*Over what range of signal level would you expect the receiver to operate? Give reasons.*

A. 5. The following factors must be taken into account when determining the signalling frequency of a 1-V.F. in-band signalling system:

(*i*) the attenuation-frequency characteristic of the circuit over which the system is to operate; frequencies at the upper or lower edges of the band may suffer such attenuation as to require too great a receiver sensitivity,

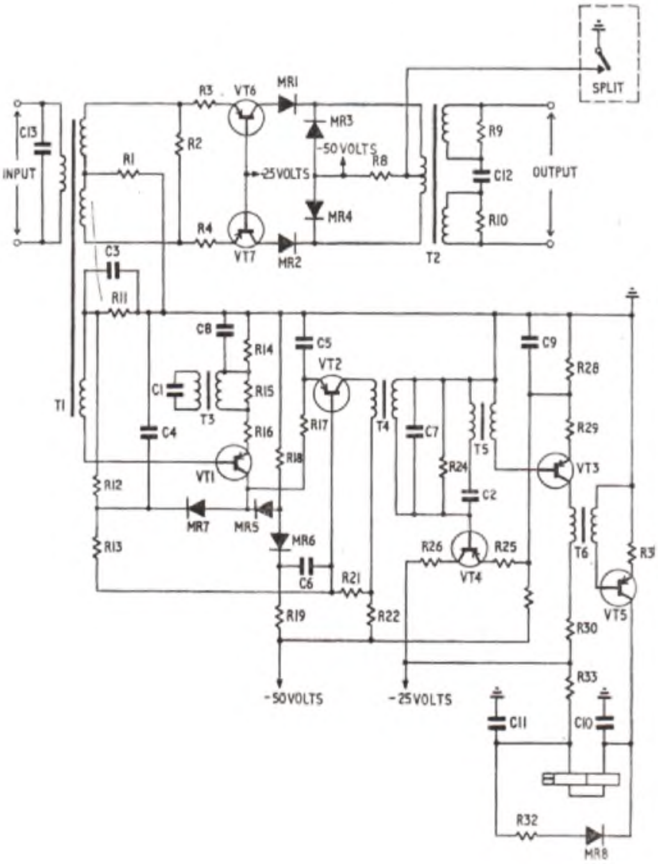
(*ii*) signal-energy content of speech; since the system is in-band, the receivers must operate in the face of speech, and be capable of differentiating between a genuine signal and a speech imitation. Extensive tests have shown that speech imitation problems can be minimized by employing frequencies above 2 kHz for signalling.

(*iii*) Power handling capacity of line plant; due to crosstalk and other considerations, the frequency employed must be related to the desired power level.

(*iv*) Noise present on the line; the choice of signalling frequency affects the band available for speech guard purposes, when using a tuned-guard circuit. The guard band should exclude regions of high noise level.

(*v*) Presence of existing V.F. signals in the network used for other services, e.g. data signals; as in-band systems must continuously monitor during a call in order to detect a genuine clear signal, the possibility of energization of the signal circuit from other services must be minimized.

Sketch (*a*) shows, in simplified form, the circuit of a transistorized




(*a*)

1-V.F. receiver (2,280 Hz). The receiver is combined with a buffer amplifier (VT6 and VT7) for insertion in the receive pair of a 4-wire circuit.

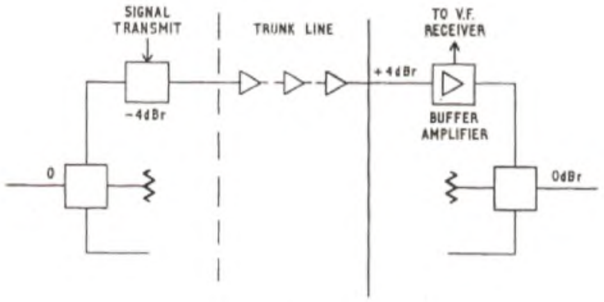
The input to the receiver, via transformer T1, is applied to the base of transistor VT1, acting as an amplifier of sufficiently high input-impedance, partly determined by resistor R6, to give the required impedance to the input terminals (600 ohms). The transformer T3, tuned by capacitor C1, introduces de-emphasis at the signal frequency in the output from transistor VT1 to achieve a suitable guard coefficient. At all levels in the signalling range, the output of transistor VT1 is clipped by the rectifier arrangement to give trapezoidal collector-current waves of constant amplitude down to the minimum input level. If the collector-current of transistor VT1 increases by more than a constant value, rectifier MR5 will be backed off and the current will flow into the base of transistor VT2. Similarly, if the collector-current decreases by more than this constant value, rectifier MR6 will be backed off and the current will flow from the base of transistor VT2 via capacitor C6.

The square-wave constant-current output from transistor VT1 is fed to the current amplifier VT2. In the collector circuit of transistor VT2, transformer T4 is tuned by capacitor C7 to the signal frequency, the output at this frequency being applied to the base of transistor VT3, resulting in a maximum current in transformer T6. At frequencies away from the signal, the guard voltage across resistor R24 is applied to the base of transistor VT4, which, acting as a rectifier, charges capacitor C9 to the peak guard-voltage, opposing the action of transistor VT3. Thus, the winding of transformer T6, in the transistor VT3 collector circuit, has a peak current at signal frequency, but this is reduced by the presence of guard frequencies; hence the speech guard action.

The rectifying action of transistor VT5 produces operation of the high-speed relay HS. Capacitors C10 and C11 are used to give equal operate and release lags to the relay for distortionless pulsing. Rectifier MR8 and resistor R32 prevent excessive voltage spikes damaging transistor VT5 at cut off.

In the buffer amplifier circuit, rectifiers MR1 and MR2 are normally forward-biassed and low impedance, and rectifiers MR3 and MR4 are back-biassed giving a high shunt-impedance. In this state the attenuation from input to output is low. When the splitting earth is applied, rectifiers MR1 and MR2 are back-biassed giving a high series impedance, and attenuation through the buffer amplifier of approximately 60 db.

Sketch (*b*) illustrates the relative levels of the signal transmit and




(*a*)

receive points for a typical application, i.e. a nominal zero-loss trunk circuit. The relevant factors in determining the levels of sensitivity for the V.F. receiver are as follows:

Maximum sending level permitted = −6 dBm0, (actual level = −10 dBm).

Permitted variation in sending level = ±0·5 dBm0.

Loss variation with time encountered on 99·9 per cent of connexions is ± 6 dB (3$\theta$ where $\theta$ = 2 dB, the standard deviation of loss).

Additional loss at 2,280 Hz due to attenuation-frequency distortion of line is 3 dB.

This gives a received level of signal at receiver of 0·5 dBm0 maximum and −15·5 dBm0 minimum.

Since the receiver is attached at a +4 dBr relative level point, actual level at the receiver input is + 4·5 dBm maximum, or −11·5 dBm minimum, i.e. 15 dB range of sensitivity.

In practice, the same design receiver may be required to operate on trunk circuits of different overall loss, e.g. −3 dB overall loss. This would increase the spread of sensitivity required to 18 dB.

*Q.* 6. *Describe what is meant by a switching matrix in a space-division type of telephone exchange.*

*A 5,000-line exchange employs crossbar selectors for the speech-path switching. Explain, with a trunking diagram, how a free speech-path is selected when setting up a call through the exchange.*

*What are the factors which limit the time available for the marking process?*

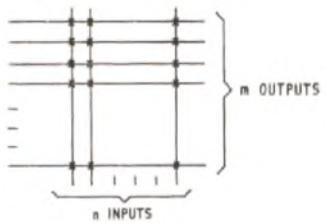
A. 6. In an automatic telephone-exchange, a both-way speech-path must be set up between a calling line and a called line. In a space-division type of exchange, this may be achieved by a number of switching stages, each consisting of speech-path contacts arranged in a co-ordinate manner (sketch (*a*)). The contacts may be relay type or electronic elements. Each line represents a speech path, the symbol X denoting a number of crosspoints which operate simultaneously. Such an "n–m" switch forms a switching-matrix.

Sketch (*b*) shows the trunking of a typical 5,000-line exchange of the ATE 5005A crossbar type. Consider a call from one subscriber to another on the same exchange. The call is set up in two steps as follows:

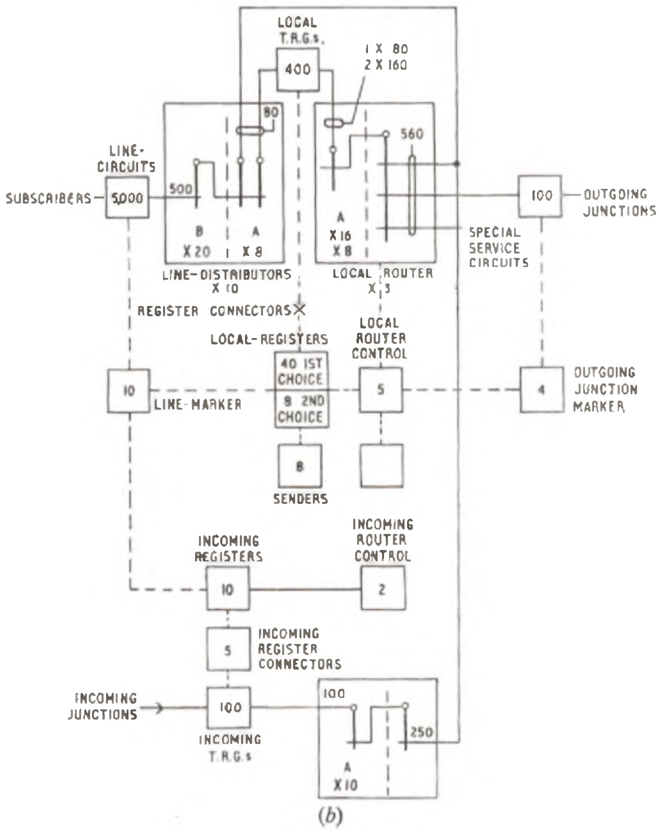
(*a*) The calling subscriber lifts his receiver, operating the LR relay in his line circuit. The A contact of relay LR marks the outlet of the B switch, and all the free links to A switches in the 500-line distributor are marked. One of the free transmission relay-groups (T.R.G.) connected to the A switches is seized, together with one of the free registers to which the T.R.G. has access, and a connexion established through the switches using the self-steering principle. At the time of seizing, one of four possible classes of caller is signalled to the register and stored. Dial tone is returned to the caller and the subscriber is free to dial.

(*b*) After sufficient information is received, the register applies to the router-control for permission to route the call. Permission having been granted, the register then applies to the appropriate 500-line marker to mark the wanted subscriber's line. The register thus gains access to the marking wire of the wanted subscriber, and encounters one of four possible conditions, i.e. subscriber free, subscriber busy, dead-number or disconnexion, the latter indicating a fault condition.

If a subscriber-free condition is found, the B-switch outlet serving the required subscriber is marked, and the mark is passed back via all the free links to the A switches in his distributor, and then by route-switches B to the route-switch A serving the T.R.G. on which the call is waiting to be set up. The arrival of the mark in route-switch A causes a signal to be sent to the router-control to indicate that a complete free-path has been selected. The router-control now takes action to operate the bridge-magnet in route-switch A. This extends the call to route-switch B (according to which bridge-magnet was operated in route-switch A) and so on through the distributor-switches A and B to the called subscriber's line. The call set-up is now complete and the called subscriber is rung.




(*a*)




(*b*)

The marking process described in (*a*) is such that only one marking operation can take place at any one time on a 500-line group. Two simultaneous calls will result in one call being delayed for a fraction of a second while the other waits for marking. Thus the rate of arrival of calls in a 500-line group limits the time available for marking.

The marking process described in (*b*) involves both a local-router and a line-marker. The router-controls for each local-router are duplicated, each register having access to either control, choosing at random if both are free. If two calls originate from the same router-switch, or are destined for subscribers in the same 500-line distributor, then one call is delayed. The total holding time of the router-control is 1 to 1·5 seconds. However, if two calls arise in a local-router on different A-switches, destined for subscribers in different 500-line groups, then the marking and switching of each call can be achieved simultaneously.

It follows, therefore, that in the exchange shown, up to six simultaneous calls can be put through the second step of the set-up operation simultaneously.

The factors affecting the time available for marking in the second stage of the set-up procedure are, therefore, the rate at which calls originate, and the extent to which these demands are distributed over non-conflicting marking paths.

*Q.* 7. *What are the properties of a transistor which make it suitable for switching applications in a telephone exchange?*

*Describe, with a circuit sketch, how loop-disconnect pulsing in a 2-wire line may be monitored, with a low-loss circuit using transistors and relays, in order that the dialled numbers may be recorded.*

A. 7. The properties of a transistor of interest for switching applications in a telephone exchange are as follows:

(*a*) It can operate as a current-switch. The change from the off-state to the on-state can be effected by a control current many times smaller than the current being switched.

(*b*) As a current-switch, the operate and release times can be very small compared with electro-mechanical devices.

(*c*) The circuitry requires only low-voltage supplies.

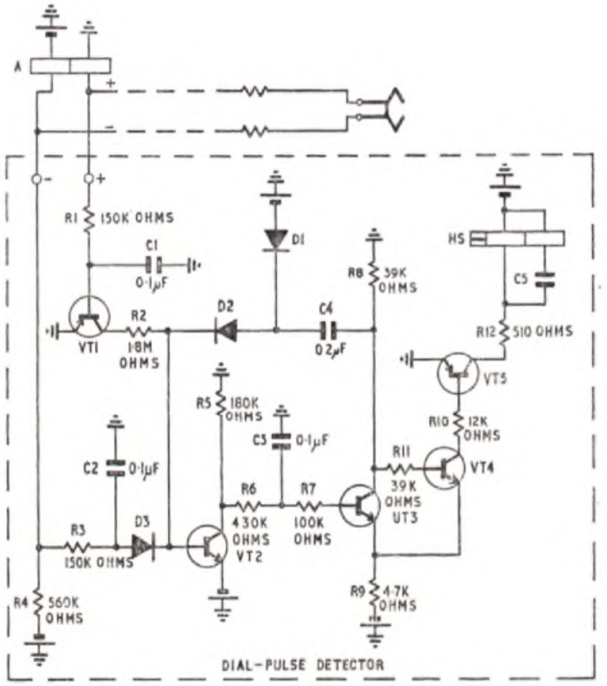
(*d*) It is small.

(*e*) It is efficient. There is no supply of heat required as in hot-valve thermionic devices.

(*f*) No maintenance is required.

(*g*) It has a long life.

The sketch shows the circuit of a transistorized monitoring-device, used for detecting dial-pulses on a 2-wire circuit. The positive and




negative wires of the input are tapped on to appropriate wires of the 2-wire circuit.

When the dial contacts are open, transistors VT1 and VT2 are cut off. Transistors UT3 and VT4 form a Schmitt trigger, the normal state (dial-contacts open) being transistor UT3 conducting and transistor VT4 cut off. With transistor VT4 cut-off, transistor VT5 is also cut off, and the high-speed relay HS is unoperated.

When the dial-contacts are closed, the potential on the positive input lead (−2 volts to −25 volts according to the line loop-resistance) will turn on transistor VT1 via resistor R1. Transistor VT1, in turning on, will turn on transistor VT2 via resistor R2. The fall in potential of the collector of transistor VT2 will, after a delay due to the time-constant of the R6-C3 resistor-capacitor combination (to prevent operation to short-duration spurious-pulses), result in transistor UT3 turning off, i.e. the turn-over of the Schmitt trigger. Transistor VT4 now conducts and its collector-current turns on transistor VT5. In the collector circuit of transistor VT5 is the coil of the high-speed relay HS, which operates.

Thus, when the dial-springs are closed, relay A operates in the 2-wire telephone-circuit, then relay HS in the dial-pulse detector is operated. When the dial springs are open, relay HS releases. Hence relay HS will follow the operation of relay A. The contacts of this relay can be connected to a suitable counting and storing apparatus. The input impedance to the dial-pulse detector is high, resulting in a negligible shunting effect.

In the circuit shown, certain additional features have been incorporated.

The circuit comprising capacitor C4 and diodes D1 and D2, provides a feedback path from the collector of transistor UT3 to the base of transistor VT2. When the trigger switches, feedback provides base-current for transistor VT2, independently of the normal input from transistor VT1, for a period of approximately 35 ms. Hence, relay HS is operated for a minimum period of 35 ms, whatever the duration of input-pulse, providing it is above the minimum operate-value. This can assist in the operation of electro-mechanical equipment connected to the contacts of relay HS.

When monitoring a 2-wire telephone circuit, several possibilities exist of spurious disturbances to the loop condition due to switching of positive and negative wires through selectors. Also, pulsing on short lines may result in an oscillatory condition about the final steady-state potentials of the positive and negative wires. Capacitors C1 and C2 and diode D3 are used to safeguard the operation against these various factors.

When selector switching results in disconnexion of both wires simultaneously for a short period, the negative potential connected to the negative input lead maintains transistor VT2 conducting.

When the negative wire is disconnected momentarily in advance of the positive wire, the positive back-e.m.f. on the negative line momentarily holds transistor VT2 conducting, via diode D3, during this period.

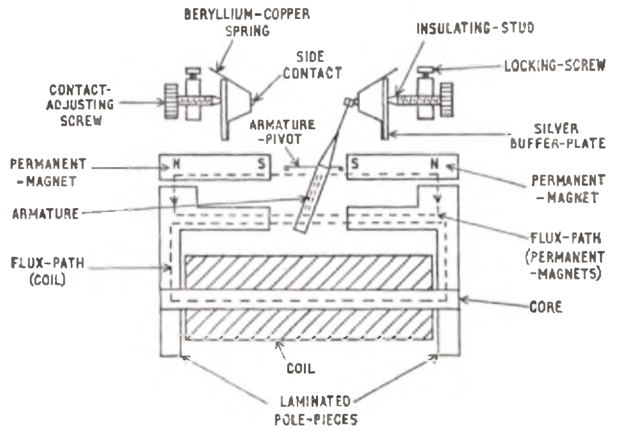
On short line oscillations, when the dial-contacts open, the positive line swings first in a positive direction, charging capacitor C1 and cutting off transistor VT1. Capacitor C1 then discharges slowly through resistor R1, to prevent the negative oscillation turning on transistor VT1 and simulating a dial make.

*Q.* 8. *Describe, with a sketch, the construction and operation of a sensitive polarized-relay for the direct reception of double-current signals in a long-distance d.c. signalling system.*

*What is the contribution of the permanent magnet?*

*Explain how eddy currents in the magnetic circuit would affect the response-time of the relay. How are these effects minimized?*

A. 8. The sketch shows the construction of a Post Office Type 2B




polarized relay.

The relay is both-side stable, the armature being held against either side-contact when the coil is unenergized due to out of balance in the flux from the permanent-magnets. When the coil is energized in such a direction that the force due to the permanent-magnet is overcome, the armature moves across the gap to the opposite side-contact. The force on the armature increases according to a square law during its transit so that the armature accelerates across the gap until it is arrested by the contact assembly. The kinetic energy imparted to the moving armature is absorbed by the beryllium-copper spring which flexes when the contacts come together causing friction between the upper end of the spring and the silver buffer-plate. Thus contact bounce is almost eliminated.

The standard contact material for the Type 2B relay is copper-palladium in the ratio 60 per cent palladium to 40 per cent copper, although platinum contacts may be fitted where heavy usage or onerous switching conditions demand a higher grade material. A contact pressure of 16 grams is required for reliable current switching and, as the force due to the permanent-magnets is insufficient to provide this, a bias-winding is generally employed to provide adequate pressure and to restore the armature, on release, to the correct position when no current is flowing in other windings.

The effect of the permanent-magnets is as follows:

(*i*) The direction of movement of the armature of the polarized-relay is dependent on the direction of the operating current.

(*ii*) The polarized-relay is more sensitive than the unpolarized type, since in the former case the operating force is dependent on the flux density due to the current alone, whereas in the latter case, the operating force is proportional to the product of flux density due to the current times the flux density due to the permanent magnet.

(*iii*) On operation, once the armature has moved past the point equidistant between the magnets, the flux due to the current is reinforced by the flux due to the nearer magnet, the attractive-force is substantially increased and the relay is more reliable.

There are three separate flux paths in the iron circuit of the relay. No flux due to the coil passes through the permanent-magnets. When the relay is operated, the rising current causes eddy-currents to be induced into the core and polepieces. These eddy-currents in turn cause a flux, which by Lenz law opposes the main flux, the operation of the relay is therefore delayed. Similarly, on release of the relay, eddy-currents cause a flux tending to prolong the main flux and the release of the relay is also delayed. Eddy-currents are kept to a minimum by using laminated polepieces and a mu-metal core.

**Q. 9.** *State Erlang's lost-call formula for a full-availability group, giving the meaning of each symbol used.*

*Five erlangs of traffic are offered to a full-availability group during the busy hour. Use the formula to calculate the sum of the traffic carried by the first three trunks.*

**A. 9.** Erlang's lost-call formula for a full-availability group is:

$$B = \frac{\dfrac{A^N}{N!}}{1 + A + \dfrac{A^2}{2!} + \dfrac{A^3}{3!} + \dots \dfrac{A^N}{N!}}$$

where $B$ = proportion of lost calls, i.e. grade of service,
$N$ = number of circuits (or trunks),
$A$ = average traffic offered (erlangs),
$!$ = factorial,
e.g. $N! = N(N-1)(N-2)\dots 3 \times 2 \times 1$.

Using the values given, the proportion of calls lost by the first three trunks is given by

$$B = \frac{\dfrac{5^3}{3!}}{1 + 5 + \dfrac{5^2}{2!} + \dfrac{5^3}{3!}},$$

$$= \frac{\dfrac{125}{6}}{1 + 5 + \dfrac{25}{2} + \dfrac{125}{6}},$$

$$= \frac{125}{6} \times \frac{6}{6 + 30 + 75 + 125},$$

$$= \frac{125}{236}.$$

$\therefore$ Total traffic lost $= 5 \times \frac{125}{236}$ erlangs.

$\therefore$ Traffic carried by first three trunks = traffic offered − traffic lost,

$$= 5 - 5 \times \frac{125}{236} = 5\left[1 - \frac{125}{236}\right],$$

$$= 5 \times \frac{111}{236} = \frac{555}{236},$$

= 2·35 erlangs to 2 places of decimals.

**Q. 10.** *Describe, with the aid of a block diagram, the principal features of an automatic routiner used for testing group selectors in a non-director type of exchange. Explain the nature of the tests which would be applied by the routiner.*

*What are the advantages and disadvantages of routine testing separate items of equipment in an automatic exchange compared with testing the operation of the exchange by means of test calls generated by an artificial traffic machine?*

**A. 10.** The principal elements of the routiner are shown in the sketch.

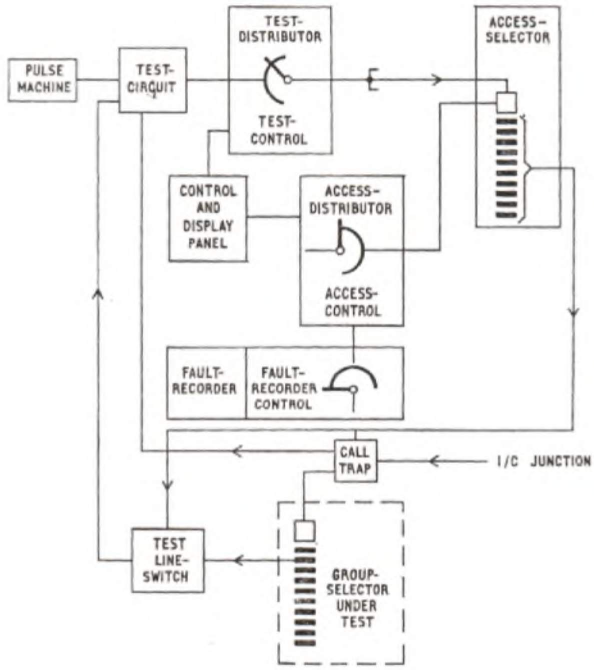
The 2-motion access-selector, serving up to 100 group-selectors, is operated by the routiner-access control, lamps on the display panel indicating which group-selector is under test.

The access-selector also extends the routiner to the test line-switch serving level 9, outlets 19 and 20 of the group-selector under test. This switch diverts the outlets from the normal trunks to the routiner test-circuit to enable the functioning of the selector to be checked.

The call trap-circuit, at the input to each group-selector, ensures that busy tone is given to any caller seizing a junction connected to a group-selector under test.

Each test is applied to the group-selector in sequence by the test-distributor. Each successful test is signalled to step the distributor. A lamp display is given on the control-panel to indicate testing progress.

Failure to step in 3–6 minutes is regarded as a fault condition, and in the case of manual operation gives visual and audible alarms on the control-panel. When operating automatically under clock control, the fault condition results in the association of the fault-recorder and the printing of a fault-docket. After printing, the access is stepped to the next group-selector and the test sequence begun.




The first test applied is the busy test and, provided the group-selector is free, it is then guarded against intrusion from traffic. Tests of polarity, continuity and absence of contact of the incoming positive and negative wires are followed by A-relay performance tests.

The group-selector is then seized and pulsed by non-leaky pulses, at 12 pulses per second with an 80 per cent break ratio, to level 9 and a check made that the first test line (outlet 19) is seized within approximately 600 ms of receipt of the last vertical pulse. This test proves that the selector cuts in correctly on level 9 and that the release lag of relay CD, plus the hunting and switching times, are within the required limits. The P-wire guard is monitored throughout the tests, continuity of the switch path is checked before the selector is released and the release sequence is also checked.

The stepping and switching test is then repeated with leaky pulses at 12 pulses per second with a 50 per cent break ratio and switching to the second test line (level 9, outlet 20) is proved in a similar manner.

The selector is then released, pulsed by leaky pulses to level 9 again and the return of busy-tone checked from the 11th rotary step.

Satisfactory release from this test completes the test program and the test circuit is released. The access-selector is advanced to the next group-selector and the test cycle is re-commenced.

The advantages of routine testing over test calls from an artificial traffic machine are as follows:

(*i*) Since only one item of equipment is involved, e.g. a group-selector, a more precise indication of the position of a fault is given.

(*ii*) The equipment is tested to its design limits, which ensure that it will operate under the most onerous conditions it is expected to meet in practice.

(*iii*) Equipment items are tested in sequence, therefore the application of tests is not dependent on traffic conditions.

(*iv*) All functions can be proved systematically.

The disadvantages of routine testing are as follows:

(*i*) The wiring between equipments is not in the path checked by the routiner, hence faults in the rack wiring can give trouble in practice but are not revealed by the routiner.

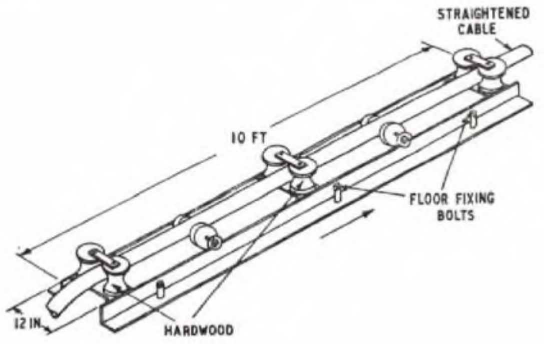
(*ii*) Unnecessary maintenance attention may be given to equipment because it is outside design limits, whereas in practice it may not cause failure due to compensating deviations in associated equipment which is tested separately by the routiner.

(*iii*) The numerous access connexions for the routiners are costly to provide.

**Students were expected to answer any six questions.**

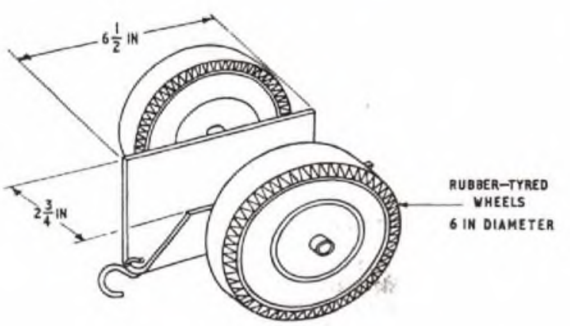
*Q. 1. Describe in detail a method by which two 176-yd lengths of 888-pair, 10-lb, P.C.Q.T., lead-sheathed cable would be installed via the shaft, in a 7-ft, concrete-lined cable tunnel 100 ft below ground. Describe briefly the cable joint which would be used and type of bracket and bearer that would be suitable.*

A. 1. The cable drum should be set up at the shaft head as close to the cable feed-pipe as possible. The cable may be fed into the bell mouth of the cable feed-pipe by hand. A brake fitted to the cable-drum device will be necessary to control the speed of entry of the cable into the pipe, which would otherwise increase as the weight of cable hanging in the pipe increased. As the cable leaves the bottom of the cable feed-pipe, it is passed through a cable guide as shown in sketch (*a*), which




(*a*)

is temporarily fixed to the floor of the tunnel. The passage of the cable through the hardwood spools corrects any sets which may have been introduced as the cable passed down the feed-pipe. When the cable has been straightened, i.e. as it leaves the guide, it is laid on cable trolleys placed at 6-ft intervals. These support it clear of the tunnel floor and allow it easy movement along the tunnel. Sketch (*b*) shows




(*b*)

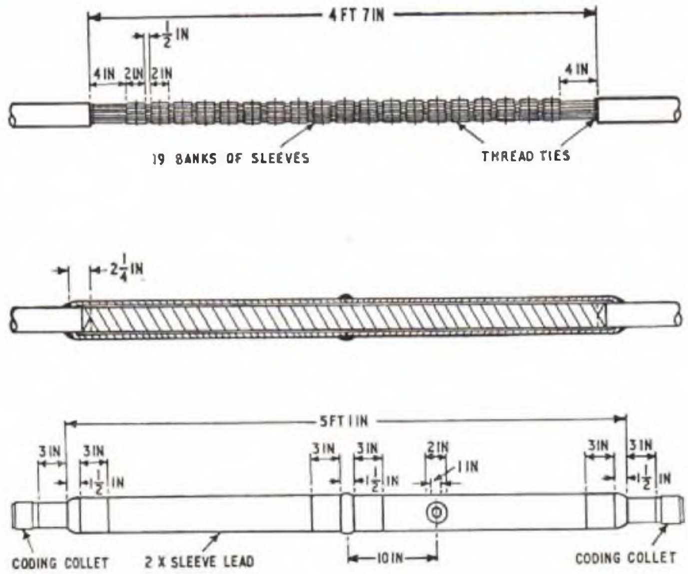
a typical cable trolley. As the cable leaves the shaft, supported on the trolleys, it is dragged manually or mechanically along the tunnel.

If at the end of the route the cable has to be brought to the surface via an exit shaft, the end of the cable is attached to a cable grip which is joined by a rope to a winch at the surface. As the cable is slowly hauled up the shaft, the trolleys are removed until the front end of the cable reaches the surface. The cable is then led to the point where it will be jointed or terminated.

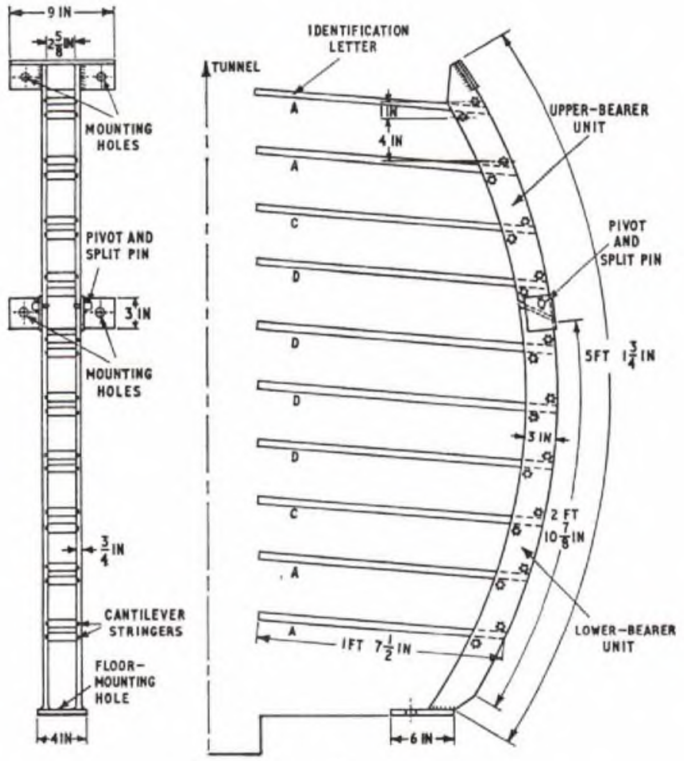
If the cable cannot be removed from the cable feed-pipe it will be necessary to cut the cable, remove it from the pipe and joint it to another length. The new length must be lowered to the foot of the shaft but not via the feed pipe. If a joint is undesirable, the cable can be lowered out of the feed pipe by continuing to draw the cable along the tunnel and out of the exit shaft. When the cable is free of the pipe the cable is drawn back up the shaft, again not via the feed pipe. Trolleys are used to carry the cable along the tunnel during this operation.

When both ends of the cable are located in the shafts, the trolleys are removed, and the cable is placed in its correct position on the cable brackets. The vertical sections of cable are then plumbed or clamped to the appropriate tacking bars.

The cable joint used would be of the tunnel type as shown in sketch (*c*). The overall diameter of this joint is considerably reduced by extending the wire joints along the sleeve for a greater distance than normal for a manhole joint. Sketch (*d*) shows the type of bracket and bearer that would be used in a concrete-lined tunnel.




(*c*)



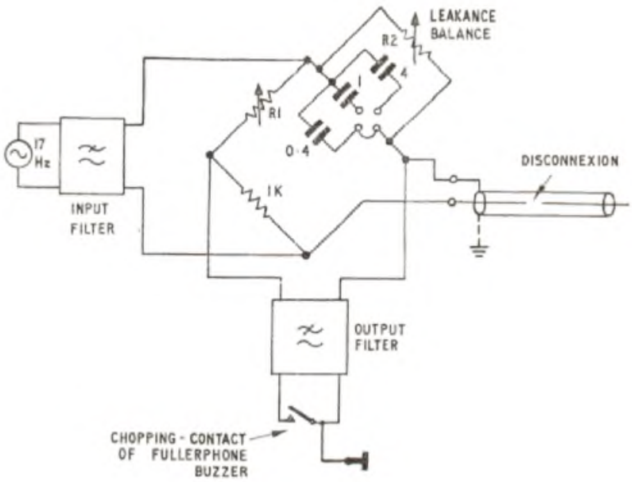

(*d*)

*Q. 2. Describe how a disconnexion in the centre conductor of a 0·375-in coaxial pair would be located using an a.c. bridge. Draw a diagram of the circuit and describe how the output from the bridge can be audibly detected if the frequency used is below the audio range.*

A. 2. A disconnexion in the centre conductor of a coaxial cable may be located by the use of a low-frequency a.c. bridge. Low frequencies enable the capacitance to be measured accurately over a greater distance, by reducing the relative effect of the cable's inductance. A frequency of 17 Hz can be used satisfactorily on long lengths of loaded or unloaded cable.

In order to detect the low-frequency output from the bridge, it is made audible by chopping at an audio frequency. This can be done using a Fullerphone buzzer vibrating at about 400 Hz, with a chopping contact in series with the receiver. A suitable test circuit is shown in the sketch.

The low-frequency supply, which may be from a hand generator, is connected to the bridge via a low-pass input filter which cuts off at about 30 Hz. The input filter eliminates harmonics which could make balancing difficult. A U-link enables one of the three standard capacitors (0·4 $\mu$F, 1 $\mu$F or 4 $\mu$F) to be selected. The line capacitance, between the inner and outer conductor up to the point of the disconnexion, is balanced against the standard capacitance by adjusting the variable resistor, R1, of value up to 1 kohm. The indicator dial of this variable resistor gives the value as a percentage of the standard capacitance selected.




The resistor, R2, of value up to 100 ohms, is adjusted to balance the leakance of the line under test.

The unbalance current from the bridge is taken through another 30 Hz low-pass filter to cut out 50 Hz induction from power lines.

If $R_1$ = value of resistor R1, $C$ = standard-capacitor value, $C_x$ = capacitance of line, then, when bridge is balanced,

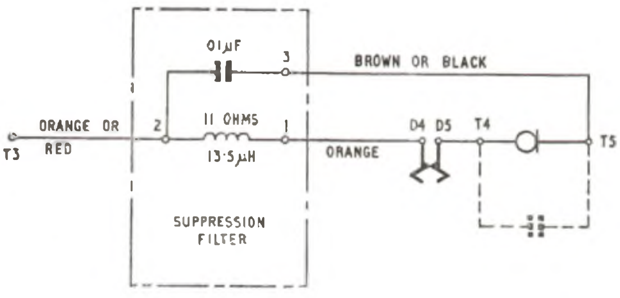
$$\frac{C_x}{C} = \frac{R_1}{1{,}000}.$$

The capacitance of a coaxial pair with a good inner conductor is determined in a similar manner. The distance to the disconnexion can then be calculated, since the capacitance between inner and outer conductors is directly proportional to the length.

*Q. 3. Explain how a radio transmitter can cause interference with a subscriber's telephone circuit. Describe the measures which are taken to eliminate interference of this type.*

A. 3. A subscriber's telephone may be subjected to interference from a high-powered radio transmitter when open-wire spans serving the telephone are in the vicinity of the transmitter. The open-wire spans act as a receiving aerial, a longitudinal e.m.f. being induced in the same direction in each wire of the span. The subscriber's circuit is normally sufficiently unbalanced, with respect to these induced e.m.f.s, for transverse currents to be created in the telephone circuit. These transverse currents give rise to radio-frequency (r.f.) potentials which, if non-linear components exist in the subscriber's circuit, may become rectified, producing audio-frequency currents and causing interference. Typical non-linear devices are the carbon-granule transmitter in the telephone and any high-resistance joints in the circuit.

Interference may be reduced by:




(*a*) *Reducing the r.f. currents flowing in the circuit.*

Air-cored r.f. inductors of 700–1,000 $\mu$H are used to limit the longitudinal r.f. current flowing in the telephone line. The overhead circuit is broken at a suitable point and an inductor is connected in series with each leg of the line. Each coil is provided with two tails of 18 s.w.g. wire to facilitate connexion, and is designed to be accommodated in the top cavity of an insulator.

(*b*) *Ensuring that no faulty joints exist in the circuit.*

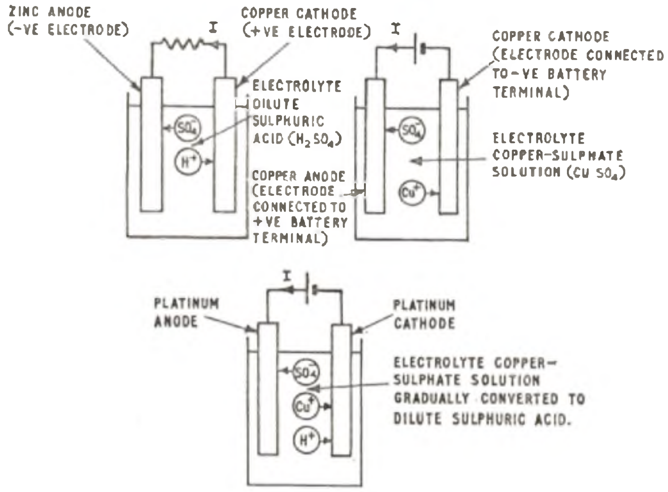
Faulty joints in the subscriber's circuit may be avoided by ensuring that joints are correctly made during installation, and that subsequent maintenance is properly carried out.

(*c*) *Fitting a suppression device in the subscriber's telephone.*

The suppression device installed in the telephone instrument consists of a 0·1 $\mu$F capacitor and a 13·5 $\mu$H r.f. inductor, and is designed to operate as a transmitter filter. It is connected as shown in the sketch. The unit eliminates most radio interference in the telephone instrument, but in exceptional cases it may be necessary to fit capacitors of value up to 0·15 $\mu$F across the transmitter (shown dotted).

*Q. 4. Show by reference to a simple electrolytic cell what is meant by polarization. Describe in detail two causes of polarization.*

A. 4. When current passes through any of the cells illustrated in the sketch, the electrode potentials change from their equilibrium




values. At the anode, the electrode potential tends to become less negative; whilst at the cathode, the electrode potential tends to become less positive. The change of potential increases as the current density at the electrode increases, but not linearly. This potential-change effect is known as polarization. Polarization is caused in the following ways.

(*a*) *Activation Polarization.* Whilst an equilibrium electrode potential is being maintained, dynamic equilibrium exists between the metal ions passing into solution (solvation) and the metal ions being deposited on the electrode (deposition). For a current to flow, a net reaction must take place, e.g. for an anodic current, the rate of solvation must exceed the rate of deposition.

If an appreciable current is to flow, the potential of the electrode must change from its equilibrium value, assisting one process and opposing the other, thus sustaining the flow of current. The difference from the equilibrium potential of the reaction is the "overpotential," and this form of polarization is called activation polarization or activation overpotential. The value of the overpotential associated with any reaction depends upon the current density, the type of reaction and, in some cases, the nature and condition of the electrode surfaces. The chief reactions at which considerable activation polarization occurs, are the anodic evolution of oxygen and the cathodic evolution of hydrogen.

(*b*) *Concentration Polarization.* The current flow through an electrolyte is maintained by the movement of positively-charged ions towards the cathode and negatively-charged ions towards the anode. If the current flow is small, the migration of ions towards their respective electrodes is sufficient to maintain the ionic concentrations around the electrode. If the current is increased, a value will be reached where the rate at which ions are involved in the reactions at anode and cathode exceeds the diffusion rate of these ions through the electrolyte. Under these circumstances, the ion concentrations of the electrolyte surrounding the anode are increased, whilst the ion concentration of the electrolyte surrounding the cathode are reduced. The anode potential thus becomes less negative and the cathode potential less positive. This type of polarization is "concentration polarization." It occurs at a critical current density, the value depending on the overall concentration of ions in the electrolyte, the speed of the ions, and the temperature of the electrolyte. The limiting current density is lower in a stagnant solution, since any movement tends to produce a more uniform concentration.

(*c*) *Ohmic Polarization.* The electrical resistance of the electrolyte at the surface of the electrode will cause changes in potential proportional to the current flowing in the cell. This potenial drop within the cell gives rise to "ohmic polarization." Factors which contribute to the electrical resistance of a cell are the concentration, the effective surface area of the electrodes (affected by irregularities in the composition of the metal surfaces), and the presence of obstructive films. The latter may be, for example, layers of oxide, insoluble salts, grease or hydrogen gas. The most important obstructive films are caused by the formation of oxides at the anode, and of hydrogen gas at the cathode.

**Q. 5.** *Give an account of the stresses which will occur in the lead sleeve of a cable joint which contains air under pressure.*

*A lead sleeve of a cable joint contains air at 9 lb/in² pressure. If the sleeve has an internal diameter of 6 in, how thick would the wall of the sleeve have to be if the stress in the lead must not exceed 150 lb/in²?*

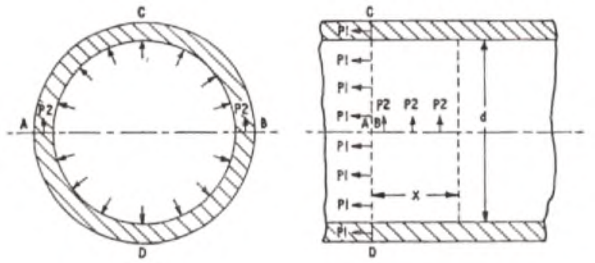
**A. 5.** The lead sleeve of a cable joint containing air under pressure is subject to two stresses, longitudinal stress and hoop stress.

The maximum safe pressure which can be resisted by a sleeve may be calculated by likening the sleeve to a thin cylindrical shell, and considering the longitudinal and hoop stresses separately.

*Longitudinal stress*

Let $P$ = internal gas pressure in lb/in²,
$t$ = thickness of the shell in inches,
$d$ = internal diameter in inches,
$p_1$ = intensity of longitudinal stress.

As shown in the sketch, the internal gas pressure will tend to tear




the shell across planes such as ABCD. Whether the ends are plane or dished, the force tending to produce such a failure will be $\frac{P\pi d^2}{4}$ pounds.

The total force resisting this pressure will be the product of the intensity of longitudinal stress and the surface area, $\pi dt$, that this force is acting on.

$$\therefore \text{ total longitudinal stress} = \pi dtp_1.$$

Equating these two forces,

$$\pi dtp_1 = \frac{P\pi d^2}{4}.$$

$$\therefore\ p_1 = \frac{Pd}{4t} \text{ lb/in}^2.$$

*Hoop stress*

Consider a length $x$ of the shell as shown in the sketch. The internal gas pressure will apply a force throughout this length tending to separate the shell into two half cylinders along the plane AB. This force of separation is $dxP$ pounds.

The stress on the sheath tending to resist this failure is the hoop stress. Let the hoop stress be $p_2$ lb/in², then the total resisting force will be the area of section of shell multiplied by $p_2 = 2xtp_2$.

Equating these two forces,

$$2xtp_2 = dxP.$$

$$\therefore\ p_2 = \frac{Pd}{2t} \text{ lb/in}^2.$$

Thus hoop stress is twice as large as the longitudinal stress, and therefore it is the hoop stress which will limit the working gas pressure.

In the example given:

$P$ = 9 lb/in²,
$p_2$ = hoop stress = 150 lb/in²,
$d$ = 6 in,
$t$ = wall thickness of sleeve in inches.

Now hoop stress $p_2 = \frac{Pd}{2t}$.

$$\therefore\ 150 = \frac{9 \times 6}{2t},$$

$$\text{and } t = \frac{9 \times 6}{300} \text{ in.}$$

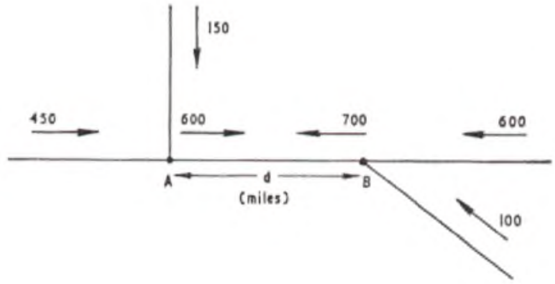
$$= \underline{0 \cdot 18 \text{ in.}}$$

**Q. 6.** *Describe the steps you would take to determine the Practical Centre of a telephone-exchange area. If a suitable site for the telephone exchange is not available at the Practical Centre, describe the next steps you would take in finding an actual site.*

**A. 6.** The first step in determining the Practical Centre for a telephone-exchange area is to obtain a forecast of the demand for telephones, and other services requiring a cable pair in the area, during the next 20 years. The forecast figures are most conveniently shown on a map of the area divided into blocks of roughly uniform density.

The forecast figures must now be converted to densities of telephones per square mile. If the map is shaded according to density, it will be possible to pick out the densest area. The Practical Centre will tend to be in the densest part of the area and if there is only one point where the principal routes intersect, that is likely to be the best Practical Centre. The next stage is to prepare a map of the area showing the distribution of subscribers and the junction forecasts, the latter being marked at the points where the existing or proposed routes enter the exchange area. By trial and error, a north-south line and an east-west line should be drawn such that each equally divides the forecast. The point at which these lines cross is the ideal location for the Practical Centre, and again if there is only one point in the vicinity where the principal routes intersect that will be the best Practical Centre.

The actual Practical Centre may now be determined using a pair-mileage assessment. A simple example to illustrate the method is shown in the sketch. The possible positions for the pair-mileage centre




are at points A and B. The figures and the arrows indicate the number of pairs required on each route at the 20-year date. The different figures for the centres at A and B have been shown in the common section AB. The distance from A to B is assumed to be $d$ miles.

Thus with the exchange at A, 700 $d$ pair-miles are required between A and B. If the exchange is at B then 600 $d$ pair-miles are required, hence B is the Practical Centre. This is readily seen to be true in this example because with the exchange at B the same number of pairs are required between A and B as in the larger spur at B. Hence the point of minimum pair-mileage must be where the length of the 100-pair spur is a minimum, i.e. at B. In calculating the pair-mileage, corrections have to be made for differences in conductor size. This is usually achieved by the use of weighting factors to bring all cable pairs to a common size, say 4 lb/mile.

If this method is inadequate, a full cost comparison of the plant required on a percentage value of annual charges basis, has to be made with the exchange at various sites.

If a suitable site is not available at B it would be necessary to search for a site as near to B as possible, preferably close to a major cable route. The extra cable cost due to departure from such a route may be considerable, but a site on "back land" may be much cheaper to buy, and may also permit the erection of a less-expensive type of building. It is convenient to prepare a "lines of search" map along the main routes to indicate the roads on which a site would be preferred, and the limits to which the initial search should extend.

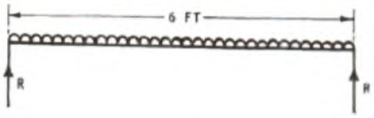
Unless any suitable site which is found is clearly the cheapest, it would be necessary to calculate the out-of-centre cabling cost on a percentage value of annual charges basis for the alternatives, so that the most economic one could be chosen.

**Q. 7.** *Explain what is meant by pre-stressed concrete. What are the relative advantages and disadvantages of pre-stressed concrete and ordinary reinforced concrete? How could the pre-stressing technique be applied to the construction of a manhole?*

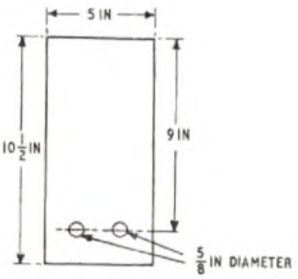
**A. 7.** See A. 10. Line Plant Practice C, 1961, *Supplement*, Vol. 55, No. 2, July 1962.

**Q. 8.** *A reinforced-concrete beam is freely supported at its ends over a clear span of 6 ft. It is of rectangular section 10½ in deep and 5 in wide with two steel reinforcing rods each ⅜ in diameter embedded 9 in from the upper face. Assuming that the modular ratio is 15 and the maximum permissible stress is limited to 20,000 lb/in² in the steel and 750 lb/in² in the concrete, calculate the safe working load that can be uniformly distributed over the beam.*

**A. 8.** Sketch (*a*) shows the beam loading and sketch (*b*) shows the cross section of the beam used.




(*a*)




(*b*)

Cross-sectional area of reinforcing steel, $A_s = 2 \times \left(\frac{5}{16}\right)^2$ in$^2$,

$= 0{\cdot}614$ in$^2$.

Ratio of area of steel to area of concrete, $p = \frac{0{\cdot}614}{9 \times 5}$.

Modulator ratio, $m = 15$,

$$\therefore\ pm = \frac{0{\cdot}614 \times 15}{9 \times 5} = 0{\cdot}204,$$

$$\therefore\ p^2m^2 = (0{\cdot}204)^2 = 0{\cdot}0416.$$

Substituting in the expression,

$$\frac{h}{d} = \sqrt{p^2m^2 + 2pm} - pm,$$

where $h$ = depth of neutral axis

and $d$ = effective depth of beam = 9 in,

$$\frac{h}{d} = \sqrt{0{\cdot}0416 + 0{\cdot}408} - 0{\cdot}204,$$

$$= 0{\cdot}467,$$

$$\therefore\ h = 0{\cdot}467 \times 9 \text{ in},$$

$$= 4{\cdot}203 \text{ in}.$$

If $M$ is the maximum bending moment when compression in concrete reaches 750 lb/in$^2$,

$$M = \tfrac{1}{2} f bh \left(d - \frac{h}{3}\right),$$

$$= \tfrac{1}{2} \times 750 \times 5 \times 4{\cdot}203\,(9 - 1{\cdot}4) \text{ lb in},$$

$$= \frac{750 \times 5 \times 4{\cdot}203 \times 7{\cdot}6}{2} \text{ lb in},$$

$$= 59{,}893 \text{ lb in}.$$

It is now necessary to ascertain that the stress in the steel does not exceed the permissible limit, when the concrete reaches its maximum compressive stress.

Maximum stress in steel, $f_s = \dfrac{M}{A_s\left(d - \frac{h}{3}\right)}$ lb/in$^2$,

$$= \frac{59{,}893}{0{\cdot}614 \times 7{\cdot}6} \text{ lb/in}^2,$$

$$= 12{,}832 \text{ lb/in}^2.$$

This is within the limit of 20,000 lb in$^2$ allowed.

For a distributed load on a beam $M = \frac{WL^2}{8}$,

where $L$ = span and $W$ = distributed load.

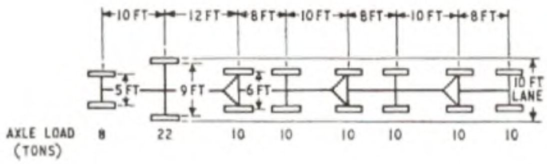
$$\therefore\ W = \frac{8M}{L^2} = \frac{8 \times 59{,}893}{72^2},$$

$$= 92{\cdot}4 \text{ lb/in}.$$

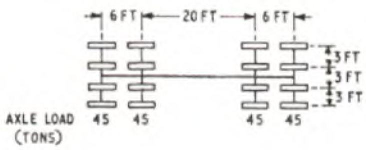
$\therefore$ Maximum distributed load on beam = 92·4 lb/in.

***Q. 9.*** *Give an account of the Standard Traffic Loadings which are used to determine the maximum traffic loads to be considered in the design of jointing chambers.*

**A. 9.** The maximum traffic loads which have to be taken into account in the design of jointing chambers are determined using vehicle axle loads specified by the Ministry of Transport. This information is normally used for bridge-design work, but applies equally to other load-bearing structures. Two such trains are in use, sketch (*a*) shows the one for normal highway loading and sketch (*b*) the one for abnormal loading. These are applied as follows:




(*a*)




(*b*)

(*a*) *Motorways and trunk roads.* The first two traffic lanes are designed for normal loading, and any further lanes for ⅓ of normal loading. The designed strength is checked for one lane carrying abnormal loading, and all other lanes ⅓ of normal loading.

(*b*) *Class I and class II roads.* The first two traffic lanes are designed for normal loading, and any further lanes for ⅓ of normal loading. The designed strength is checked for one lane carrying ⅚ of the abnormal loading and all other lanes ⅓ of normal loading.

(*c*) *Class III roads.* As for class I and class II roads, but no allowance is made for abnormal loading.

Consideration of the Ministry of Transport trains enables structural members of all sizes to be designed. For large load-carrying members the loadings approximate to uniform distribution and have been converted into standard loading curves, but for slabs which are supported on all four sides in such a way that the distance between supports in one direction is less than twice the corresponding distance in the other, the loading curves cannot be used, and individual wheel loads must be considered. Where individual wheel loads are considered, the following conditions are specified:

(*a*) two wheel loads each to be 11¼ tons in line, transversely spaced 3 ft apart in the direction of travel, and each having a contact area of 15 ft × 3 in,

(*b*) dispersal under the wheel loads to be taken at 45°,

(*c*) 25 per cent overstress to be allowable, when considering the effects of the 11¼-ton wheel loads,

(*d*) no allowance to be made for impact.

*Footway Loads*

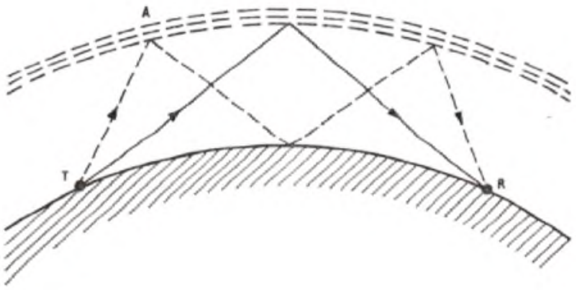
The normal footway load used in bridge design is 80 lb/ft$^2$, but this is increased to 100 lb/ft$^2$ where crowds may be encountered. For road verges where traffic may occasionally be driven off the main road, a single wheel load of 4 tons (including impact) must be allowed for. In this case a stress distribution at an angle of 45° and an overstress of 25 per cent may be allowed as above. The area of contact is taken as a circle of 12 in diameter.

**Students were expected to answer any six questions**

*Q.* 1. *Explain how selective fading can arise on a long-distance short-wave radio link. How can its occurrence be minimized by a suitable choice of frequency?*

*Briefly outline some radio transmission and reception techniques which reduce the effects of selective fading in (a) telephony, (b) telegraphy.*

A. 1. Selective fading commonly arises on short-wave radio links because of multipath propagation. The sketch gives an example in




which transmission via the ionosphere occurs both by the main one-hop path and also by a subsidiary two-hop path. The two-hop path is the longer, so that the signal received by this route is delayed relative to the main signal. Signals received over the two paths add vectorially at the receiver, so that the resultant signal strength depends on their amplitudes, and their relative phase, which is a function of the time-delay difference and the frequency within the bandwidth of the transmission. For example, if the path time-delay difference, $\tau$, is 1 ms, the signals will be in phase at frequencies separated by

$$\frac{1}{\tau} = \frac{1}{10^{-3}} = 1{,}000 \text{ Hz},$$

throughout the bandwidth of the signal. Midway between these frequencies the signals are in phase opposition, tending to cancel each other out, so that a frequency-selective fading pattern exists. Because of the unstable nature of the ionosphere, the path-delay difference is usually varying, with the result that the carrier and all the frequency components in the sidebands of the signal fade in the manner described, but they go through their minimum values at different times.

Selective fading can generally be minimized by choosing a frequency close to the maximum usable frequency, since multipath propagation is then less likely to occur. In the example shown in the sketch the frequency could be increased until the unwanted two-hop path completely penetrated the layer at A and was lost; this would leave only the one-hop path which, because of its greater angle of incidence, would continue to be reflected by the layer.

In telegraphy the signal occupies only a narrow bandwidth and diversity reception—either space or frequency—is effective in minimizing selective fading. In space diversity the signal is received simultaneously on two aerials several wavelengths apart, each having its own receiver. The aerials provide signals which fade independently of each other, and the output circuit is switched automatically from one receiver to the other according to which is providing the stronger signal at any given instant. In frequency diversity, the signal is transmitted simultaneously on two frequencies which are sufficiently separated to fade independently. Only one aerial is needed for reception and is connected to two receivers, one tuned to each frequency. The output is switched between these receivers as in space-diversity reception.

Frequency-shift, or two-frequency keying, has an inherent advantage over on–off keying, since with appropriate separation between the mark and space frequencies the two frequencies tend to fade independently; this gives a measure of in-built frequency diversity which can be utilized by suitable detection circuits.

In telephony, because of the larger bandwidth of the signal, diversity reception is of little help. The use of single side-band rather than double side-band transmission considerably reduces the distortion caused by selective fading of the carrier frequency. Compressor-expander techniques such as Lincompex, in which information defining the amplitude of the speech signal is transmitted on a separate narrow-band f.m. channel within the speech band, have proved highly successful in combatting the effects of selective fading in speech. The use of highly-directive arrays of aerials, steerable in the vertical plane, enables the angle corresponding to the dominant mode of propagation to be selected and other modes suppressed, thus minimizing multipath and selective fading.

*Q.* 2. *An earth station is receiving transmissions from a space-research satellite on a frequency of* 136 *MHz* (*Mc/s*). *The satellite is at a range of* 500 *km and its transmitter supplies* 0·5 *W into an aerial having a gain of* 3 *dB with reference to an isotropic aerial. Assuming free-space propagation, and taking the impedance of free-space as* $120\pi$ *ohms, calculate:*

*(a) the power flux density in watts/m², *

*(b) the field strength in* $\mu V/m$,

*at the earth station.*

*If the aerial at the earth station has a gain of* 20 *dB with reference to an isotropic aerial, what is the signal power received? (The effective absorbing area of an isotropic aerial is* $\lambda^2/4\pi$.)

A. 2. (*a*) The satellite has an aerial gain of 3 dB, hence the effective radiated power (e.r.p.) in the direction of the earth station is twice the power supplied to the aerial, i.e. $0{\cdot}5 \times 2 = 1$ watt e.r.p. As seen from the earth station, therefore, the satellite is equivalent to an isotropic source radiating 1 watt. An isotropic source would radiate equally in all directions, so that the power flux density at a distance $r$ would be equal to the power of the source, $P$, divided by the surface area of a sphere of radius $r$. Thus,

$$\text{power flux density} = \frac{P}{4\pi r^2}.$$

Substituting $P = 1$ watt, and $r = 500 \times 10^3$ m,

$$\text{gives, power flux density at earth station} = \phi = \frac{1}{4\pi(5 \times 10^5)^2},$$

$$= \underline{0{\cdot}317 \times 10^{-12} \text{ watts/m}^2}.$$

(*b*) If $E$ is the field-strength in volts/m and the impedance of free space is $Z = 120\pi$ ohms, then

$$\frac{E^2}{Z} = \text{power flux density in watts/m}^2.$$

$\therefore$ At the earth station,

$$E = \sqrt{120\pi \times 0{\cdot}317 \times 10^{-12}} = 11 \times 10^{-6} \text{ volts/m} = \underline{11\,\mu\text{V/m}}.$$

Received power = (power flux density) × (effective absorbing area of aerial).

$$\text{Now wavelength, } \lambda = \frac{300}{\text{frequency (MHz)}} = \frac{300}{136} = 2{\cdot}22\,\text{m},$$

$$\text{and effective absorbing area of isotropic aerial} = \frac{\lambda^2}{4\pi} = 0{\cdot}39 \text{ m}^2.$$

$\therefore$ Power received by an isotropic aerial

$$= 0{\cdot}317 \times 10^{-12} \times 0{\cdot}39 = 12{\cdot}2 \times 10^{-14} \text{ watts}.$$

Since the earth-station aerial has a gain of 20 dB, it will receive 100 times as much power as an isotropic aerial.

$\therefore$ Power received by earth station $= P_R$,

$$= 12{\cdot}2 \times 10^{-14} \times 100 = \underline{12{\cdot}2 \times 10^{-12} \text{ watts}}.$$

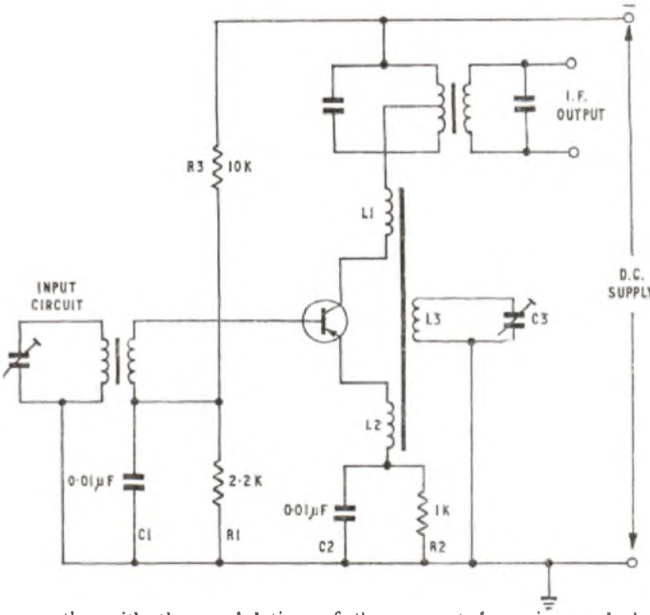
*Q.* 3. *Draw the circuit diagram of a self-oscillating transistor frequency-changer stage for a radio receiver, showing the biasing arrangements and giving typical component values. Explain the operation of the circuit.*

*Explain why, and in what circumstances, cross modulation would be likely to occur in such a stage.*

A. 3. In the circuit shown in the sketch, a single transistor functions both as oscillator and frequency-changer. The frequency of oscillation is determined by inductor L3 in parallel with capacitor C3. Inductors L1 and L2 in the collector and emitter circuits provide positive feedback through the transistor to overcome losses, and hence to maintain the oscillation. The amplitude is stabilized by the biasing resistor R2 in parallel with capacitor C2 in the emitter circuit. Bias stabilization for the base circuit is through resistors R3 and R1, which form a potential divider across the d.c. supply. Resistor R1 is by-passed by capacitor C1 and therefore has no effect on the radio frequency (r.f.) currents flowing in the base circuit.

The input circuit is tuned to the signal frequency, the variable capacitor being ganged to that in the oscillator circuit. Padding and trimming capacitors are included (not shown in the sketch) to give accurate tracking over the required frequency range. A separate winding, coupled to the input tuned circuit, transfers the signal to the base of the transistor. Between base and emitter, the signal is effectively in series with the oscillator voltage developed across inductor L2. Under working conditions, the collector current is a non-linear function of the effective voltage between base and emitter, and thus contains components at the sum and difference frequencies of the oscillator and signal frequencies. The intermediate frequency (i.f.) output transformer is tuned to the difference frequency. The tapped connexion to the primary is arranged to provide a correct match to the transistor output impedance at the i.f., while presenting a very low impedance at the oscillator frequency.

Cross modulation occurs if a strong signal is present on a frequency close to that of the wanted signal and the selectivity of the input tuned circuit cannot provide sufficient rejection. In these circumstances the conversion conductance of the frequency-changer tends to vary in



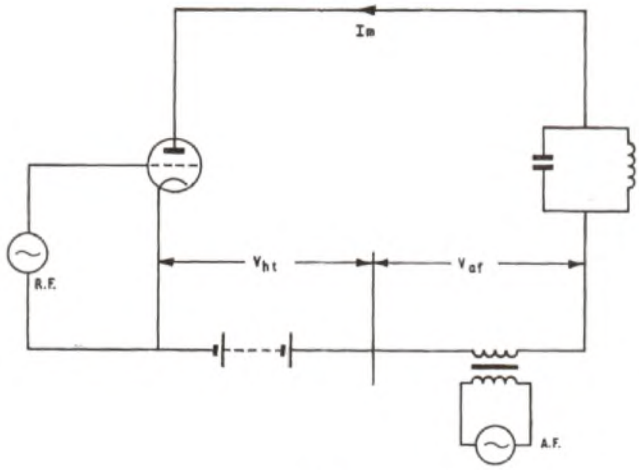

sympathy with the modulation of the unwanted carrier, and the interfering signal can thus become superimposed on the wanted signal.

*Q.* 4. *An amplitude-modulated transmitter has an anode-modulated Class-C output stage in which an audio-frequency sine wave of* 3 *kvolts peak value is developed across the secondary of the modulating transformer in series with the* 5 *kvolts h.t. supply. The stage has an anode efficiency of* 75 *per cent and delivers* 1·5 *kwatts of carrier power into the tank circuit. Calculate:*

(*a*) *the depth of modulation,*
(*b*) *the mean anode current,*
(*c*) *the power supplied by the modulator,*
(*d*) *the total r.f. power delivered to the tank circuit.*

*State the assumptions you have made in these calculations.*

A. 4. The basic circuit is shown in sketch (*a*). The anode supply




(*a*)

for the r.f. amplifier stage is effectively $V_{ht} + V_{af}$, the latter being the audio-frequency voltage developed across the secondary of the modulating transformer. The r.f. voltage at the anode is approximately proportional to the supply voltage and thus varies in accordance with the waveform shown in sketch (*b*).

(*a*) Depth of modulation

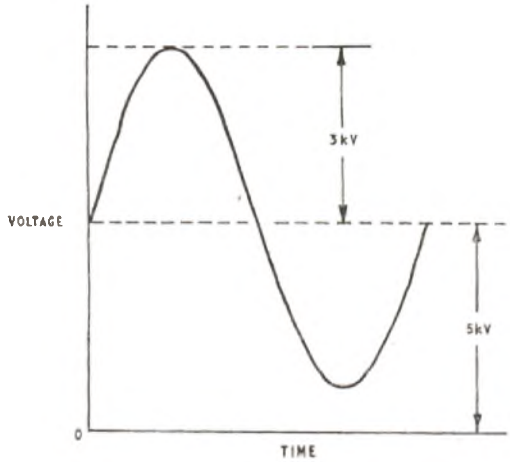
$$= \frac{\text{peak a.f. voltage}}{V_{ht}} = \frac{3{,}000}{5{,}000} \times 100 = \underline{60 \text{ per cent.}}$$

(*b*) If the modulation is removed, the power output is that of the carrier only.

The power drawn from the supply $= V_{ht} \times$ mean anode current ($I_m$).

Thus, $I_m \times V_{ht} \times 0{\cdot}75 = 1{\cdot}5$ kW,

$$\text{and, } I_m = \frac{1{,}500}{5{,}000 \times 0{\cdot}75} = \underline{0{\cdot}4 \text{ amps}}$$




(*b*)

(*c*) The sideband power is drawn from the modulator, which therefore supplies

$$1{,}500 \times \frac{(0{\cdot}6)^2}{2} \times \frac{1}{0{\cdot}75} = \underline{360 \text{ watts.}}$$

Alternatively, since the load across the modulating transformer is given by $V_{ht}/I_m = 12{,}500$ ohms, the power supplied is

$$\frac{(3{,}000)^2}{2 \times 12{,}500} = \underline{360 \text{ watts.}}$$

(*d*) The total r.f. power delivered to the tank circuit is

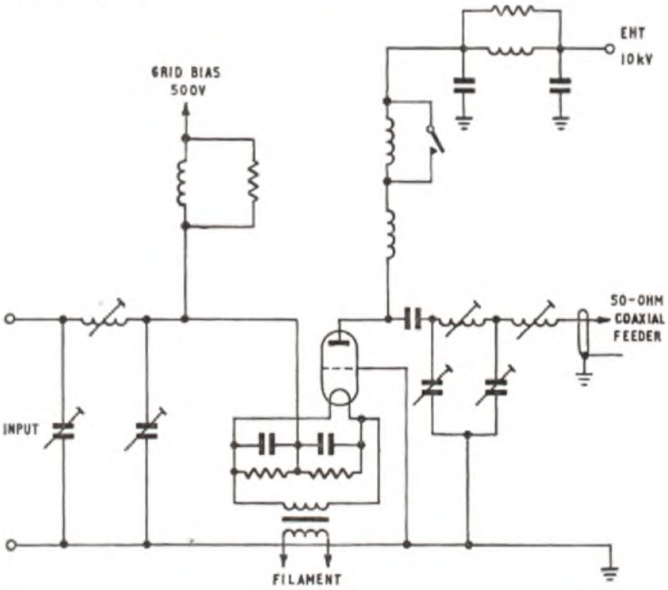
$$1{,}500 + 360 \times 0{\cdot}75 = 1{,}500 + 270 = \underline{1{,}770 \text{ watts.}}$$

The main assumptions are that the stage works at constant efficiency, and that the amplitude of the r.f. output is directly proportional to the anode supply voltage.

*Q.* 5. *Describe, with the aid of a circuit diagram, the output stage of a high-power short-wave transmitter suitable for long-distance point-to-point services. Show the arrangements for matching the output to a 50-ohm coaxial feeder.*

*In what ways are the performance requirements of the output stage different for* (*a*) *200-baud single-channel frequency-shift telegraphy,* (*b*) *4-channel independent-sideband telephony?*

A. 5. The sketch shows the basic circuit of an output stage suitable




for telegraphy or telephony transmissions of 20 kW to 30 kW power, at any frequency from 4 MHz to 27 MHz. A single grounded-grid vapour-cooled triode is used with the anode supply in shunt feed, and $\pi$-configuration tuned circuits at the input and output. The grounded-grid circuit, operating under Class-B conditions, provides the linearity necessary for telephony transmission, and in addition gives high stability without the need for neutralization.

The $\pi$-tuned circuits each consist of a series inductor with a shunt capacitor at both ends. All three elements are variable, so that a wide frequency range can be covered while the on-load Q-factor and power-conversion efficiency remain constant. Impedance transformation of the coupling circuits is determined by the ratio of the shunt capacitances. The input tuned-circuit provides matching between the penultimate and final stages. The output tuned-circuit similarly matches the anode impedance to the 50-ohm coaxial feeder, with the aid of a further variable series inductor for loading adjustment.

The 10 kvolts e.h.t. supply is fed to the anode through a decoupling filter and a series choke. The choke is provided with tapping points so that, according to the frequency in use, sections of it may be switched out to eliminate resonance due to stray capacitance. The resistor shunting the inductor of the decoupling filter serves a similar purpose by damping unwanted resonances. The anode is connected to the output tuned-circuit through a capacitor which blocks the valve d.c. For simplicity of operation the various tuning elements are ganged and in modern transmitters, servo-controlled motor drives are employed so that tuning and matching are carried out automatically when a frequency change is initiated.

The differences in performance requirements are summarized below:

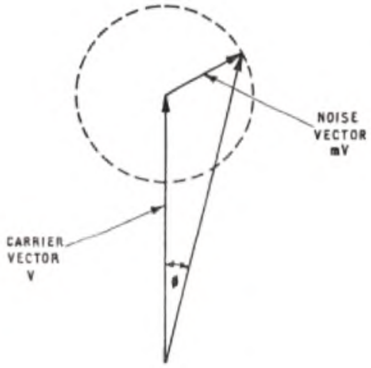
(*a*) The frequency-shift telegraphy signal would require a bandwidth of only a few hundred Hertz. As the signal is of constant amplitude, linearity of the output stage is relatively unimportant. Class-C amplification could therefore be used, with consequent high efficiency. Because the signal amplitude and power consumption are constant the h.t. supply would not require a high degree of regulation.

(*b*) The 4-channel independent side-band telephony transmission would require a bandwidth of 12 kHz. Linearity is essential to prevent distortion and crosstalk between the channels. Class-B amplification could be used to obtain linearity with good efficiency. The signal amplitude and the power consumption are constantly varying and good h.t. regulation would therefore be necessary.

*Q. 6. Explain why f.m. transmission can give an improved signal-to-noise ratio compared with a.m. transmission of the same carrier power. What characteristics of f.m. transmission determine the magnitude of this improvement?*

*An f.m. radio link having a deviation ratio of* 10 *is to transmit speech occupying the audio band up to* 3 kHz (*kc/s*). *What r.f. bandwidth would normally be used for this transmission? What would be the effect on* (a) *the r.f. bandwidth, and* (b) *the signal-to-noise ratio, if the deviation ratio were reduced to* 5?

A. 6. Consider two radio systems, one a.m. and the other f.m., in which carriers of equal power are modulated by an audio-frequency signal. The peak a.f. signal produces 100 per cent modulation of the a.m. carrier and a frequency deviation, $\Delta f$, of the f.m. carrier. Both receivers have the same level of random noise, which is uniform over the frequency bandwidth, so that the noise components at all frequencies are of equal amplitude. In sketch (*a*) the carrier voltage is

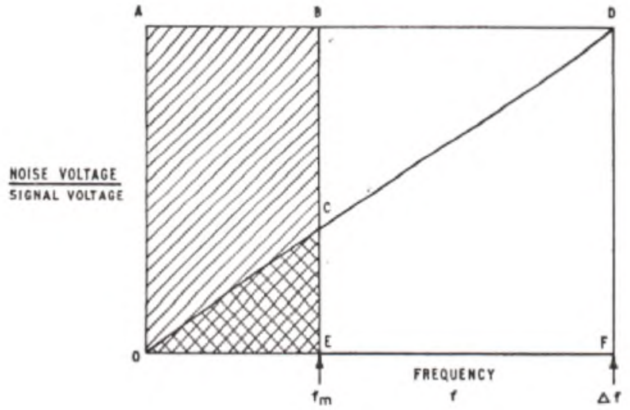



(*a*)

represented by the fixed vector $V$, and the voltage of the noise component at any frequency is the vector $mV$, which rotates at a speed determined by its frequency difference from the carrier. The latter produces both amplitude and phase modulation, the depth of modulation being independent of the frequency of the noise component. Since the amplitude of the noise is normally very much smaller than that of the carrier, both types of modulation are approximately sinusoidal, the depth of amplitude modulation is $m$, and the peak phase deviation is $m$ (radians).

In the a.m. system, the detector ignores the phase variations but responds to the amplitude modulation. Thus, in the audio-frequency output, the amplitude of a noise component is proportional to $m$, and is constant with frequency. This is indicated in sketch (*b*) by the line AB. Noise components at frequencies higher than the maximum modulating signal frequency, $f_m$, can be ignored since they are suppressed by a low-pass filter following the detector.

In the f.m. system, the amplitude modulation is removed by a limiter, leaving only the phase modulation. The discriminator




(*b*)

responds to frequency variations, so that its output is proportional to the rate of change of phase. For noise separated from the carrier by a frequency difference, $f$,

$$\text{phase deviation} = \phi = m \sin 2\pi f t,$$

and,

$$\text{frequency deviation} = \frac{1}{2\pi} \times \frac{d\phi}{dt} = mf \cos 2\pi f t.$$

The frequency deviation due to the noise is therefore directly proportional to its frequency separation from the carrier. The noise voltage at the output of the discriminator thus increases linearly with frequency as indicated by the line OD in sketch (*b*). As noise components higher than $f_m$ can be ignored, the noise affecting the output is indicated by the triangle OCE, and is clearly much less than the corresponding noise in the a.m. system which is indicated by the rectangle OABE. This shows that f.m. is capable of giving an improved signal-to-noise ratio compared to a.m., though at the expense of occupying a wider transmission bandwidth.

The f.m. characteristics which determine the magnitude of the improvement can be obtained by considering the signal-to-noise power ratio, S/N, in each case.

For the a.m. system, $S/N_{am} \propto \dfrac{1}{(\mathrm{BE})^2}$.

For the f.m. system, $S/N_{fm} \propto \dfrac{1}{(\mathrm{CE})^2}$.

$\therefore$ The f.m. improvement ratio $\dfrac{S/N_{fm}}{S/N_{am}} \propto \left(\dfrac{\mathrm{BE}}{\mathrm{CE}}\right)^2$.

Now, since triangles ODF and OCE are similar,

$$\frac{\mathrm{BE}}{\mathrm{CE}} = \frac{\mathrm{DF}}{\mathrm{CE}} = \frac{\mathrm{OF}}{\mathrm{OE}} = \frac{\Delta f}{f_m}.$$

$$\therefore \frac{S/N_{fm}}{S/N_{am}} \propto \left(\frac{\Delta f}{f_m}\right)^2.$$

The f.m. improvement thus depends upon the ratio of the frequency deviation, $\Delta f$, to the maximum modulating frequency $f_m$; that is, it depends upon the deviation ratio. It can be shown that the f.m. improvement, expressed in decibels, is

$$20 \log_{10} \sqrt{3} \times \frac{\Delta f}{f_m}.$$

Let $\Delta f$ = maximum frequency deviation,
$f_m$ = maximum modulating frequency,
$D$ = deviation ratio,
$B$ = bandwidth.

Then, $D = \dfrac{\Delta f}{f_m}$,

and, $B \simeq 2(\Delta f + f_m)$.

When the deviation ratio is 10,

$$\Delta f = 3 \times 10 = 30 \text{ kHz},$$

$$\text{and} \quad B = 2(30 + 3) = \underline{66 \text{ kHz.}}$$

If the deviation ratio is reduced to 5,

(*a*) $\Delta f = 3 \times 5 = 15$ kHz,

$$\text{and,} \quad B = 2(15 + 3) = \underline{36 \text{ kHz.}}$$

Thus, the r.f. bandwidth would be reduced to a little more than half its original value.

(*b*) The signal-to-noise power ratio is proportional to $D^2$ and would therefore be reduced to

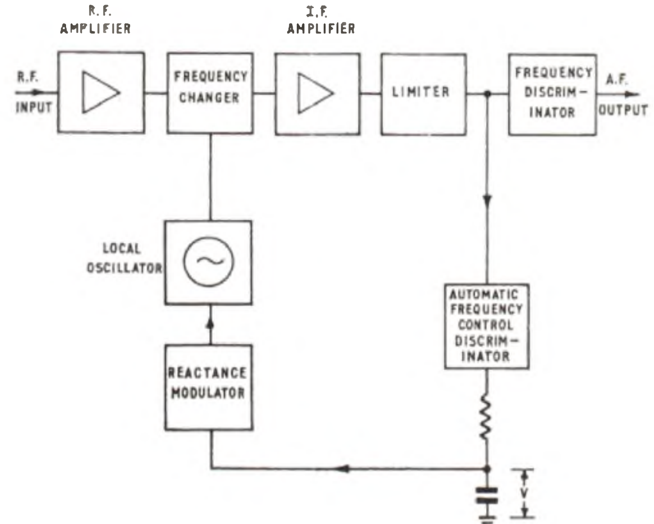
$$\left(\frac{5}{10}\right)^2 = \tfrac{1}{4} \text{ of the original value.}$$

The signal-to-noise ratio thus becomes $10 \log_{10} 4 = \underline{6 \text{ dB worse}}$.

*Q. 7. With the aid of a block-schematic diagram, describe the application of automatic frequency control to an f.m. radio receiver. Why is a.f.c. more necessary at v.h.f. than at lower frequencies?*

*An a.f.c. discriminator produces 1 volt of control bias for a frequency error of 50 kHz (kc/s), and the controlled oscillator is shifted by 250 kHz/V. Calculate the tuning error if the oscillator would have drifted 20 kHz from the correct frequency without a.f.c.*

**A. 7.** An arrangement for automatic frequency control (a.f.c.) of an f.m. radio receiver is shown in block-schematic form in the sketch.




The signal from the i.f. amplifier passes through a limiter to remove any amplitude modulation, and is then demodulated by a frequency discriminator from which the a.f. signal is taken. Part of the signal from the limiter is also fed into an a.f.c. discriminator, the output of which is filtered to remove all traces of the a.f. modulation. The d.c. voltage which remains is proportional to the difference between the mean frequency of the modulated i.f. signal and the nominal i.f. centre frequency to which the a.f.c. discriminator is tuned. The magnitude of this voltage thus depends on the frequency error of the i.f. signal, and its sign indicates whether the frequency is above or below the correct value.

The error voltage is then used to control the frequency of the local oscillator by varying the effective reactance of the tuned circuit with a device such as a reactance-modulator circuit. The frequency of the oscillator is thus controlled by the d.c. voltage in a manner which tends to correct the frequency error. With this system, a frequency error can never be fully corrected, since with zero error the control voltage itself would be zero. However, the action of the circuit considerably reduces the magnitude of any frequency errors, whether due to initial setting inaccuracies or to subsequent drift of the transmitter frequency or the receiver local-oscillator frequency.

The use of a.f.c. is more necessary at v.h.f. than at lower frequencies because a small percentage frequency drift results in a larger error in Hertz, with consequent greater distortion. For example, a drift of 0·01 per cent in a receiver working at 100 MHz would correspond to an error of 10 kHz, while at 1 MHz the error would be only 100 Hz.

In the problem, let the resulting frequency error with a.f.c. applied be $f$ (kHz). Since the frequency error without a.f.c. would be 20 kHz, the oscillator frequency is shifted by $20 - f$ (kHz) by the control voltage, $V$. Thus,

$$V = \frac{20 - f}{250}.$$

But $V$ is also the voltage produced by the discriminator by the frequency error $f$ (kHz).

$$\therefore\ V = \frac{f}{50} = \frac{20 - f}{250},$$

$$\text{from which, } 5f = 20 - f,$$

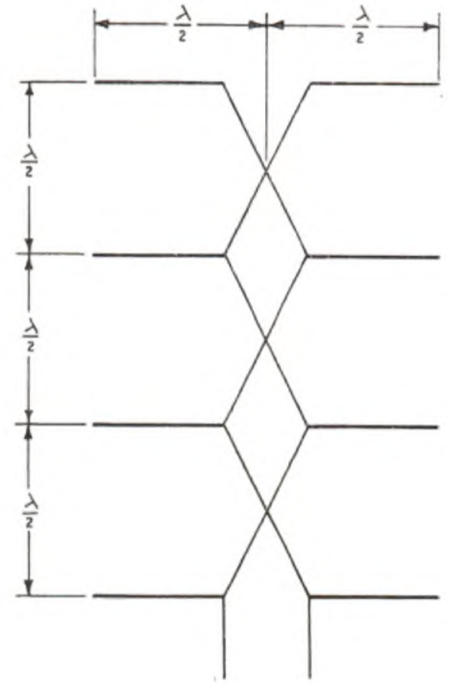
$$\text{and frequency error, } f = \frac{20}{6} = \underline{3 \cdot 33 \text{ kHz}}.$$

*Q. 8. Sketch the arrangement of a broadside aerial array comprising four vertical bays, each of four full-wave horizontal dipoles with reflectors.*

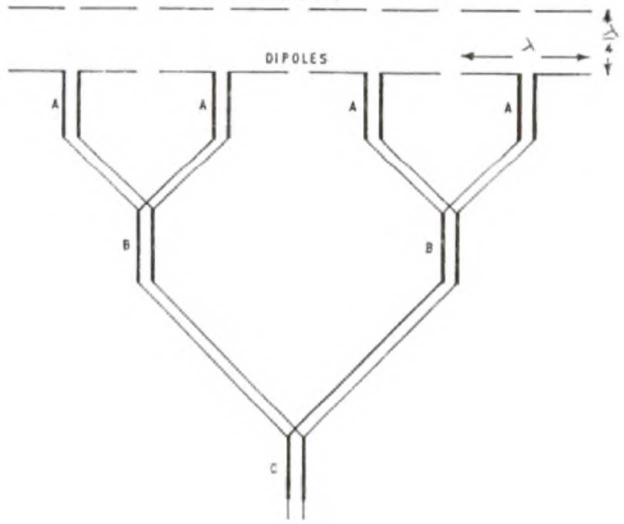
*Show (a) the dimensions and spacings in wavelengths, and (b) an arrangement for feeding the array from a 600-ohm balanced transmission line.*

*Estimate the gain in decibels of such an array relative to a single full-wave dipole. By how much would the gain be reduced if the reflectors were removed?*

**A. 8.** Sketch (*a*) shows the arrangement of one vertical bay of




(*a*)




(*b*)

aerials, and sketch (*b*) shows the plan view of four bays and the feeder layout. The length of each dipole element is slightly less than half the free-space wavelength. To obtain the required broadside radiation all dipoles in the array must be energized in phase. The feeder line in each vertical bay is therefore cross-connected between dipoles to compensate for the phase shift of 180° due to the half-wavelength of feeder. The lengths of the feeders from the common input to the foot of each bay must be equal.

If it is required to use 600-ohm feeder throughout, quarter-wave matching sections may be included as shown in sketch (*b*). A typical value for the impedance of each dipole would be 4,000 ohms, making the impedance of each bay 1,000 ohms. The impedance of the quarter-wave matching sections labelled A would therefore be

$$Z_A = \sqrt{1{,}000 \times 600} = 760 \text{ ohms.}$$

At points B and C the impedance to be matched to the line consists of two 600-ohm feeders in parallel, so that these matching sections must have an impedance of

$$Z_B = Z_C = \sqrt{300 \times 600} = 425 \text{ ohms.}$$

The array consists of 16 dipoles and will therefore have an effective power gain of approximately 16 times that of a single dipole. This is doubled by the presence of the reflectors, giving a total power ratio of 32 times, or $10 \log_{10} 32 = 15$ dB gain relative to a single full-wave dipole. If the reflectors were removed, the gain would be reduced by 3 dB.

**Q. 9.** *Explain what is meant by the sensitivity, and the selectivity, of a radio receiver.*

*Describe a method of measuring the gain/frequency characteristics of a receiver. Sketch, on the same axes, typical gain/frequency characteristics of receivers suitable for:*

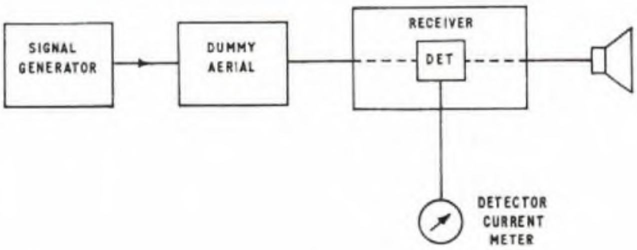
(*a*) *a.m. telephony in which the maximum modulating frequency is* 3 *kHz,*

(*b*) *f.m. telephony having a maximum frequency deviation of* 15 *kHz and a maximum modulating frequency of* 3 *kHz.*

**A. 9.** The *sensitivity* of a receiver is its ability to accept very weak input signals and give a satisfactory signal-to-noise ratio at the output. It is usually expressed as the minimum input signal, in microvolts, that will give the standard output signal-to-noise ratio under specified conditions of modulation.

The *selectivity* of a receiver is its ability to reject signals in frequency channels adjacent to the wanted transmission. It can be expressed in terms of the relative levels of the wanted and unwanted signals and their frequency separation.

The arrangement shown in sketch (*a*) may be used to measure the




(*a*)

gain/frequency selectivity characteristic of a receiver. The signal generator output is unmodulated and is coupled into the receiver through a dummy aerial having the same impedance as the type of aerial normally used. The signal level at the detector is monitored by a d.c. milliammeter connected in series with the diode load. The receiver automatic gain control must be made inoperative, since its action would prevent a true characteristic being obtained. The measurement procedure is as follows. The signal generator is set to the required frequency and the receiver tuned to this signal. The signal generator output level is then adjusted to give a convenient reference reading on the diode meter. The signal frequency is now varied to explore the characteristic both above and below the initial frequency. At each frequency setting the signal generator output level is adjusted to bring the diode current back to its original reference value. The increase of signal level in decibels is equal to the reduction in receiver gain relative to the initial midband frequency. The gain/frequency characteristic may then be plotted as in sketch (*b*).

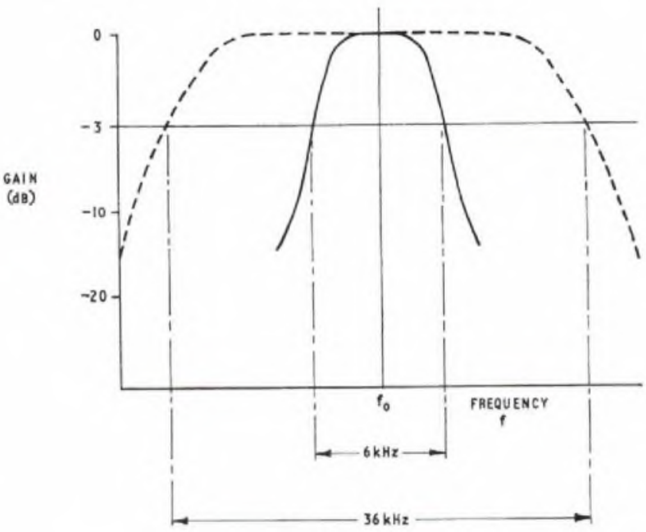
A different method, which is particularly useful when aligning a receiver, is to use a generator which is frequency modulated by a saw-tooth waveform and hence sweeps repeatedly across the i.f. bandwidth. The output from the detector is displayed on an oscilloscope which has its time base synchronized to the saw-tooth modulation cycle. An immediate display of the selectivity characteristic is thus obtained.

The characteristic must be sensibly flat over the bandwidth of the transmission, and should then cut off rapidly in order to suppress adjacent-channel signals. For the examples given the bandwidths are:

(*a*) for the case of a.m. telephony,

bandwidth = 2 × maximum modulating frequency,

= 2 × 3 kHz = 6 kHz.




(*b*)

(*b*) for the case of f.m. telephony,

bandwidth = 2 × (maximum frequency deviation + maximum modulating frequency),

= 2 × (15 + 3) = 36 kHz.

The required characteristics are shown in sketch (*b*).

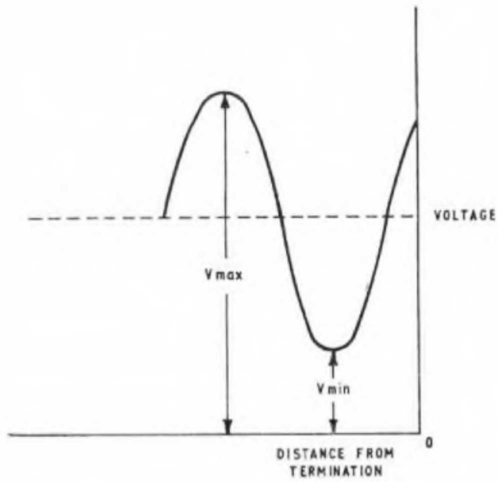
**Q. 10.** *Describe the circumstances in which a standing wave can arise on a transmission line, and state the meaning of standing wave ratio.*

*An h.f. transmission line of negligible loss has a characteristic impedance of* 600 *ohms and is terminated by an aerial. Calculate the standing wave ratio along the line when the aerial impedance is:*

(*a*) 500 *ohms,*

(*b*) (400 + j300) *ohms.*

**A. 10.** A standing wave can arise on a uniform transmission line if the terminating impedance is not matched to the characteristic impedance of the line. Because of the mismatch, some of the energy is reflected at the termination and travels back to the source. The resultant r.m.s. voltage at any point on the line is then the sum of the forward and reflected waves, the magnitude of the latter depending on the degree of mismatch. As the two waves are travelling in opposite directions they will be in phase at some points and in anti-phase at others, so that the resultant voltage varies with distance along the line forming a standing wave as shown in the sketch.




The ratio of the voltage (or current) at a point of maximum amplitude, to that at a point of minimum amplitude, is known as the *standing wave ratio.* In the sketch, the voltage standing wave ratio is $V_{max}/V_{min}$.

If a line of characteristic impedance $Z_O$ is terminated in an impedance $Z_R$, the voltage reflexion coefficient is

$$\rho = \frac{Z_R - Z_O}{Z_R + Z_O}.$$

and the standing wave ratio is

$$S = \frac{1 + |\rho|}{1 - |\rho|}.$$

(*a*) Substituting $Z_O = 600$ ohm and $Z_R = 500$ ohm, gives

$$\rho = \frac{500 - 600}{500 + 600} = -\frac{1}{11},$$

$$\text{and, } S = \frac{12}{11} \times \frac{11}{10} = \underline{1 \cdot 2.}$$

(Note that, in this case, since $Z_O$ and $Z_R$ are both wholly resistive, the expression simplifies to $S = Z_O/Z_R$.)

(*b*) Substituting $Z_R = (400 + j300)$ gives,

$$\rho = \frac{400 + j300 - 600}{400 + j300 + 600} = \frac{-2 + j3}{10 + j3}.$$

$$\therefore\ |\rho| = \frac{\sqrt{4 + 9}}{\sqrt{100 + 9}} = 0 \cdot 34,$$

$$\text{and, } S = \frac{1 \cdot 34}{0 \cdot 66} = \underline{2 \cdot 0.}$$

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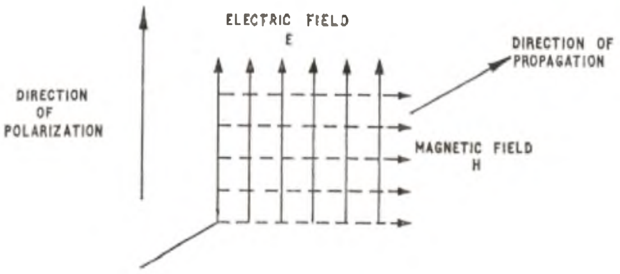
## BASIC MICROWAVE COMMUNICATIONS 1968

**Students were expected to answer any six questions.**

Q. 1. *Explain with the aid of a diagram what is meant by the polarization of a plane electro-magnetic wave. Describe how in microwave practice the polarization of a plane wave can be determined.*

*A horizontally polarized plane wave falls obliquely on a perfectly-conducting horizontal surface. Explain why reflexion occurs and account for any phase change in the reflected wave.*

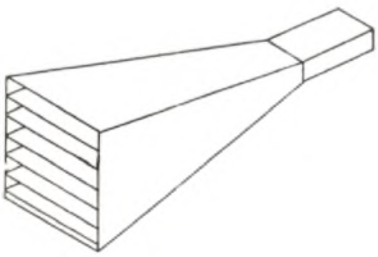
A. 1. At a fixed point in space a plane electro-magnetic wave consists of alternating electric and magnetic fields at right-angles to each other and in a plane which is perpendicular to the direction of propagation (see sketch (*a*)). The two fields must be in phase with




(*a*)

each other for the propagation of energy to take place. The direction of polarization is defined as that of the electric field and is usually either vertical or horizontal.

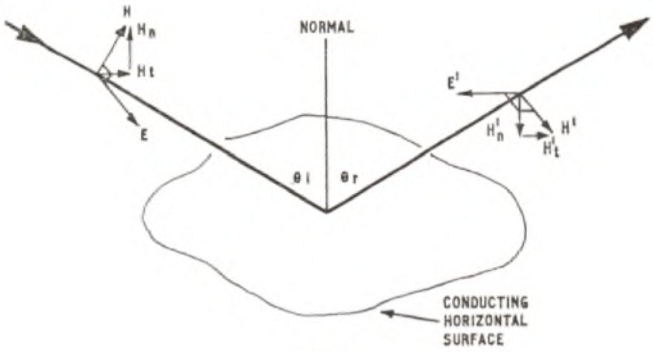
The polarization of a plane wave can be determined using a small dipole with a diode detector and micro-ammeter, or an open-ended waveguide which can be excited by the dominant mode only. As a refinement, a waveguide horn with a polarizing grating of narrow metal strips, can be added. The grating ensures that the horn accepts only the component of field polarized perpendicular to the metal strips (see sketch (*b*)). The waveguide, which is connected to a detector



(*b*)

indicating the field strength, is rotated in the transverse plane until a zero reading is obtained. The plane of polarization is then parallel with the metal strips of the grating.

When a wave falls on a perfectly-conducting surface the boundary conditions must be observed at the surface. In sketch (*c*) the electric field of the horizontally-polarized wave falls tangentially onto the horizontal reflecting surface. The boundary conditions state that the tangential electric field, *E*, must be zero close to the surface, so that an equal and opposite field ($E' = -E$) must be present to meet this condition. The magnetic field can be resolved into two components, $H_t$ tangential to the surface and $H_n$ normal to the surface. From




(*c*)

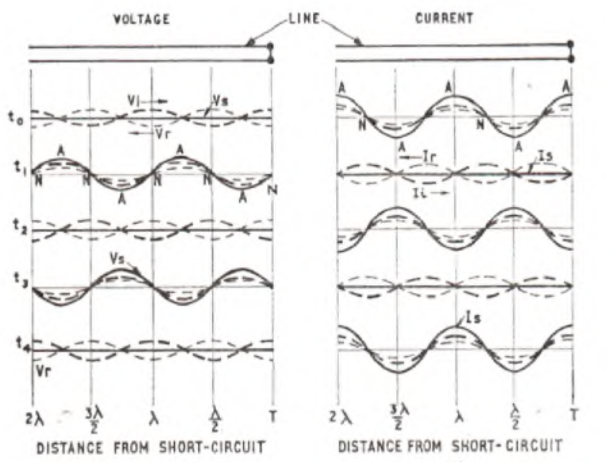
the boundary conditions, $H_t$ will produce a current in the skin of the surface at right-angles to the direction of $H_t$ which, in turn, will produce a magnetic field $H_t'$ displaced by a further 90 degrees to the current, such that $H_t'$ will be in antiphase with $H_t$. Similarly, a magnetic field cannot exist normal to a conducting surface and $H_n$ must also be balanced by an equal and opposite field $H_n'$. The components of the incident wave thus produce equal and opposite components which propagate a reflected wave in antiphase with the incident wave. The direction of reflexion will obey the optical laws of reflexion, i.e. the two rays and the normal to the surface at the point of reflexion lie in the same plane, and $\theta_i = \theta_r$.

Q. 2. *Describe with the aid of diagrams how a standing wave is set up in an open-wire transmission line terminated in a short-circuit. Indicate clearly the position of both voltage and current nodes and antinodes on the line for a distance of two wavelengths measured from the termination.*

*Describe an experiment to illustrate this phenomenon. State how this could be used to determine the frequency of the source used to energize the line.*

A. 2. Sketch (*a*) shows the progress of an incident voltage wave $V_i$ for times $t_0$ to $t_4$. The voltage at the short-circuit must always be zero. At time $t_0$ the incident voltage at the termination is at a maximum and is positive. This must be balanced by a reflected wave $V_r$ which, at this instant, is negative and at its maximum. At time $t_1$ the incident wave has progressed a quarter of a wavelength to the right while the reflected wave has moved a quarter of a wavelength to the left. At this instant both waves are in phase and add to produce the standing wave $V_s$, double the value of $V_i$ and $V_r$. After each wave has progressed a further quarter wavelength at time $t_2$, the voltages are in antiphase and the sum of the two waves is zero at all points. At $t_3$ both waves are again in phase and the standing wave resulting from the addition of the two components is at a maximum, but of opposite polarity to the wave at time $t_1$. At $t_4$ the two waves are again in antiphase and the standing wave is zero at all points on the line. These conditions are identical with those of $t_0$ and the process repeats itself. The points of maximum oscillation are called antinodes and are marked as A in the sketches. The points where the voltage is always zero are called nodes and are shown as N. The short-circuit terminations are designated T.

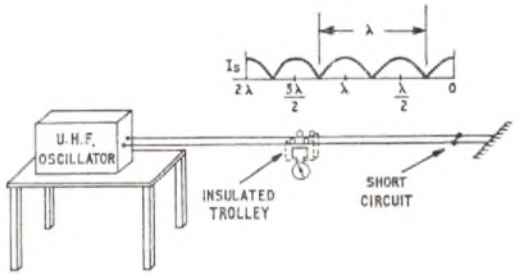
The current waves, shown in sketch (*b*), can be determined by considering conditions at the short-circuit. At the short-circuit termination the voltage must always be zero and the equal division of energy between the electric and magnetic fields is destroyed. All




(a) (b)

the energy is therefore stored in the magnetic field. The maximum current at the short-circuit is twice $I_i$ and $I_r$ and at time $t_0$ results in the standing current wave $I_s$. At time $t_1$ the incident current wave is zero at the termination and the amplitude of the reflected wave is also zero. The two waves are in anti-phase and the standing wave at all points is zero. The process continues as for the voltage waves, and the current antinodes and nodes have again been marked as A and N, respectively.

An experiment illustrating this phenomenon is shown in sketch (*c*).




(c)

An open-wire transmission line (Lecher line) is fed from a u.h.f. oscillator at a frequency of several hundred Megahertz. An insulated trolley can be moved along the line and holds a loop of wire close to one of the wires of the line. The loop is connected to a detector and micro-ammeter and as the trolley is moved along the line, the nodes and antinodes of current are shown on the micro-ammeter as zero and maximum readings, respectively.

The velocity of propagation will be practically that of light, at the frequency used. The wavelength, $\lambda$, is the distance between two alternate nodes (see sketch (*c*)) and the frequency of the source used to energize the line is given by:

$$f = \frac{c}{\lambda}.$$

Where $f$ = frequency (Hz),
$c$ = speed of light (m/sec),
$\lambda$ = wavelength (m).

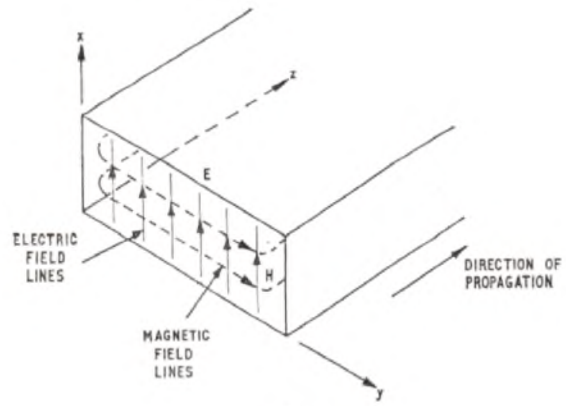
Q. 3. *Account for the fact that T.E.M. waves cannot be propagated in a hollow waveguide.*

*Calculate the lowest frequency at which it is possible to transmit energy in a hollow rectangular waveguide having internal dimensions* $7{\cdot}5$ *cm* $\times$ $3{\cdot}4$ *cm.*

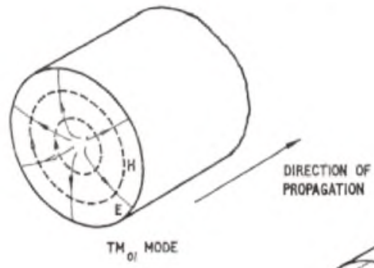
*In a carefully drawn three-dimensional diagram show the pattern of the E and H lines in a rectangular waveguide energized in the dominant mode.*

**A. 3.** The T.E.M. wave has both electric and magnetic fields in phase with each other, and at right-angles to the direction of propagation. An attempt to contain a plane-polarized T.E.M. wave in a rectangular guide results in the electric field being tangential to two sides of the guide and the magnetic field being normal to these sides. Neither of these boundary conditions are admissible and the magnetic field must curve round until it is tangential with the side wall, similarly the E-field must fall to zero at the side walls. Propagation is then in the $TE_{10}$ mode as shown in sketch (*a*). T.E.M. waves are propagated along a coaxial line but if the centre conductor is omitted, the transverse electric field can no longer terminate on the inner conductor and has to curve into the axis of the waveguide as shown in sketch (*b*).

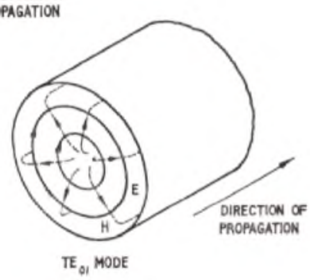
An attempt to reverse the relative positions of the magnetic and electric fields is shown in sketch (*c*). The magnetic field must




(a)




(b)

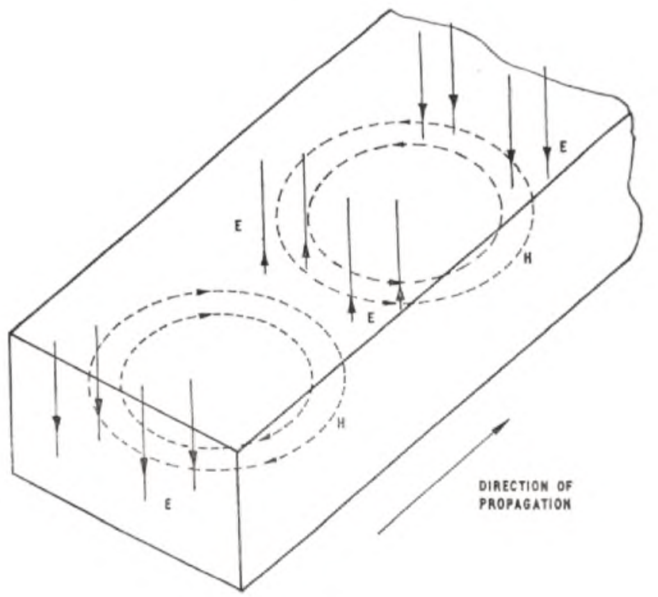



(c)

curve round and run parallel to the axis both at the walls and on the axis of the tube. Propagation is then in the $TE_{01}$ mode.

Alternatively, the field patterns which can be produced in hollow guides can be considered as the addition of two waves propagated at an angle to each other and successively reflected from the opposite sides of the guide. The direction of propagation is the bisector of this angle, and the propagated wave must inevitably contain a longitudinal component of either the electric or magnetic field. T.E.M. waves cannot therefore be propagated in a hollow waveguide.

The lowest frequency ($f_c$) at which it is possible to transmit energy




(d)

in a hollow rectangular waveguide is given when half a free-space wavelength is equal to the wide dimension of the guide.

$$f_c = \frac{c}{2a}.$$

Where, $f_c$ = critical frequency (Hz),
$c$ = speed of light (cm/sec),
$a$ = wide dimension of guide (cm).

Substitution of values gives,

$$f_c = \frac{3 \times 10^{10}}{2 \times 7 \cdot 5},$$
$$= 2 \times 10^9 \text{ Hz},$$
$$= \underline{2 \text{ GHz}}.$$

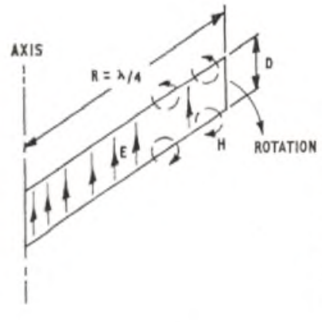
The dominant mode in a rectangular waveguide is the $TE_{10}$ mode and sketch (*d*) shows the corresponding pattern of the E and H lines.

*Q.* 4. *By regarding a hollow cylinder as a quarter-wave section of open-wire transmission line with a short-circuited termination, rotated about an axis through its open end, show that the cavity will resonate at a frequency determined by its dimensions.*

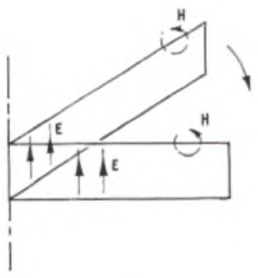
*Sketch, in separate diagrams, the patterns of the E and H fields, assuming that the depth of the cavity is very much less than its diameter.*

*Explain carefully why conventional LC resonant circuits cannot be used at microwave frequencies.*

A. 4. The quarter-wave section of open-wire transmission line is shown rotated about its axis in sketch (*a*). It will be assumed initially




(*a*)



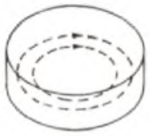

(*b*)

that its length, $R$, is large compared with the spacing, $D$, between the wires. The rotation of the line about the axis will generate a closed hollow cylinder and it can be seen in sketch (*b*) that the vertical components of the magnetic field will cancel to produce the magnetic field shown in sketch (*d*). There will be no magnetic field external to the cylinder, because the radial currents on the inside of the cylinder will balance the magnetic field which is inside the cylinder and tangential to the surface.

The resonant frequency of the cavity will be similar to that of the quarter-wave line which generated the cylinder, and, in particular, they are both inversely proportional to the radius. Further, the resonant frequency of the line will be independent of the spacing between the wires, where $R$ is very much greater than $D$, and likewise the resonant frequency of the cavity is independent of its depth $D$. The electric and magnetic fields in the cavity are shown in sketch (*c*) and sketch (*d*). The two fields are in time quadrature.

E FIELD

(*c*)



H FIELD

(*d*)

A conventional LC resonant circuit consists of an inductor and a capacitor. The values of both components could be reduced in order to increase the resonant frequency. To reach microwave frequencies, however, the inductor would have to be reduced to a plain piece of wire with a capacitance between its ends not negligible compared with the inductance, and the capacitor would have well separated plates of small area with the self-inductance of the plates and leads similarly not negligible. The resonant circuit would thus have distributed inductance and capacitance and no longer resemble a conventional LC circuit. It would also be difficult to achieve a low-Q circuit, required for wide-band applications, since the Q of a resonant circuit using a plain conductor as the inductor increases with the frequency ($f$). This phenomenon is due to the fact that at very high frequencies current flows close to the surface of a conductor, the depth of penetration being inversely proportional to $\sqrt{f}$. Thus, as the frequency is increased, the resistance, $R$, is proportional to $\sqrt{f}$, while the inductive reactance, $\omega L$, is proportional to $f$. The $Q = \frac{\omega L}{R}$ is therefore roughly proportional to $\sqrt{f}$.

*Q.* 5. *What is meant by an isotropic radiator?*

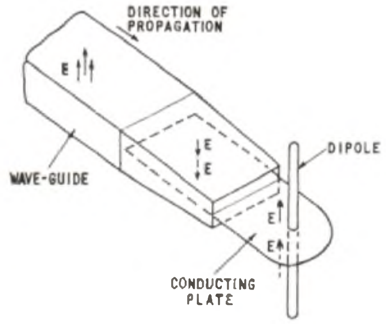
*A half-wave dipole has a power gain of* 1·64, *a beam width of approximately* 78 *degrees and a driving impedance of about* 72 *ohms. Explain what is meant by each of these parameters.*

*Describe, with the aid of a diagram showing the method of mounting, how a half-wave dipole radiator can be energized from a hollow rectangular waveguide carrying the dominant mode.*

A. 5. An isotropic radiator is an aerial which radiates uniformly in all directions.

The power radiated from a dipole is distributed nonuniformly, being greatest in directions normal to the dipole, and zero in line with the dipole. The ratio of the power radiated normal to the dipole, to the average power radiated, in terms of flux per unit solid angle, is termed the power gain, given here as 1·64. The beam width measured in a plane containing the dipole, is 78 degrees, this is the angle between the directions at which the radiated power has dropped to half its maximum value. The ratio of voltage to current at the input to the aerial represents the driving impedance, given as about 72 ohm. This is nearly equal to the radiation resistance of the aerial, which is the power radiated divided by the mean-square value of the current entering the aerial.

The sketch shows a method of energizing a half-wave dipole radiator




from a hollow rectangular waveguide carrying the dominant mode. The dipole is mounted on a conducting plate, which being perpendicular to the electric field has little effect on radiation. The electric field is split into two parts each of which energizes half the dipole in the correct relative phase. The waveguide is slightly tapered to improve matching.

*Q.* 6. *What is meant by the terms:*

(*a*) *instantaneous frequency,*
(*b*) *frequency deviation,*
(*c*) *modulation index,*

*as used in frequency modulation?*

*A 5 kHz (kc/s) tone of amplitude* 20 *volts drives an f.m. transmitter to its maximum permitted deviation with modulation index* 15. *For an* 8 *kHz modulating signal of amplitude* 12 *volts:*

(i) *calculate the frequency deviation and modulation index,*
(ii) *describe the frequency spectrum of the radiated signal.*

*What bandwidth is acceptable in practice for the transmitter?*

A. 6. (*a*) The instantaneous frequency of a frequency-modulated wave is the rate of change of phase measured in radians per second divided by $2\pi$, at any given instant.

(*b*) The frequency deviation is the peak value of the difference between the instantaneous frequency and the carrier frequency.

(*c*) The modulation index for sinusoidal waves is the ratio of the frequency deviation to the modulating frequency.

(*i*) The frequency deviation, $\Delta f = f_p m_f$

where, $f_p$ = modulation frequency,
$m_f$ = modulation index.

Substituting the values given,

$$\Delta f = 5 \times 15$$
$$= 75 \text{ kHz}.$$

The maximum permitted deviation is produced by a tone of amplitude 20 volts.

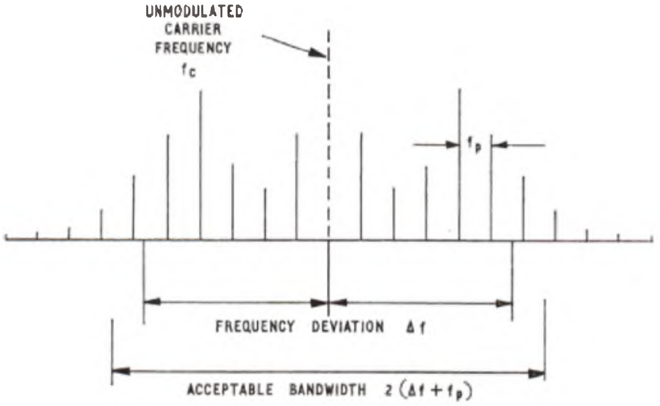
A tone of amplitude 12 volts will cause a frequency deviation of

$$75 \times \frac{12}{20} = \underline{45 \text{ kHz.}}$$

For an 8 kHz modulating signal the modulation index $m_f = \frac{45}{8}$

$$= \underline{5 \cdot 625}.$$

(*ii*) The frequency spectrum of the radiated signal is shown in the sketch. It consists of an infinite series of pairs of sidebands placed




at multiples of the modulating frequency ($f_p$) away from the carrier. Most of the energy is contained within the frequency range $f_c \pm \Delta f$ and the amplitude of the sidebands falls away rapidly beyond this range. The amplitude of the carrier will be practically zero when the modulation index is 5·6.

A bandwidth of $2(\Delta f + f_p)$ is acceptable for commercial telephony and assuming for the present transmitter a maximum modulating frequency of 10 kHz and a maximum deviation of 75 kHz, a bandwidth of 170 kHz would be required.

*Q.* 7. *Show that if a receiver has a gain G and a noise factor N, then the product G (N — 1) is proportional to the internal noise. Deduce an expression for the output noise power from a receiver in terms of its gain, bandwidth and noise factor when the input is white noise.*

*Briefly explain why high gain with low noise factor is of greater importance in the input stage than in subsequent stages of a receiver.*

**A. 7.** The available thermal noise power from a source (white noise) $= kTB$,

where, $k$ = Boltzman's constant (Joules/°K),
$T$ = absolute temperature (°K),
$B$ = bandwidth (Hz).

By definition, the noise factor,

$$N = \frac{\text{actual output noise power}}{\text{theoretical minimum output noise power}},$$

$$= \frac{N_0}{kT_0BG}, \qquad \ldots\ldots (1)$$

where, $N_0$ = actual output noise power,
$T_0$ = working temperature = 290° K,
$G$ = power gain.

The output noise power is composed of amplified input noise ($kT_0BG$) and amplified internal noise, so that the internal noise power, $N_i$, is the difference given by

$$N_i = N_0 - kT_0BG,$$
$$= (N-1)\,kT_0BG \qquad \ldots\ldots (2)$$

i.e. $\underline{\text{internal noise} \propto G(N-1)}$.

The output noise power of a receiver when the input is white noise, from equation (1), is

$$\underline{N_0 = NkT_0BG}.$$

Let the noise factor of the input stage be $N_1$, and of the subsequent system $N_2$, and the corresponding power gains be $G_1$ and $G_2$. It is assumed that the overall bandwidth, $B$, is limited by that of the subsequent system.

The input termination supplies a noise power $kTB$ which is amplified and appears at the output as $kTBG_1G_2$.

From equation (2) the internal noise from stage 1 is $(N_1 - 1)\,kTBG_1$, and appears at the output of the system as $(N_1 - 1)\,kTBG_1G_2$.

The overall noise factor $N_{12}$ is the ratio of actual output noise power to the output noise power originating from the input termination.

$$\therefore N_{12} = \frac{\text{total noise output}}{kTBG_1G_2},$$

$$= \frac{kTBG_1G_2 + (N_1 - 1)\,kTBG_1G_2 + (N_2 - 1)\,kTBG_2}{kTBG_1G_2},$$

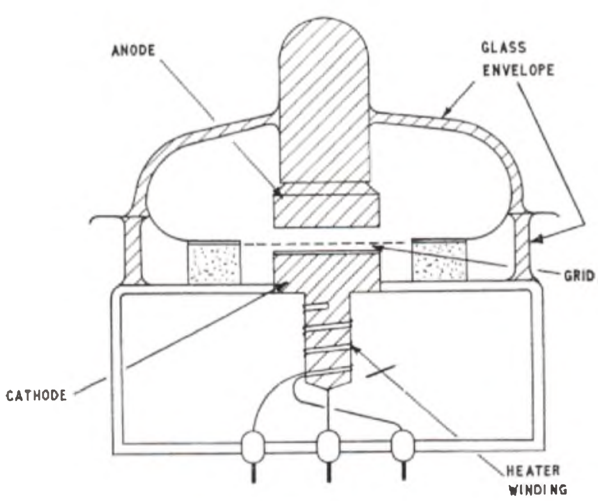
$$= N_1 + \frac{N_2 - 1}{G_1}. \qquad \ldots\ldots (3)$$

Equation (3) shows that to achieve a low overall noise factor, it is essential that the noise factor of the first stage, $N_1$, should be low, and its gain, $G_1$, should be high. The effects of the noise factor of the subsequent system, $N_2$, is reduced by a factor of at least $1/G_1$ compared with the noise factor of the input stage, $N_1$. The gain of the subsequent system, $G_2$, has no effect on the overall noise factor.

*Q.* 8. *Describe, with the aid of a diagram, the construction of a disc-seal (planar) triode. By means of a circuit diagram, show how this type of valve is used in a cascode amplifier.*

*In the first stage of a microwave receiver, what are the advantages of this circuit over (a) a pentode, and (b) a grounded-grid triode amplifier?*

**A. 8.** Sketch (*a*) shows the construction of a disc-seal triode. The




(*a*)

cathode is of nickel, ground flat and sprayed with an oxide coating to a fine tolerance. It has a separate heater winding.

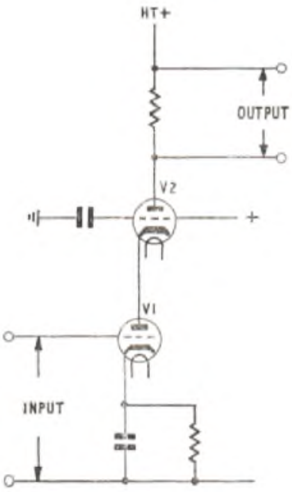
The grid is tightly wound round a grid-frame, using fine wire at a spacing of about 1,000 turns per inch. The grid-cathode spacing is a fraction of a thousandth of an inch. The anode is set to a distance of about 0·01 in from the grid. The mounting is arranged so that the spacings can be adjusted, and the whole assembly is compressed together with springs to allow for differential thermal expansion. The unit is sealed by glass and is finally evacuated, baked and sealed. The close spacing of the electrodes is essential to reduce the effect of transit time at high frequencies. The disc-seal arrangement reduces the inductance of the leads to the electrodes, and in practice the discs form part of the resonant cavities in which the valve is mounted.

The principle of the cascode amplifier is shown in sketch (*b*). The input stage, V1, is a triode with an earthed cathode, and its anode is connected to the cathode of V2 which has its grid earthed to r.f. currents.

The grounded-grid stage, V2, reduces the capacitive feedback from output to input, without introducing the partition noise which would be produced by the screen current in a pentode. The effect of internally generated shot noise in V2 is made insignificant by the anode resistance of V1, which provides negative feedback for V2.

The noise factor of the complete circuit approaches the theoretical value for V1 used as a triode, but without the undesirable effect of anode-grid capacitance.

(*a*) The cascode amplifier has a lower noise factor than a pentode amplifier, which suffers from partition noise produced by the screen grid. The high gain available from a pentode is unusable at microwave frequencies, because the resonant cavities which have to be used are essentially low-impedance devices. The disc-seal triode has the




advantage that no screen-grid supply is necessary. Nothing is lost by the omission of the suppressor grid, because the r.f. output must be relatively low to obtain optimum efficiency.

(*b*) The input impedance to the cascode circuit is purely capacitive, whereas the grounded-grid triode draws power from the source, this may reduce the $Q$ of the input resonant cavity, an important factor if a high $Q$ is required.

*Q.* **9.** *Explain in detail why the gain of a simple R-C coupled amplifier stage decreases at frequencies above the audio range. By considering an input train of rectangular pulses of such an amplifier in terms of its fundamental and harmonic components, show the effect of reduced high-frequency gain.*

*With the aid of diagrams, describe and explain how high-frequency compensation is provided in a video amplifier.*

*Q.* **10.** *The input signal to the circuit shown is a train of square pulses as indicated in Fig. 1.*

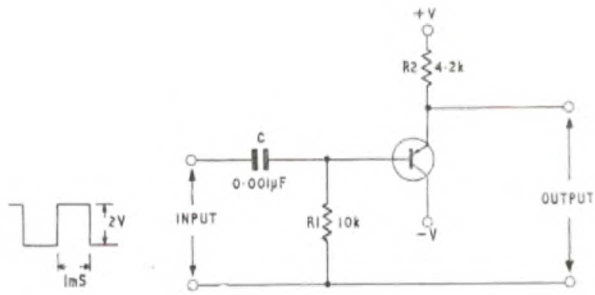



Fig. 1

*Accompanied by a brief explanation, draw to scale a diagram showing input and output signal waveforms. Calculate the instantaneous output voltage 10 μs after the instantaneous rise of the input voltage, assuming that the output can rise to 0·7 volt.*

*Why, for a device of this nature, should the input be taken from a source having a low shunt capacitance?*

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## LINE TRANSMISSION C, 1968

**Students were expected to answer any six questions**

*Q.* **1.** *Write down the expressions for the characteristic impedance and the propagation coefficient of a uniform transmission line in terms of its primary coefficients. Hence derive the expression for the attenuation coefficient and the phase-change coefficient. State clearly the units commonly used.*

*A coaxial cable has a loss of 5 dB/mile at 1 MHz (Mc/s) and of this loss 1 dB/mile is attributable to the dielectric. Estimate the loss at 4 MHz, stating clearly the assumptions made.*

**A. 1.** For a uniform transmission line the characteristic impedance, $Z_0$, may be expressed as:

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \text{ ohms} \qquad \ldots\ldots (1)$$

and the propagation coefficient $\gamma$ as:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \qquad \ldots\ldots (2)$$

where the primary coefficients are $R$ ohms/mile, $L$ henrys/mile, $G$ mhos/mile and $C$ farads/mile.

The propagation coefficient is a complex quantity and $\gamma = \alpha + j\beta$. The real part $\alpha$ is the attenuation coefficient in nepers/mile and the imaginary part $\beta$ is the phase-change coefficient in radians/mile.

Equation (2) can be expressed in polar form:

$$\gamma = \sqrt[4]{((R^2 + \omega^2L^2)(G^2 + \omega^2C^2))} \;\Big/ \frac{1}{2}\left[\tan^{-1}\frac{\omega L}{R} + \tan^{-1}\frac{\omega C}{G}\right].$$

$$\text{Thus, } |\gamma| = \sqrt[4]{((R^2 + \omega^2L^2)(G^2 + \omega^2C^2))},$$

$$\text{but, } \gamma = \alpha + j\beta, \text{ and } |\gamma| = \sqrt{\alpha^2 + \beta^2}.$$

$$\text{Thus, } \sqrt{\alpha^2 + \beta^2} = \sqrt[4]{((R^2 + \omega^2L^2)(G^2 + \omega^2C^2))}.$$

$$\therefore\ \alpha^2 + \beta^2 = \sqrt{(R^2 + \omega^2L^2)(G^2 + \omega^2C^2)}. \qquad \ldots\ldots (3)$$

$$\text{But, } \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}.$$

$$\therefore\ \alpha^2 + 2j\alpha\beta - \beta^2 = RG + j\omega LG + j\omega CR - \omega^2LC.$$

Equating real parts,

$$\alpha^2 - \beta^2 = RG - \omega^2LC. \qquad \ldots\ldots (4)$$

Adding equations (3) and (4) gives,

$$2\alpha^2 = \sqrt{(R^2 + \omega^2L^2)(G^2 + \omega^2C^2)} + (RG - \omega^2LC).$$

$$\therefore\ \alpha = \underline{\sqrt{\tfrac{1}{2}[\sqrt{(R^2 + \omega^2L^2)(G^2 + \omega^2C^2)} + (RG - \omega^2LC)]} \text{ nepers/mile}}$$

Subtracting equation (4) from equation (3) gives,

$$2\beta^2 = \sqrt{(R^2 + \omega^2L^2)(G^2 + \omega^2C^2)} - (RG - \omega^2LC).$$

$$\therefore\ \beta = \underline{\sqrt{\tfrac{1}{2}[\sqrt{(R^2 + \omega^2L^2)(G^2 + \omega^2C^2)} - (RG - \omega^2LC)]} \text{ radians/mile}}$$

For a coaxial cable working at 1 MHz it can be assumed that,

$$\alpha \simeq \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{C}{L}} \text{ nepers/mile.}$$

(See A. 1, Line Transmission C, 1966, *Supplement*, Vol. 60. p. 28, July 1967.)

In this expression the first term (containing $R$) represents the series loss and the second term (containing $G$) represents the shunt or dielectric loss. At high frequencies it can be assumed that the skin effect is fully operative and that series loss is proportional to $\sqrt{f}$ where $f$ is the frequency. It can further be assumed that the dielectric loss is proportional to $f$.

Therefore, total loss = series loss + shunt (or dielectric) loss,

$= k_1\sqrt{f} + k_2 f$ where $k_1$ and $k_2$ are constants.

For the coaxial cable at 1 MHz:

Total loss = 5 dB/mile,

dielectric loss = 1 dB/mile = $k_2 \times 10^6$.

$\therefore$ series loss = 4 dB/mile = $k_1 \times 10^3$.

Hence, $k_1 = 4 \times 10^{-3}$, and $k_2 = 1 \times 10^{-6}$.

Hence at 4 MHz:

$$\text{total loss} = 4 \times 10^{-3} \times \sqrt{4 \times 10^6} + 1 \times 10^{-6} \times 4 \times 10^6,$$

$$= 8 + 4,$$

$$= \underline{12 \text{ dB/mile}}.$$

*Q.* **2.** *A cable pair has the following primary coefficients,*

| | |
|---|---|
| *$R$ = 40 ohms/mile* | *$L$ = 1 mH/mile* |
| *$G$ is negligible* | *$C$ = 0·1 μF/mile* |

*This pair is to be loaded, using 88-mH coils spaced 1·136 miles apart, in order to reduce the attenuation over part of the frequency range.*

*Calculate the attenuation coefficient of the pair at a frequency of 5,000/2π Hz (c/s) before and after loading, if the total series resistance of each coil is 5 ohms.*

*Estimate the cut-off frequency of the loaded pair and explain how* *this would be affected if, owing to an error in installation, one coil were omitted and the pair jointed straight through instead.*

A. 2. The attenuation $\alpha$ may be derived from the propagation coefficient $\gamma$ where:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}.$$

$$\text{Thus, } \gamma^2 = (R + j\omega L)(G + j\omega C).$$

Before loading:

$$\begin{aligned}\gamma^2 &= (40 + j5{,}000 \times 1 \times 10^{-3})(0 + j5{,}000 \times 0\cdot1 \times 10^{-6}),\\ &= j40 \times 5{,}000 \times 10^{-7} - 5 \times 5{,}000 \times 10^{-7},\\ &= j20 \times 10^{-3} - 25 \times 10^{-4},\\ &= 10^{-2}(j2 - 0\cdot25),\\ &= 2 \times 10^{-2} \,\underline{/97\cdot2^\circ}.\end{aligned}$$

$$\therefore\ \gamma = 0\cdot14 \,\underline{/48\cdot6^\circ}.$$

$$\text{But } \gamma = \alpha + j\beta.$$

$$\therefore\ \alpha = 0\cdot14 \cos 48\cdot6^\circ \text{ and } \beta = 0\cdot14 \sin 48\cdot6^\circ.$$

Hence, $\alpha = 0\cdot14 \times 0\cdot6613 = \underline{0\cdot093}.$

After loading:

$$R' = 40 + \frac{5}{1\cdot136} = 44 \text{ ohms/mile},$$

$$L' = 1 + \frac{88}{1\cdot136} = 79 \text{ mH/mile},$$

where $R'$ and $L'$ are the effective primary coefficients after loading.

$$\text{Thus } \gamma^2 = (44 + j5{,}000 \times 79 \times 10^{-3})(0 + j5{,}000 \times 10^{-7}).$$

$$\therefore\ \gamma = 0\cdot446 \,\underline{/86\cdot8^\circ},$$

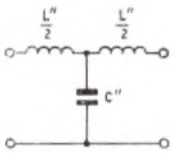
$$\text{and, } \alpha = 0\cdot446 \times 0\cdot0558 = \underline{0\cdot0249.}$$

[*Note.* After loading it would have been permissible to use the approximate expression:

$$\alpha = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{C}{L}}$$

since $L$ is comparatively high.]

The cut-off frequency, $f_c$, is determined by the capacitance $C''$ of a loading section and the inductance $L''$ of a loading coil as shown in the sketch.




$$f_c = \frac{1}{\pi\sqrt{L''C''}} \text{ Hz}.$$

$$\text{Thus, } f_c = \frac{1}{\pi\sqrt{88 \times 10^{-3} \times 0\cdot1 \times 1\cdot136 \times 10^{-6}}},$$

$$= \underline{3{,}180 \text{ Hz}}$$

If a loading coil were to be omitted the capacitance $C''$ would be doubled and that part of the line would become a low-pass filter with a cut-off frequency which was lower than that of the remainder by a factor of $\frac{1}{\sqrt{2}}$, i.e. 0·707.

Therefore, new cut-off frequency, $f_c'$, would be given by:

$$f_c' = \frac{1}{\pi\sqrt{88 \times 10^{-3} \times 0\cdot2 \times 1\cdot136 \times 10^{-3}}},$$

$$= \underline{2{,}260 \text{ Hz}}.$$

*Q. 3. What are the causes of crosstalk between pairs in a multipair cable? Explain how such cables are balanced and utilized for (a) audio working, and (b) carrier working.*

A. 3. Crosstalk may be caused by:

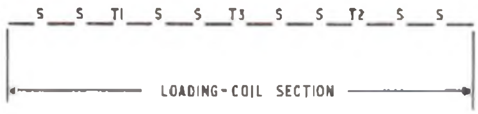
(*i*) capacitance unbalance,
(*ii*) resistance unbalance,
(*iii*) inductive coupling,
(*iv*) low insulation, and
(*v*) wire-to-wire contacts.

(*i*), (*ii*) and (*iii*) can be minimized by close control in manufacture and by subsequent installation procedures. Causes (*iv*) and (*v*) are matters of sound cable design and good maintenance.

Capacitance unbalance is generally the most serious cause of crosstalk and on installation the pairs or quads are tested and jointed in such a way as to reduce unbalance to a minimum.

For notes on cable manufacture and on the balancing and utilization of carrier cables see A. 4, Line Transmission C, 1967, *Supplement*, Vol. 61, p. 31, July 1968.

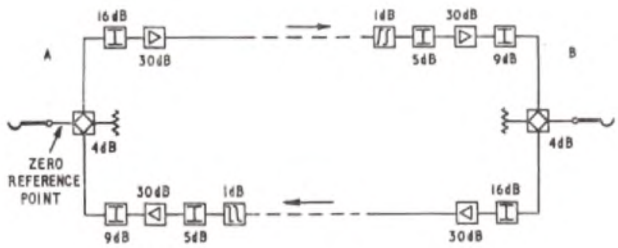
For audio cables it is usual to use the standard loading coil section as the unit for balancing and the jointing arrangement is shown in the sketch. This assumes the section to be made up from 12 manufactured




lengths each of about 176 yards. The lengths are first jointed in groups of three by systematic joints (marked S). These are made to a pre-determined schedule whereby quads which are adjacent in one length are jointed to quads which are remote in the next length thus reducing the accumulated crosstalk coupling between quads. The groups of three are then tested for capacitance unbalance from points T1 and T2 and from the results a selected joint is designed for each. Selection involves the choice of quads on either side of the joint which when joined together (with crosses between wires if required) give the minimum of resultant capacitance unbalance. Finally the process is repeated by testing, selecting and jointing at point T3.

*Q. 4. An audio-frequency 4-wire circuit with 2-wire ends is to be provided between terminal repeater stations 16 miles apart. The interconnecting cable has an attenuation of 1·5 dB/mile at 3·4 kHz (kc/s) and 0·4 dB/mile at 0·3 kHz. Draw a block schematic diagram of the complete circuit and a level diagram for one direction of transmission, assuming the circuit to be lined up to 3 dB loss between the 2-wire ends.*

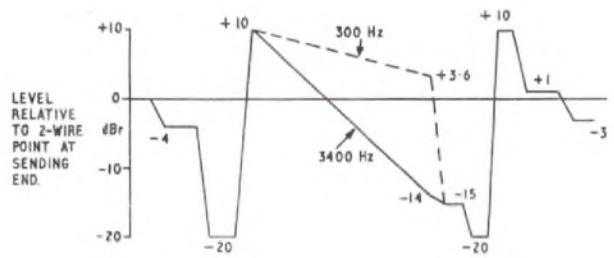
A. 4. Sketch (*a*) shows two terminal repeater stations A and B.




(*a*)

Each has a 4-wire repeater and a 2-wire to 4-wire termination with hybrid transformer and balancing network. Each amplifier has a fixed gain of 30 dB and an adjustable attenuator is connected to its input to allow the overall gain to be set to a particular figure. The attenuation/frequency characteristic of each line equalizer compensates the characteristic of the cable thus making the attenuation of the whole circuit independent of frequency.

Sketch (*b*) shows the level diagram for the A to B direction of




(*b*)

transmission. That for the B to A direction would be identical but reversed in sense. The theoretical loss from 2-wire to 4-wire (or 4-wire to 2-wire) for a hybrid transformer when balanced is 3 dB, and an extra 1 dB allows for the ordinary transformer losses and imperfections in the balance. A loss of 1 dB at 3,400 Hz has been allowed for the equalizer. In this case it would be designed for a loss of 18·6 dB at 300 Hz.

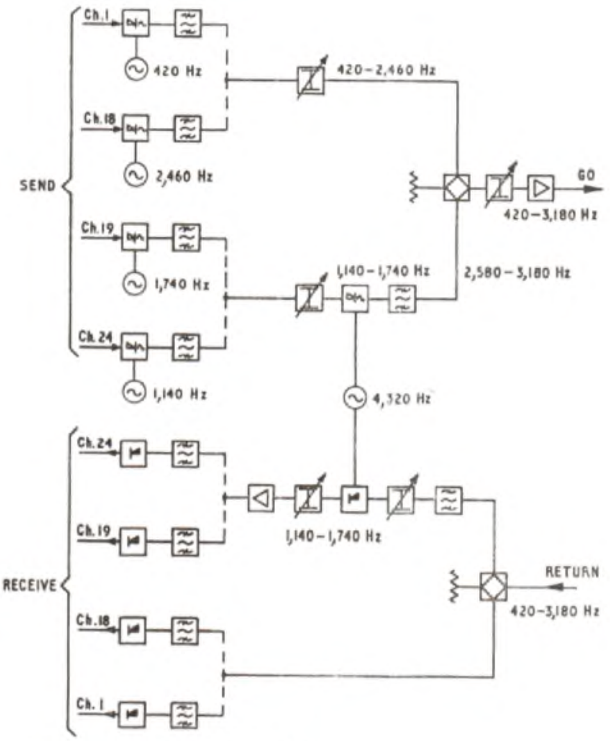
The maximum level at any point is +10 dBr and the minimum is −20 dBr. These are the usual figures adopted in planning, higher levels might lead to crosstalk to other channels whilst lower levels might give rise to a poor signal-to-noise ratio. A separate 9 dB attenuator at the output of the receiving amplifier is necesseary to give the required level of −3 dBr at the 2-wire point. This could not be achieved by increasing the value of the attenuator at the amplifier input without dropping the input level below the −20 dBr allowed.

*Q. 5. Draw a block schematic diagram for one end of a 24-channel amplitude-modulated voice-frequency telegraph system and explain its operation.*

*Explain what factors influence the choice of the carrier frequencies and the spacing between channels.*

*Show how the band-pass filter requirements differ from those used in a carrier telephone system.*

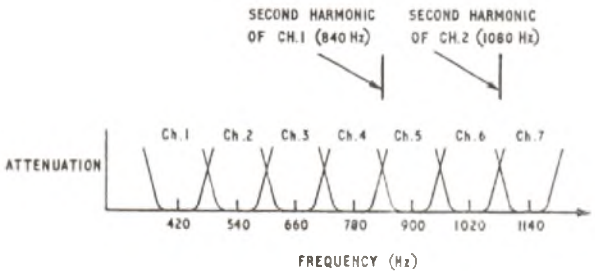
A. 5. The block schematic diagram for one end of a typical 24-channel amplitude-modulated voice-frequency telegraph system is shown in sketch (*a*). This is basically an 18-channel system in the




(*a*)

frequency range 420 Hz to 2,460 Hz plus a six-channel system in the range 1,140 Hz to 1,740 Hz. A group modulator at 4,320 Hz converts the latter to a higher range so that the whole 24-channel system has line frequencies spaced 120 Hz apart between 420 Hz and 3,180 Hz. The detailed operation is described in A. 4, Telegraphy C, 1960, *Supplement*, Vol. 54, p. 31, July 1961.

For a telegraph system working at 50 bauds a channel width of 120 Hz is required and the chosen frequencies are all odd harmonics of 60 Hz. Use of odd harmonics minimizes the possibility of inter-channel interference because the second harmonic of any channel frequency falls midway between two higher channels as shown in sketch (*b*).




(*b*)

In a carrier telephony system it is economic to make the best use of the bandwidth available by employing filters which are comparatively expensive but which have a very sharp cut off. In a telegraph system such filters are not justified and a more gradual cut off is acceptable. For further information see A. 10, Line Transmission C, 1966, *Supplement*, Vol. 60, p. 32, July 1967.

*Q. 6. Describe the tests which would be necessary to determine the probable transmission performance of a new design of telephone intended for use in an existing network.*

*What is the effect of sidetone on (a) the speaker, and (b) the listener at the distant end?*

A. 6. When assessing a new design of telephone for use in an existing network it is essential to consider its performance when connected to any existing type of exchange transmission-bridge and when switched through to any other existing telephone set. Objective tests are essential during the design stages but they must be supported by subsequent subjective tests which take account of human factors.

*Objective tests*

The response of the transducers themselves can be tested in the laboratory. For the transmitter this may be done at various frequencies by using a small loudspeaker issuing a pure tone in front of the transmitter and measuring its electrical output at various values of line-feeding current over a range of sound pressures from the loudspeaker.

For the receiver, an artificial-ear is used to simulate the acoustic-impedance of the human ear and thus to load the receiver correctly. The artificial-ear is equipped with a microphone to measure the sound pressure and thus the response of the receiver at various frequencies can be assessed. These tests can be repeated using the complete telephone set connected to a variety of artificial exchange-lines and transmission-bridges and measuring the sending and receiving response relative to the exchange termination.

*Subjective testing*

A method which has been in use for many years involves the assessment of articulation by setting up a telephone connexion in the laboratory and using trained testing crews to transmit meaningless words (logatoms) over a circuit having an adjustable attenuation. The quality of the telephone set is judged by the percentage of meaningless words which are received correctly at the distant end. Various degrees of impairment (such as room noise) can be applied to simulate real conditions of use. This method takes account of the human element but nevertheless it remains somewhat artificial as the level of speech and the position of the speaker's mouth relative to the transmitter are closely controlled.

In recent years an improved method has been developed in which pairs of representative subjects are asked to use a telephone circuit and to give an opinion as to its quality. They may be asked to solve simple puzzles by passing information over the circuit or they may be asked merely to converse. These methods have the merit of being realistic because the subjects are free to hold the telephone handset in the way they prefer and to speak at any level they wish.

*Effect of sidetone*

When a telephone is in use, the speaker hears his own words very faintly in his own receiver. The level of this sidetone is a matter for telephone circuit design but it depends to some extent upon the degree of impedance matching between the telephone set and the exchange line. Sidetone has a reassuring effect upon the speaker because it gives him a feeling that the telephone is working properly. However, if the sidetone is excessive, he will gain the impression that he is speaking too loudly and he will lower his voise to the possible detriment of the listener at the distant end. The listener generates no sidetone because he himself is silent, but he will hear as sidetone the local room noise picked up by his own transmitter.

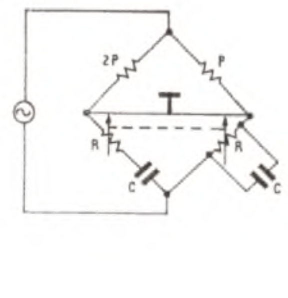
*Q. 7. Describe a typical national country-wide telephone network, showing the maximum permissible transmission loss allowable in the various parts of the system.*

*What factors are important from the transmission viewpoint in the design of the line between the customer and his local exchange?*

A. 7. For a description of a typical national country-wide telephone network see A. 5, Line Transmission C, 1965, *Supplement*, Vol. 59, p. 49, October 1966.

*Q. 8. Draw the circuit of an a.c. bridge designed for the measurement of audio frequencies. Derive the expression for the measured frequency in terms of the bridge components and explain what precautions should be taken to ensure an accurate result.*

A. 8. Any bridge which only balances at a single frequency can be used to measure the frequency of the source provided the value of each component of the bridge is known. The sketch shows the



arrangement of the Wien frequency-bridge which was specifically designed for the purpose and is in common use for frequency measurements.

The fixed resistance arms are in the ratio 2 : 1. The two variable resistance arms are equal in value and ganged so that they can be adjusted simultaneously to achieve a balance. The capacitors are of the same fixed value which can be changed to alter the frequency range of the bridge.

At balance:

$$\frac{2P}{\frac{1}{R}+j\omega C} = P\left(R+\frac{1}{j\omega C}\right). \qquad \ldots\ldots (1)$$

$$\therefore \frac{2P}{P} = \left(R+\frac{1}{j\omega C}\right)\left(\frac{1}{R}+j\omega C\right).$$

$$\therefore 2 = 1 + \frac{1}{j\omega CR} + j\omega CR + 1. \qquad \ldots\ldots (2)$$

Hence, $-\frac{1}{j\omega CR} = j\omega CR,$

giving $\omega^2C^2R^2 = 1.$

$$\therefore f = \frac{1}{2\pi CR}.$$

The reason for using the ratio 2 : 1 for the fixed resistance arms can be seen by putting $X$ and $Y$ in place of $2P$ and $P$. Then equation (2) becomes,

$$\frac{X}{Y} = 1 + \frac{1}{j\omega CR} + j\omega CR + 1.$$

Equating imaginary parts gives:

$$-\frac{1}{j\omega CR} = j\omega CR, \quad \text{as before.}$$

Equating real parts gives:

$$\frac{X}{Y} = 2.$$

Both real and imaginary parts must be equal for balance.

In order to ensure an accurate result the bridge components should be of high quality and properly calibrated. Also, a balanced and screened transformer should be interposed between the source whose frequency is to be measured and the bridge so as to avoid errors due to stray capacitance. Detection of the balance point can be made easier by including a band-pass filter in the connexion to the source or the detector in order to exclude harmonic interference.

*Q.* **9.** *Explain, with diagrams, how you would measure the insertion loss/frequency characteristic of an audio-frequency line equalizer.*

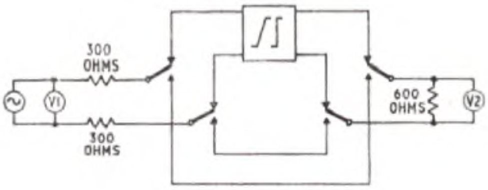
*Describe carefully the steps you would take to ensure reasonable accuracy of measurement.*

A.9. The insertion-loss of a line-equalizer is the ratio between the power delivered to the load before the equalizer is inserted into the line and the power delivered to the load after the equalizer has been inserted into the line. This loss can be measured directly by using the circuit shown in sketch (*a*). The oscillator output can be switched alternately between the direct path and the path through the equalizer. Then, if the oscillator output is adjusted before each reading to give a constant input voltage $V_1$ we have:

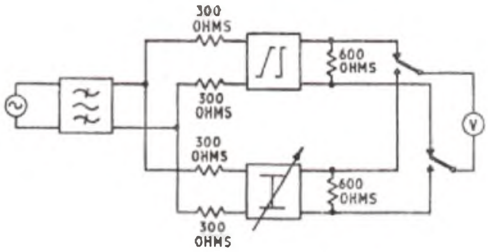
$$\text{Insertion loss} = 20\log_{10}\frac{V_2\,(\text{direct})}{V_2\,(\text{via equalizer})}.$$

The accuracy of this method depends upon the calibration and stability of the output voltmeter.

Sketch (*b*) shows another method which depends upon comparison




(*a*)




(*b*)

between the line-equalizer and an adjustable attenuator. In this case the output voltmeter is switched alternately between the equalizer and the adjustable attenuator. The attenuator is then altered until the output from each path gives the same reading on the meter. The accuracy of this method depends only on the calibration of the adjustable attenuator; calibration and stability of the output voltmeter are unimportant.

In each sketch a balanced 600-ohm line-equalizer is shown. For both methods of measurement it is important to avoid errors due to harmonics of the testing frequency and this may be done by using a voltmeter which is frequency selective or by connecting a band-pass filter to the oscillator output. The voltmeters should have a high impedance compared with 600 ohms.

*Q.* **10.** *Describe, with sketches, how racks of transmission equipment are interconnected and explain the precautions taken to minimize mutual interference between input circuits, output circuits and power supplies.*

*In recent years it has been possible to reduce the space occupied by transmission terminal equipment. Explain the developments which have allowed this to be done.*

A. **10.** A description of typical transmission equipment assembly is given in A. 3, Line Transmission C, 1962, *Supplement*, Vol. 56, p. 22, July 1963.

When designing cable runs between individual racks it is essential to use separate cables for input circuits, output circuits and power supplies and to space them apart as far as possible. Power supplies should be decoupled to avoid mutual interference caused by internal impedance of the power source.

In recent years the increasingly widespread use of transistors and miniaturized components has enabled the space occupied by transmission equipment to be reduced. Introduction of the transistor in place of the thermionic valve has led to a much lower power consumption and it has thus been possible to use smaller components and to mount more assemblies on a single rack without risk of over-heating. Power supplies have been simplified as separate high tension and low tension sources are no longer needed.

## ELEMENTARY TELECOMMUNICATIONS PRACTICE, 1969

**Students were expected to answer any six questions**

*Q.* **1.** *A local automatic telephone area has three exchanges, one large and two small. Explain in general terms:*

(*a*) *a possible numbering scheme to route the local traffic within the area,*
(*b*) *a possible local-junction line-plant layout,*
(*c*) *the factors which determine the number of circuits to be provided in the local-junction line plant,*
(*d*) *arrangements which may be adopted to check that the local-junction line-plant provision is adequate to carry the traffic,*
(*e*) *how the long-distance traffic outgoing from and incoming to the local area may be dealt with.*

A. **1.** Assume the three exchanges to be A (large) and B and C (small).

(*a*) A common (linked) numbering scheme could apply to the local area, each exchange being identified by the initial digit(s). By way of example, assume exchange A to be 5,000 lines, and exchanges B and C to be 800 lines each, allowing for growth. A 4-digit local-numbering scheme could be used, and typically, exchange A could be 2XXX–6XXX, exchange B 7XXX, and exchange C 8XXX, the numbering scheme for the whole area being 2XXX–8XXX. In all cases the initial digit would identify the called exchange and would determine the routing of traffic within the area.

(*b*) It is usually uneconomic to fully interconnect all the exchanges in an area, and tandem switching is usually adopted. Here shorter, fewer and more efficiently worked junction circuits take the place of individual links interconnecting all the exchanges. Additional switching equipment is required, but in general the cost of this is outweighed by the saving in junction circuits.

In the case given, exchange A could be a tandem switching centre. The local-junction line-plant layout would then consist of junction links A-B and A-C. Traffic B-C (and C-B) would be tandem switched on A. Should there be a significant volume of traffic B-C (or C-B), an alternative arrangement would be to provide a direct B-C junction link, in addition to the A-B and A-C links, and to dispense with the tandem switching on A. Tandem switching on the main large exchange is, however, the more usual economic arrangement.

(*c*) Line-plant provision is determined by the magnitude of the telephone traffic to be carried expressed in traffic units, which in turn is determined by the number and duration of calls (usually during the busy hour) to be carried by the line plant. In practice, this means the number of simultaneous calls likely to arise during the busy hour.

(*d*) For economic reasons, a certain proportion of calls (the grade of service) is allowed to fail during the busy hour due to insufficient switching plant or line plant. Overflow meters (e.g. connected to the eleventh step of two-motion selectors giving access to the junction circuits) indicate the number of calls lost due to inadequate line-plant provision. The readings are compared with standard data which indicates the readings which would give the minimum permissible grade of service for the condition concerned. If this grade of service is below the standard, additional line plant may be necessary.

(*e*) Telecommunication networks usually consist of a long-distance network and a number of local networks. The interface between the two consists of a number of group switching centres (GSCs), each of which deals with the long-distance traffic outgoing from, and incoming to, one or more local centres. The GSCs have access to both the long distance and relevant local networks, and incorporate special switching equipment to deal with the long-distance traffic.

Assuming that subscriber trunk dialling applies, the subscriber's trunk number would consist of:

(*i*) a unique initial digit (e.g. 0) which identifies that the call is long distance. This serves as an access digit to route the call from the calling local area to the home GSC,

(*ii*) area code digits which identify the distant GSC area to which the call is to be routed,

(*iii*) the local number.

The home GSC receives the called-area code and local-number digits, routes the call over the long-distance network to the distant GSC of the required area, transmits forward the called local number to the distant GSC, and determines the charge.

On incoming long-distance traffic, the home GSC receives the called local-number digits from the distant GSC and automatically sets up the incoming call in the local area.

(*to be continued*)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