SUPPLEMENT

TO THE POST OFFICE ELECTRICAL ENGINEERS' JOURNAL

Vol. 74 Part 1 April 1981

ISSN 0309-2720

TEC & CGLI: GUIDANCE FOR STUDENTS

Contents

CITY AND GUILDS OF LONDON INSTITUTE

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.

MATHEMATICS C 1980

Students were expected to answer any 6 questions

Q1 (a) In a line test, the loop resistance R ohms and the distance x kilometres to a fault are related by the equations

$$R - 8x = 146, \text{ and}$$

$$8(21 - x) + \frac{8Rx}{R + 8x} = 136.$$

Simplify the second equation, and from the two equations derive a quadratic equation in x. Hence evaluate x and R.

(b) Oscillations occur in a circuit provided the roots of the equation $Lz^2 + Rz + 1/C = 0$ are complex. Give the condition for this involving L, R and C.

Display these roots in the complex plane, when $L = 0{\cdot}8 \times 10^{-3}$, $C = 2{\cdot}5 \times 10^{-6}$ and $R = 12{\cdot}8$.

A1 (*a*)

$$R - 8x = 146. \qquad \ldots\ldots (1)$$

$$8(21 - x) + \frac{8Rx}{R + 8x} = 136.$$

$$\therefore \quad 21 - x + \frac{Rx}{R + 8x} = 17. \qquad \ldots\ldots (2)$$

From equation (2),

$$(21 - x)(R + 8x) + Rx = 17(R + 8x).$$

$$\therefore \quad 21R + 168x - Rx - 8x^2 + Rx = 17R + 136x.$$

$$\therefore \quad 8x^2 - 32x - 4R = 0, \text{ or}$$

$$2x^2 - 8x - R = 0. \qquad \ldots\ldots (3)$$

But, from equation (1), $R = 8x + 146$.

Substituting for R in equation (3),

$$2x^2 - 8x - 8x - 146 = 0, \text{ or}$$

$$x^2 - 8x - 73 = 0.$$

$$\therefore \quad x = \frac{8 \pm \sqrt{(64 + 292)}}{2},$$

$$= \frac{8 \pm \sqrt{356}}{2},$$

$$= \frac{8 \pm 18{\cdot}868}{2},$$

$$= 13{\cdot}434 \text{ or } -5{\cdot}434.$$

The negative root is disregarded as inapplicable and, hence,

$x = 13{\cdot}434$. Substituting for x in equation (1),

$$R = 8 \times 13{\cdot}434 + 146 = 107{\cdot}472 + 146 = 253{\cdot}472.$$

Thus, the required equation is $\underline{x^2 - 8x - 73 = 0}$, whence $x = \underline{13{\cdot}434 \text{ km}}$ and $R = \underline{253{\cdot}47\ \Omega}$.

(*b*) $$Lz^2 + Rz + 1/C = 0.$$

In the general solution to a quadratic equation $ax^2 + bx + c$,

$$x = \frac{-b \pm \sqrt{(b^2 - 4ac)}}{2a}.$$

The condition for the roots to be complex is therefore that $b^2 < 4ac$. Hence, in the given equation, this condition is met when

$$\frac{4L}{C} > R^2 \text{ or } \underline{4L > CR^2}.$$

For the given conditions,

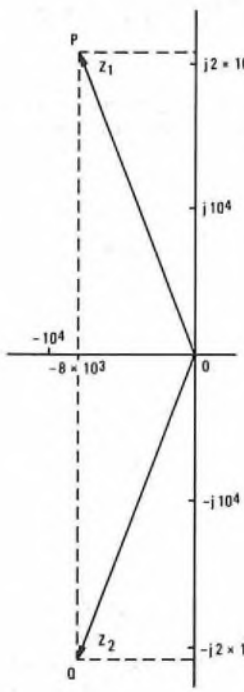
$$z = \frac{-12{\cdot}8 \pm \sqrt{\left(12{\cdot}8^2 - \dfrac{4 \times 0{\cdot}8 \times 10^{-3}}{2{\cdot}5 \times 10^{-6}}\right)}}{2 \times 0{\cdot}8 \times 10^{-3}}.$$

$$= \frac{-12{\cdot}8 \pm \sqrt{(163{\cdot}84 - 1280)}}{1{\cdot}6 \times 10^{-3}},$$

$$= \frac{-12{\cdot}8 \pm \mathrm{j}\sqrt{1116{\cdot}16}}{1{\cdot}6 \times 10^{-3}},$$

$$= \underline{-8000 \pm \mathrm{j}20\,881}.$$

These roots are shown in the sketch by the vectors OP and OQ.



Q2 (*a*) *Express* $(1 + x)^{-1\cdot4}$ *by the binomial series as far as the term in* x^3. *What limitations are there on the value of x for this series to be valid?*

(*b*) *The pressure p and the volume v of a quantity of gas vary according to the formula* $pv^{1\cdot4}$ = *constant. Use the series derived in part (a) to calculate to 2 significant figures the percentage change in the pressure when the volume is decreased by 4%.*

A2 (*a*) $$(1 + x)^{-1\cdot4} = 1 - 1{\cdot}4x + \frac{1{\cdot}4 \times 2{\cdot}4}{1 \times 2}x^2 - \frac{1{\cdot}4 \times 2{\cdot}4 \times 3{\cdot}4}{1 \times 2 \times 3}x^3 + \ldots$$

$$= \underline{1 - 1{\cdot}4x + 1{\cdot}68x^2 - 1{\cdot}904x^3 + \ldots}.$$

For the series to be valid, $\underline{|x| < 1}$.

(*b*) Let $pv^{1\cdot4} = k$.

Then, $p = kv^{-1\cdot4}$.

If the volume is decreased by 4%, the new volume, v_1, $= v - 0{\cdot}04v$, $= v(1 - 0{\cdot}04)$.

The new pressure, p_1, is given by

$$p_1 = kv_1 = k\{v(1 - 0{\cdot}04)\}^{-1\cdot4},$$
$$= kv^{-1\cdot4}(1 - 0{\cdot}04)^{-1\cdot4},$$
$$= p(1 - 0{\cdot}04)^{-1\cdot4}.$$

From part (*a*),

$$(1 - 0{\cdot}04)^{-1\cdot4} \approx 1 + 1{\cdot}4 \times 0{\cdot}04 + 1{\cdot}68 \times 0{\cdot}04^2 + 1{\cdot}904 \times 0{\cdot}04^3,$$
$$= 1 + 0{\cdot}056 + 0{\cdot}002688 + 0{\cdot}00012185,$$
$$= 1{\cdot}058810.$$

Hence, the pressure increases by $\underline{5{\cdot}9\%}$, correct to 2 significant figures.

Q3 (*a*) *Using logarithms to base 10 only, evaluate N when* $\log_{12} N = -0{\cdot}18$.

(*b*) *The surface of an exponential horn is obtained by revolving about the x-axis the curve* $y = ae^{x/k}$ *from* $x = 0$ *to* $x = l$.

(*i*) *If the diameter of the horn at* $x = 0$ *is d centimetres and at* $x = l$ *is D centimetres, express k in terms of l, D and d.*

(*ii*) *Sketch the section of the horn in the (x, y) plane, given* $a = 2$, $k = 15$ *and* $l = 33$.

A3 $\log_{12} N = -0{\cdot}18.$

$\therefore\quad N = 12^{-0\cdot18}.$

$\therefore\ \log_{10} N = -0{\cdot}18 \log_{10} 12,$

$= -0{\cdot}18 \times 1{\cdot}07918,$

$= -0{\cdot}19425 = \bar{1}{\cdot}80575.$

$\therefore\quad N = \underline{0{\cdot}63937}.$

(*b*) (*i*) $$y = ae^{x/k}.$$

When $x = 0$, $y = ae^0 = a$.

$y = a$ becomes the radius of the horn at $x = 0$ and hence $d/2 = a$.

When $x = l$, $y = ae^{l/k}$.

$$\therefore\quad D/2 = ae^{l/k}.$$

But, $$a = d/2.$$

$$\therefore\quad \frac{D/2}{d/2} = e^{l/k}, \text{ or}$$

$$\frac{D}{d} = e^{l/k}.$$

$$\therefore\quad l/k = \log_e \frac{D}{d}.$$

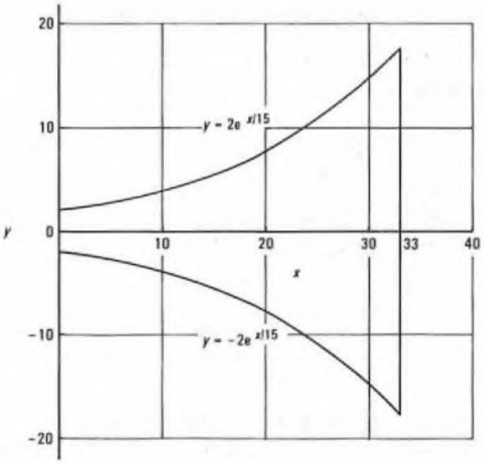
$$\therefore\quad \underline{k = \frac{l}{\log_e \dfrac{D}{d}}}.$$

(*ii*) $$y = 2e^{x/15}.$$

The curve may be sketched from the following table of values.

x	0	6	12	18	24	30	33
$x/15$	0	0·4	0·8	1·2	1·6	2·0	2·2
$e^{x/15}$	1	1·492	2·226	3·320	4·953	7·390	9·025
y	2	2·98	4·45	6·64	9·91	14·78	18·05

The section of the horn is shown in the sketch, the upper curve being the graph of $y = 2e^{x/15}$ and the lower that of $y = -2e^{x/15}$.



$$\therefore \quad \sin\theta\cos\alpha - \cos\theta\sin\alpha = -\frac{3}{5\cdot 3852},$$

where $\alpha = \tan^{-1}\frac{2}{5} = 22°$ to the nearest degree.

$$\therefore \quad \sin(\theta - \alpha) = -0\cdot 5571,$$

or $\theta - \alpha = -180° + 34°$ or $-34°$.

$$\therefore \quad \theta = -146° + 22° \text{ or } -34° + 22°,$$

$$= \underline{-12° \text{ or } -124° \text{ to the nearest degree.}}$$

Q5 *(a) Draw the polar curve $r = 4(1 + \cos\theta)$ from $\theta = -\pi$ to $\theta = +\pi$. Mark the points A, B, C on the curve at which $\theta = \frac{\pi}{3}, \frac{\pi}{6}$ and $\frac{3\pi}{4}$ respectively.*

(b) Choosing the line $\theta = 0$ as the x-axis and the line $\theta = \frac{\pi}{2}$ as the y-axis, deduce the cartesian equation of the curve in part (a). Name the type of curve.

Q4 *(a) If $\sin\alpha = \frac{20}{29}$ and α is obtuse, express $\tan 2\alpha$ as a vulgar fraction.*

(b) Prove that $\cos\left(\theta + \frac{2\pi}{3}\right) + \cos\theta + \cos\left(\theta - \frac{2\pi}{3}\right) = 0.$

(c) Solve the trigonometrical equation $2\cos\theta = 5\sin\theta + 3$ giving θ between $-180°$ and $+180°$ to the nearest degree.

A4 *(a)*

$$\sin\alpha = 20/29.$$

$$\therefore \quad \cos\alpha = \frac{\sqrt{(29^2 - 20^2)}}{29},$$

$$= \frac{\sqrt{(841 - 400)}}{29},$$

$$= \frac{\sqrt{441}}{29} = \frac{21}{29}.$$

Now,
$$\tan 2\alpha = \frac{\sin 2\alpha}{\cos 2\alpha},$$

$$= \frac{2\sin\alpha\cos\alpha}{\cos^2\alpha - \sin^2\alpha},$$

$$= \frac{2 \times \frac{20}{29} \times \frac{21}{29}}{\frac{21^2}{29^2} - \frac{20^2}{29^2}},$$

$$= \frac{40 \times 21}{41 \times 1} = \underline{\frac{840}{41}}.$$

(b)
$$\cos\left(\theta + \frac{2\pi}{3}\right) + \cos\theta + \cos\left(\theta - \frac{2\pi}{3}\right) =$$

$$\cos\theta \times \cos\frac{2\pi}{3} - \sin\theta \times \sin\frac{2\pi}{3} + \cos\theta +$$

$$\cos\theta \times \cos\frac{2\pi}{3} + \sin\theta \times \sin\frac{2\pi}{3},$$

$$= 2\cos\theta\cos\frac{2\pi}{3} + \cos\theta,$$

$$= 2\cos\theta \times (-\tfrac{1}{2}) + \cos\theta = 0. \qquad \text{QED}$$

(c)
$$2\cos\theta = 5\sin\theta + 3,$$

or
$$5\sin\theta - 2\cos\theta = -3.$$

$$\therefore \quad \sqrt{(5^2 + 2^2)}\left(\sin\theta \times \frac{5}{\sqrt{29}} - \cos\theta \times \frac{2}{\sqrt{29}}\right) = -3,$$

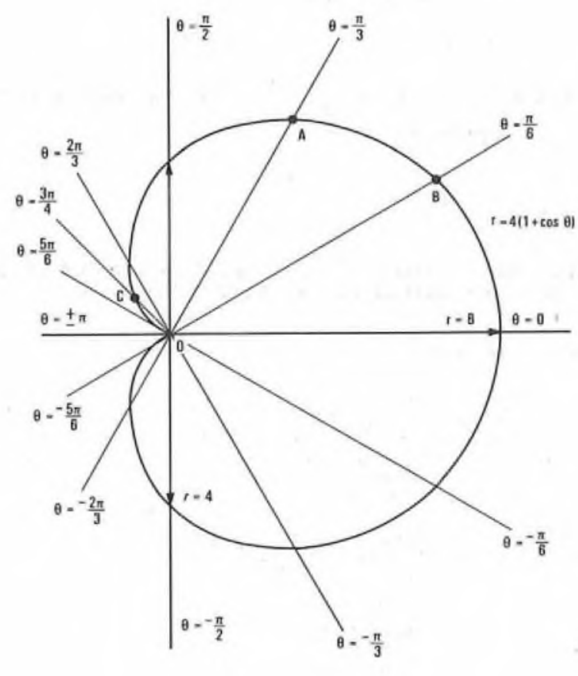
$$= 5\cdot 3852\left(\sin\theta \times \frac{5}{5\cdot 3852} - \cos\theta \times \frac{2}{5\cdot 3852}\right).$$

A5 *(a)* The curve may be drawn from the set of values in the following table.

θ	$-\pi$	$-\frac{5\pi}{6}$	$-\frac{2\pi}{3}$	$-\frac{\pi}{2}$	$-\frac{\pi}{3}$	$-\frac{\pi}{6}$	0
$\cos\theta$	-1	$-0\cdot 866$	$-0\cdot 5$	0	$0\cdot 5$	$0\cdot 866$	1
$1 + \cos\theta$	0	$0\cdot 134$	$0\cdot 5$	$1\cdot 0$	$1\cdot 5$	$1\cdot 866$	2
r	0	$0\cdot 536$	$2\cdot 0$	$4\cdot 0$	$6\cdot 0$	$7\cdot 464$	8

Since $\cos\theta = \cos(-\theta)$, the values of r from $\theta = \pi$ to $\theta = 0$ will be the same, at the same intervals, as those in the table.

The polar curve is shown in sketch *(a)*, in which the points A, B and C are shown.



(a)

(*b*) If $P(r, \theta)$ is any point on the curve of $r = 4(1 + \cos\theta)$, as shown in sketch (*b*), and PN is the perpendicular from P to the x-axis, then, from triangle PON,

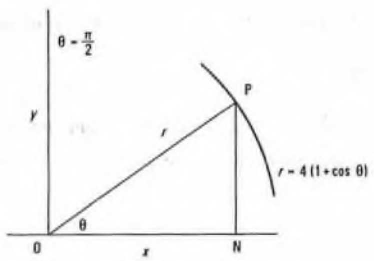
$$x = r\cos\theta \text{ and } y = r\sin\theta.$$

Also, $\quad r^2 = x^2 + y^2.$

But, $\quad r = 4 + 4\cos\theta.$

$$\therefore \quad \sqrt{(x^2+y^2)} = 4 + 4\frac{x}{r},$$

$$= 4 + \frac{4x}{\sqrt{(x^2+y^2)}}. \qquad \ldots\ldots (1)$$



(*b*)

Multiplying equation (1) by $\sqrt{(x^2 + y^2)}$ gives

$$x^2 + y^2 = 4\sqrt{(x^2+y^2)} + 4x,$$

$$= \underline{4\{x + \sqrt{(x^2+y^2)}\}}.$$

This is the cartesian equation of the polar curve, which is known as a *Cardioid.*

Q6 (a) From first principles derive a formula for differentiating the quotient of two functions.

(b) Obtain, and simplify an expression for $\frac{dy}{dx}$ *when*

(i) $y = \cot x$, *and*

(ii) $y = \dfrac{1 + 2x}{(x + 2)^2}$.

(c) Sketch the graph of $y = \dfrac{1 + 2x}{(x+2)^2}$ *marking any maxima, minima, points of inflexion and asymptotes.*

A6 (*a*) Let $y = \dfrac{u}{v}$.

Suppose that x increases by a small amount, δx, and let δu, δv and δy be the corresponding changes in u, v and y respectively.

Then, $\quad y + \delta y = \dfrac{u + \delta u}{v + \delta v},$

or, $\quad \delta y = \dfrac{u + \delta u}{v + \delta v} - y,$

$$= \frac{u + \delta u}{v + \delta v} - \frac{u}{v},$$

$$= \frac{v(u + \delta u) - u(v + \delta v)}{v(v + \delta v)},$$

$$= \frac{v\delta u - u\delta v}{v(v + \delta v)}.$$

$$\therefore \frac{\delta y}{\delta x} = \frac{v\dfrac{\delta u}{\delta x} - u\dfrac{\delta v}{\delta x}}{v(v + \delta v)}.$$

As $\delta x \to 0$, $\dfrac{\delta u}{\delta x} \to \dfrac{du}{dx}$, $\dfrac{\delta v}{\delta x} \to \dfrac{dv}{dx}$, $\dfrac{\delta y}{\delta x} \to \dfrac{dy}{dx}$, and $\delta v \to 0$.

Hence, in the limit, when δx is zero,

$$\frac{dy}{dx} = \frac{v\dfrac{du}{dx} - u\dfrac{dv}{dx}}{v^2},$$

or, $$\underline{\frac{d}{dx}\left(\frac{u}{v}\right) = \frac{1}{v^2}\left(v\frac{du}{dx} - u\frac{dv}{dx}\right)}.$$

(*b*) (*i*) $\quad y = \cot x = \dfrac{\cos x}{\sin x}.$

$$\therefore \quad \frac{dy}{dx} = \frac{\sin x\, d(\cos x) - \cos x\, d(\sin x)}{\sin^2 x},$$

$$= \frac{-\sin^2 x - \cos^2 x}{\sin^2 x},$$

$$= -\frac{1}{\sin^2 x} = \underline{-\operatorname{cosec}^2 x}.$$

(*ii*) $$\frac{dy}{dx} = \frac{(x+2)^2\, d(1 + 2x) - (1 + 2x)\, d(x + 2)^2}{(x+2)^4},$$

$$= \frac{(x+2)^2 \times 2 - (1 + 2x) \times 2 \times (x + 2)}{(x + 2)^4},$$

$$= \frac{2x + 4 - 2 - 4x}{(x+2)^3},$$

$$= \frac{-2x + 2}{(x+2)^3},$$

$$= \underline{\frac{-2(x-1)}{(x+2)^3}}.$$

(*c*) Maxima or minima occur where $\dfrac{dy}{dx} = 0$; that is, when

$$-2(x - 1) = 0 \text{ or } x = 1.$$

$$\frac{d^2y}{dx^2} = \frac{(x+2)^3\, d(-2x + 2) - (-2x + 2)\, d(x+2)^3}{(x+2)^6},$$

$$= \frac{-2(x+2)^3 + (2x - 2) \times 3 \times (x+2)^2}{(x+2)^6},$$

$$= \frac{-2(x+2) + 3(2x-2)}{(x+2)^4},$$

$$= \frac{4x - 10}{(x+2)^4}.$$

When $x = 1$, $\dfrac{d^2y}{dx^2}$ is negative and, hence, a maximum value occurs when $x = 1$.

Points of inflexion occur where $\dfrac{d^2y}{dx^2} = 0$; that is, when $4x - 10 = 0$ or $x = 2\frac{1}{2}$.

Just below $x = 2\frac{1}{2}$, at, say, $x = 2\frac{1}{4}$,

$$\frac{d^2y}{dx^2} = \frac{9 - 10}{(4\frac{1}{4})^4}\text{; that is, it is negative.}$$

Just above $x = 2\frac{1}{2}$, at, say, $x = 2\frac{3}{4}$,

$$\frac{d^2y}{dx^2} = \frac{11 - 10}{(4\frac{3}{4})^4}\text{; that is, it is positive.}$$

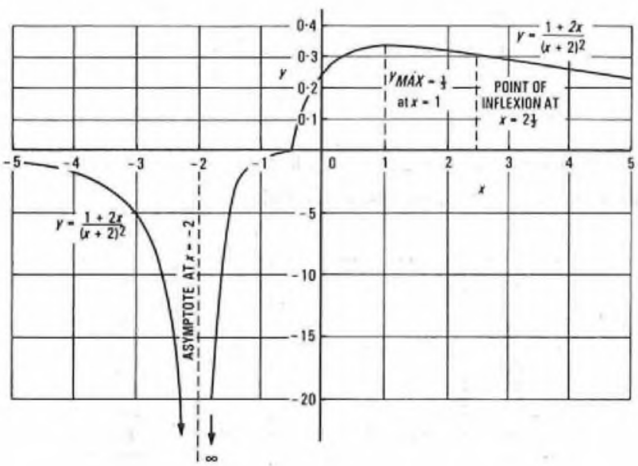
Since the sign of $\dfrac{d^2y}{dx^2}$ changes from negative to positive, a point of inflexion must occur at $x = 2\frac{1}{2}$.

The curve is plotted from the following table of values.

The graph is shown in the sketch, with the salient points marked. In addition, the x-axis becomes asymptotic to the curve for both very large negative and positive values of x.

x	-5	-4	-3	$-2\frac{1}{2}$	-2	$-1\frac{1}{2}$	-1	$-\frac{1}{2}$
$1+2x$	-9	-7	-5	-4	-3	-2	-1	0
$x+2$	-3	-2	-1	$-\frac{1}{2}$	0	$\frac{1}{2}$	1	$1\frac{1}{2}$
$(x+2)^2$	9	4	1	$\frac{1}{4}$	0	$\frac{1}{4}$	1	$2\frac{1}{4}$
y	-1	$-1\frac{3}{4}$	-5	-16	$-\infty$	-8	-1	0

x	0	$\frac{1}{2}$	1	2	3	4	5
$1+2x$	1	2	3	5	7	9	11
$x+2$	2	$2\frac{1}{2}$	3	4	5	6	7
$(x+2)^2$	4	$6\frac{1}{4}$	9	16	25	36	49
y	$\frac{1}{4}$	$0\cdot32$	$0\cdot3\dot{3}$	$0\cdot3125$	$0\cdot28$	$\frac{1}{4}$	$0\cdot224$



When $x \gg 1$, $y \approx \dfrac{2x}{x^2}$,

$$= \frac{2}{x}.$$

Hence, as $x \to \infty$, $y \to 0$.

Q7 (*a*) *Differentiate with respect to t*

(*i*) $y = 4e^{-3t}\sin 2t$,

(*ii*) $y = \log_e(1 + t^2)$, *and*

(*iii*) $y = \sqrt{(1 + t^3)}$.

Simplify the results where appropriate.

(*b*) *If* $y = x^{-\frac{1}{2}}\cos x$ *prove that* $x^2\dfrac{d^2y}{dx^2} + x\dfrac{dy}{dx} + (x^2 - \frac{1}{4})y = 0$.

A7 (*a*) (*i*) $\dfrac{dy}{dx} = 4e^{-3t} \times \cos 2t \times 2 - 3 \times 4e^{-3t}\sin 2t,$

$$= \underline{4e^{-3t}(2\cos 2t - 3\sin 2t)}.$$

(*ii*) $\dfrac{dy}{dx} = \underline{\dfrac{2t}{1 + t^2}}.$

(*iii*) $\dfrac{dy}{dx} = \frac{1}{2}(1 + t^3)^{-\frac{1}{2}} \times 3t^2,$

$$= \underline{\frac{3t^2}{2\sqrt{(1 + t^3)}}}.$$

(*b*) $\dfrac{dy}{dx} = x^{-\frac{1}{2}} \times (-\sin x) - \frac{1}{2}x^{-3/2} \times \cos x,$

$$= -x^{-\frac{1}{2}}(\sin x + \tfrac{1}{2}x^{-1} \times \cos x).$$

$$\frac{d^2y}{dx^2} = -x^{-\frac{1}{2}}(\cos x - \tfrac{1}{2}x^{-2} \times \cos x - \tfrac{1}{2}x^{-1} \times \sin x) +$$
$$\tfrac{1}{2}x^{-3/2}(\sin x + \tfrac{1}{2}x^{-1} \times \cos x),$$
$$= \cos x(-x^{-\frac{1}{2}} + \tfrac{1}{2}x^{-5/2} + \tfrac{1}{4}x^{-5/2}) +$$
$$\sin x\,(\tfrac{1}{2}x^{-\frac{1}{2}} + \tfrac{1}{2}x^{-3/2}),$$
$$= \cos x\,(-x^{-3/2} + \tfrac{3}{4}x^{-5/2}) + \sin x \times x^{-3/2}.$$

Hence, $x^2\dfrac{d^2y}{dx^2} + x\dfrac{dy}{dx} + (x^2 - \frac{1}{4})y =$

$$\cos x\,(-x^{3/2} + \tfrac{3}{4}x^{-\frac{1}{2}}) + \sin x \times x^{\frac{1}{2}} -$$
$$x^{\frac{1}{2}}(\sin x + \tfrac{1}{2}x^{-1} \times \cos x) + (x^2 - \tfrac{1}{4}) \times x^{-\frac{1}{2}} \times \cos x.$$
$$= \cos x\,(-x^{3/2} + \tfrac{3}{4}x^{-\frac{1}{2}} - \tfrac{1}{2}x^{-\frac{1}{2}} + x^{3/2} - \tfrac{1}{4}x^{-\frac{1}{2}}) +$$
$$\sin x\,(x^{\frac{1}{2}} - x^{\frac{1}{2}}),$$
$$= 0.$$

QED

Q8 (*a*) *Evaluate* $\displaystyle\int_1^3 \frac{5\,dx}{2x + 3}$.

(*i*) *by formal integration, and*

(*ii*) *using Simpson's Rule with 5 ordinates.*

(*b*) *Express the function* $y = 8\cos 4x \sin 2x$ *as the sum or difference of 2 sinusoids, and hence calculate the mean value of y from* $x = 0$ *to* $x = \pi/3$.

A8 (*a*) (*i*) $\displaystyle\int_1^3 \frac{5\,dx}{2x + 3} = \frac{5}{2}\Big[\log_e(2x + 3)\Big]_1^3,$

$$= \frac{5}{2}(\log_e 9 - \log_e 5),$$

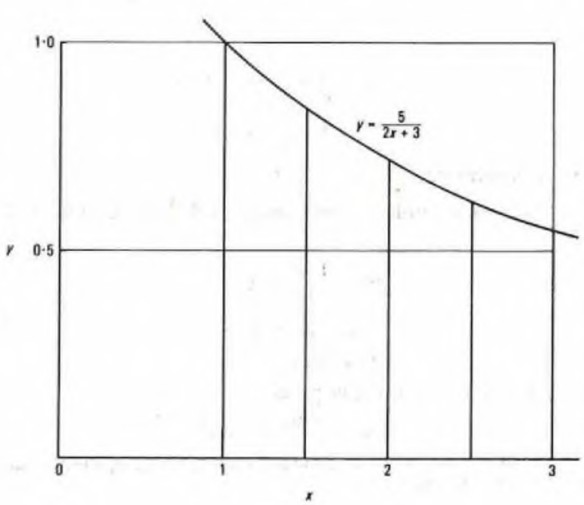
$$= \frac{5}{2}(2\cdot19722 - 1\cdot60944),$$

$$= \frac{5}{2} \times 0\cdot58778,$$

$$= \underline{1\cdot46945}.$$

(*ii*) Between the limits of 1 and 3, five ordinates are obtained by taking equal steps of 0·5, derived in the following table.

x	1	1·5	2	2·5	3
$2x + 3$	5	6	7	8	9
$\dfrac{1}{2x + 3}$	0·2	0·1667	0·1429	0·125	0·1111
$y = \dfrac{5}{2x + 3}$	1·0	0·8335	0·7145	0·625	0·5555



By Simpson's rule,

Area under the curve $= \dfrac{\text{width of strip}}{3}$ [sum of first and last ordinates $+ 2 \times$ sum of other odd ordinates $+ 4 \times$ sum of even ordinates],

$$= \frac{0 \cdot 5}{3}\,[1 \cdot 0 + 0 \cdot 5555 + 2 \times 0 \cdot 7145 + 4\,(0 \cdot 8335 + 0 \cdot 625)],$$

$$= \frac{0 \cdot 5}{3} \times 8 \cdot 8185.$$

$$= \underline{1 \cdot 46975.}$$

(*b*) $y = 8 \cos 4x \sin 2x,$

$= 4 \times 2 \cos 4x \sin 2x.$

From the product-to-sum formula,

$2 \cos A \sin B = \sin (A + B) - \sin (A - B).$

$\therefore \quad y = 4\,\{\sin (4x + 2x) - \sin (4x - 2x)\},$

$= \underline{4\,(\sin 6x - \sin 2x).}$

$$\text{Mean value} = \frac{\int_0^{\pi/3} 4\,(\sin 6x - \sin 2x)\,dx}{\pi/3},$$

$$= \frac{3}{\pi} \times 4\left[-\frac{\cos 6x}{6} - \frac{\cos 2x}{2}\right]_0^{\pi/3},$$

$$= \frac{12}{\pi}\left(-\frac{1}{6}\cos 2\pi - \frac{1}{2}\cos\frac{2\pi}{3} + \frac{1}{6}\cos 0 + \frac{1}{2}\cos 0\right),$$

$$= \frac{12}{\pi}\left(-\tfrac{1}{6} - \tfrac{1}{2} \times -\tfrac{1}{2} + \tfrac{1}{6} + \tfrac{1}{2}\right),$$

$$= \frac{12}{\pi} \times \tfrac{3}{4},$$

$$= \frac{9}{\pi} = \underline{2 \cdot 8648.}$$

Q9 (*a*) *Find the equation of the curve through the point (0, 10) such that*

$$\frac{dy}{dx} = 12e^{-3x} + 5.$$

Sketch the curve, showing both positive and negative values of x.

(*b*) *Calculate the volume of revolution generated when the curve* $y = 2x + \dfrac{12}{x}$ *from x = 1 to x = 4 is rotated through 360° about the x-axis.*

A9

$$y = \int (12e^{-3x} + 5)\,dx,$$

$$= \frac{12e^{-3x}}{-3} + 5x + C,$$

where C is a constant.

Substituting the coordinates of point (0, 10) in the expression for y gives

$$10 = \frac{12e^0}{-3} + C,$$

$$= -4 + C.$$

$$\therefore \quad C = 10 + 4 = 14.$$

Hence, the equation of the curve is

$$\underline{y = -4e^{-3x} + 5x + 14.}$$

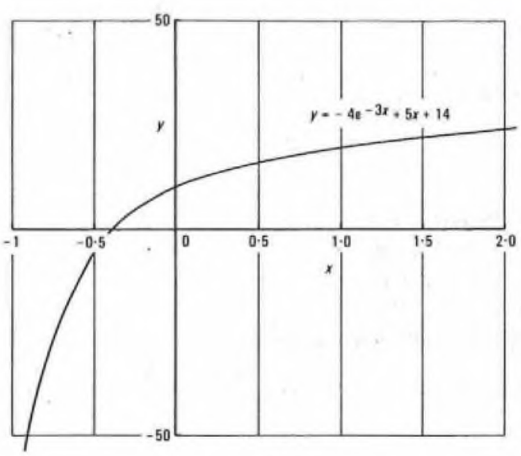
The graph of the function is shown in sketch (*a*), plotted from the following table of values.

x	-1	$-0 \cdot 5$	0	$0 \cdot 5$
$-3x$	3	$1 \cdot 5$	0	$-1 \cdot 5$
e^{-3x}	$20 \cdot 09$	$4 \cdot 48$	1	$0 \cdot 223$
$-4e^{-3x}$	$-80 \cdot 36$	$-17 \cdot 92$	-4	$-0 \cdot 89$
$5x$	-5	$-2 \cdot 5$	0	$2 \cdot 5$
14	14	14	14	14
y	$-71 \cdot 36$	$-6 \cdot 42$	10	$15 \cdot 61$

x	1	$1 \cdot 5$	2
$-3x$	-3	$-4 \cdot 5$	-6
e^{-3x}	$0 \cdot 050$	$0 \cdot 011$	$0 \cdot 0025$
$-4e^{-3x}$	$-0 \cdot 20$	$-0 \cdot 044$	$-0 \cdot 010$
$5x$	5	$7 \cdot 5$	10
14	14	14	14
y	$8 \cdot 18$	$21 \cdot 46$	$23 \cdot 99$

(*b*) The curve of $y = 2x + \dfrac{12}{x}$, plotted from $x = 1$ to $x = 4$, is shown in sketch (*b*), plotted from the following table.

x	1	2	3	4
$2x$	2	4	6	8
$\dfrac{12}{x}$	12	6	4	3
y	14	10	10	11



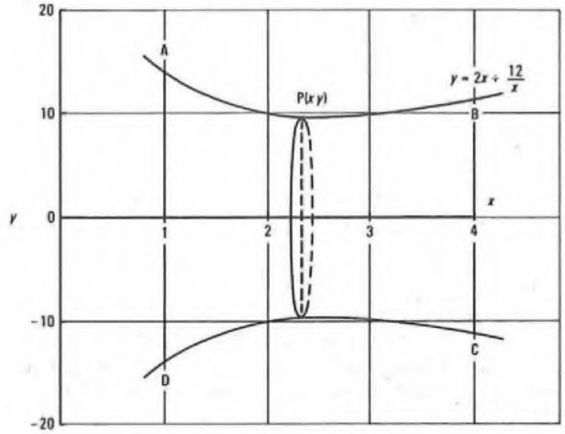
(*a*)

When the curve is rotated through 360° about the x-axis, it generates a solid depicted by ABCD in sketch (*b*).

At any point, P(x, y), on the curve a circular area πy^2 will be generated and the total volume from $x = 1$ to $x = 4$ is given by

$$\int_1^4 \pi y^2\,dx,$$

$$= \int_1^4 \pi\left(2x + \frac{12}{x}\right)^2 dx,$$



(b)

$$= 4\pi \int_1^4 (x^2 + 12 + 36x^{-2})\, dx,$$

$$= 4\pi\left[\frac{x^3}{3} + 12x + \frac{36x^{-1}}{-1}\right]_1^4,$$

$$= 4\pi\left(\frac{64}{3} + 48 - \frac{36}{4} - \frac{1}{3} - 12 + 36\right),$$

$$= 4\pi(21 + 72 - 9),$$

$$= \underline{336\pi = 1055{\cdot}58.}$$

Q10 (*a*) *Express* $Z = \dfrac{3 + j2}{7 - j} + \dfrac{7}{j4}$ *in the form* $r\angle\theta$.

(*b*) *A resistor R ohms and an inductor L henrys connected in series provide an impedance* Z_1 *at a frequency of* ω *radians/second.*

Connected in parallel they provide an impedance Z_2 *at the same frequency.*

(*i*) *Show that the phasors* Z_1 *and* $\dfrac{1}{Z_2}$ *are at right angles to one another.*

(*ii*) *Prove that the product* Z_1Z_2 *is directly proportional to the frequency.*

A10 (*a*)

$$Z = \frac{3 + j2}{7 - j} + \frac{7}{j4},$$

$$= \frac{(3 + j2)(7 + j)}{(7 - j)(7 + j)} - \frac{j7}{4},$$

$$= \frac{21 + j14 + j3 - 2}{49 + 1} - \frac{j7}{4},$$

$$= \frac{19 + j17}{50} - \frac{j7}{4},$$

$$= \frac{38 + j34 - j175}{100},$$

$$= 0{\cdot}38 - j1{\cdot}41,$$

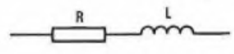
$$= \sqrt{\{(0{\cdot}38)^2 + (1{\cdot}41)^2\}} \angle \tan^{-1}\frac{-1{\cdot}41}{0{\cdot}38},$$

$$= \sqrt{(0{\cdot}1444 + 1{\cdot}9881)} \angle \tan^{-1} -3{\cdot}7105,$$

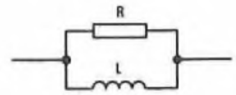
$$= \sqrt{2{\cdot}1325} \angle 360° - 74° 55',$$

$$\approx \underline{1{\cdot}46 \angle 285°.}$$

(*b*) (*i*) The series and parallel circuit are shown in sketches (*a*) and (*b*) respectively.



(*a*)



(*b*)

$$Z_1 = R + j\omega L,$$

$$= \sqrt{(R^2 + \omega^2L^2)} \angle \tan^{-1}\frac{\omega L}{R}.$$

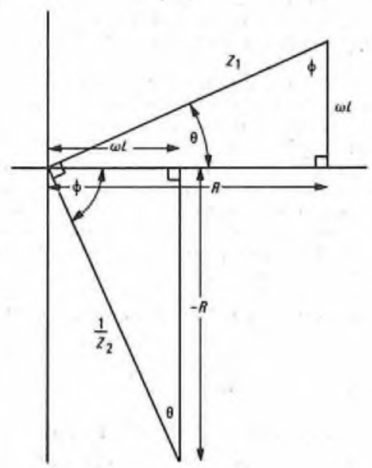
The phasor triangle is shown in sketch (*c*), such that $\tan\theta = \dfrac{\omega L}{R}$.

Also, $$\frac{1}{Z_2} = \frac{1}{R} + \frac{1}{j\omega L},$$

$$= \frac{1}{R} - \frac{j}{\omega L},$$

$$= \sqrt{\left\{\left(\frac{1}{R}\right)^2 + \left(\frac{1}{\omega L}\right)^2\right\}} \angle \tan^{-1}\left(-\frac{R}{\omega L}\right).$$

The phasor triangle of $\dfrac{1}{Z_2}$ is shown also in sketch (*c*), such that $\tan\phi = \dfrac{-R}{\omega L}$.



(*c*)

It is clear from sketch (*c*) that the two triangles are congruent and hence the angle between the phasors, $\theta + \phi = 90°$.

(*ii*) $$\frac{1}{Z_2} = \frac{1}{R} + \frac{1}{j\omega L},$$

$$= \frac{R + j\omega L}{j\omega LR}.$$

$$\therefore \quad Z_2 = \frac{j\omega LR}{R + j\omega L}.$$

Hence, $$Z_1Z_2 = (R + j\omega L) \times \frac{j\omega LR}{R + j\omega L},$$

$$= j\omega LR.$$

$$\therefore \quad Z_1Z_2 \propto \omega \propto f.$$

QED

Students were expected to answer any 6 questions

Q1 (*a*) *For communication at a frequency of 5 GHz, explain why*

(*i*) *power of only a few watts will suffice for a short range point-to-point link,*
(*ii*) *a bandwidth of at least 50 MHz can be made available,*
(*iii*) *internal receiver noise is usually of more importance than external noise, and*
(*iv*) *tropospheric conditions may affect reception.*

(*b*) *A signal-to-noise ratio of 15 dB is required at the output of a microwave receiver which has a bandwidth of 18 MHz and a noise factor of 8 dB. Determine whether this can be obtained if the input signal power to the receiver is 3×10^{-11} W (kT may be assumed to be 4×10^{-21} J).*

A1 (*a*) (*i*) At a frequency of 5 GHz an input power of a few watts will normally suffice for a short-range, point-to-point, line-of-sight radio link because practical aerials, which may have an area of several square metres, have high gain and are highly directive. This, when combined with the high sensitivity of typical receivers at 5 GHz, more than overcomes the typical path loss encountered on a short link. This loss would amount to about 120 dB for a 5 km path length.

(*ii*) A bandwidth of 50 MHz is readily available at 5 GHz, both in terms of the circuit component Q-factors and the percentage bandwidth that such a figure represents of the centre frequency; that is, 50 MHz is only 1% of the 5 GHz centre frequency.

(*iii*) Modern microwave radio-link receivers at 5 GHz typically may have a noise figure of about 8–10 dB; that is, an equivalent noise temperature of about 2000–3000°K. This noise temperature far exceeds that due to external man-made and natural sources of excess noise delivered to the receiver and even further discrimination against external noise sources is provided by the high directivity of the aerials employed.

(*iv*) At microwave frequencies the propagation velocity of the radio waves is affected by the varying water-vapour content of the intervening atmosphere. The radio waves undergo refraction between transmitter and receiver. Under normal weather conditions the aerials are aligned to give optimum results, but should certain sharp changes in the water-vapour content in the line-of-sight path take place, various abnormal radio propagation conditions may occur. These may include a complete loss of the signal over the link; reception of the signal via various paths, which then subsequently constructively and destructively interfere with each other, producing rapid fading at the receiver; or even super-enhancement of the signal. Precipitation in the line-of-sight path produces a slight attenuation of the radio signal due to absorption and scattering at 5 GHz, but its effect only becomes severe at frequencies greater than about 10 GHz.

(*b*) By definition,

$$\text{Noise Factor, } F, = \frac{\text{Input signal-to-noise ratio}}{\text{Output signal-to-noise ratio}} = \frac{\text{SNR}_{\text{in}}}{\text{SNR}_{\text{out}}}.$$

For a signal to be detected, the input signal-to-noise ratio must therefore exceed (15 + 8) dB = 23 dB.

If the signal power, $P_{\text{in}} = 3 \times 10^{-11}$ W and the receiver has a bandwidth, B, of 18 MHz, the applied input signal-to-noise ratio will be

$$\frac{P_{\text{in}}}{kTB}, = \frac{3 \times 10^{-11}}{4 \times 10^{-21} \times 18 \times 10^{6}} = 416 \cdot 67 = 26 \cdot 2 \text{ dB}.$$

The above signal is therefore detectable in the microwave receiver.

Q2 (*a*) *State why the velocity of propagation of electromagnetic waves on a solid coaxial cable is less than the free-space velocity.*

(*b*) *A 1·5 m length of coaxial cable having a characteristic impedance of 75 Ω is terminated in a 15 Ω resistor and energized at 35 MHz. The velocity of propagation in the cable is $2 \cdot 1 \times 10^{8}$ m/s. Determine*

(*i*) *the reflection coefficient of the load,*
(*ii*) *the standing-wave ratio,*
(*iii*) *the wavelength on the cable,*
(*iv*) *the input impedance of the cable, and*
(*v*) *the voltage at the cable input when it is connected to a 30 V 35 MHz source which has an internal impedance of 75 Ω.*

A2 (*a*) The velocity of propagation, v, of electromagnetic waves on a solid dielectric filled coaxial cable is defined as

$$v = \frac{c}{\sqrt{\epsilon_r}},$$

where c = velocity of light, and
ϵ_r = relative permittivity of the dielectric.

Typically, for solid polyethylene-filled cable, $\epsilon_r = 2 \cdot 3$, and therefore the velocity of propagation on such a coaxial cable is only 0·66 of the velocity of light.

(*b*) (*i*) A 1·5 m length of solid dielectric-filled coaxial cable of characteristic impedance 75 Ω can be assumed to be lossless. With a resistive load of 15 Ω terminating the transmission line, the voltage reflection coefficient, ρ, at the load will be

$$\rho = \frac{R_L - Z_0}{R_L + Z_0} = \frac{15 - 75}{15 + 75} = -\underline{\frac{2}{3}} = \frac{V_r}{V_i},$$

where V_r is the reflected voltage, V_i is the incident voltage, R_L is the terminating resistance and Z_0 is the characteristic impedance of the cable.

(*ii*) The standing wave ratio, S,

$$= \frac{1 + |V_r/V_i|}{1 - |V_r/V_i|} = \underline{5.}$$

(*iii*) The wavelength of the signal, λ, on the solid cable will be shorter than on an air-filled cable because of the slower velocity of propagation.

$$\lambda = \frac{v}{f} = \frac{2 \cdot 1 \times 10^{8}}{35 \times 10^{6}} = \underline{6 \text{ m.}}$$

(*iv*) Since the wavelength on the cable is 6 m and the cable is 1·5 m long, the electrical length of the cable is $\lambda/4$. When the length of a lossless transmission line is exactly an odd multiple of quarter wavelengths the input impedance, Z, is given by

$$Z = \frac{Z_0^2}{R_L} = \underline{375\ \Omega.}$$

(*v*) From part (*iv*), the cable has an input impedance of 375 Ω at 35 MHz. When a voltage source of 30 V, with an internal impedance of 75 Ω, is connected to the input, the actual voltage at this point is given by

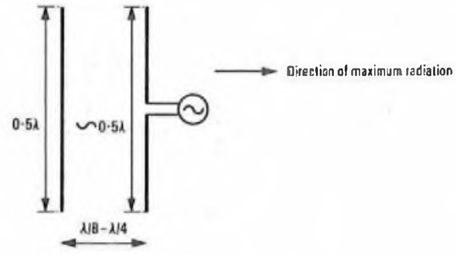
$$V_{\text{ip}} = \frac{30 \times 375}{75 + 375} = \underline{25 \text{ V.}}$$

Q3 (*a*) *Explain, with the aid of a diagram, why the gain and directivity of a $\frac{\lambda}{2}$ dipole are increased by the use of a rod reflector. Why are the length and spacing of the reflector important?*

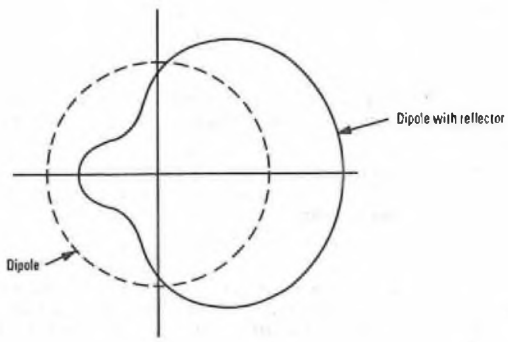
(*b*) *Describe how a $\frac{\lambda}{2}$ dipole with a rod reflector may be mounted at the mouth of a rectangular waveguide to provide a satisfactory radiator.*

A3 (*a*) As shown in sketch (*a*), an undriven rod reflector, $\lambda/2$ long, may be placed behind and parallel to a driven $\lambda/2$ dipole. A current will be induced on the reflector by the fields of the driven element. Such an element is referred to as a *parasitic* element. To obtain an increase in directivity, and hence forward gain, from this arrangement the impedance of the parasitic element must be adjusted until it has a positive reactance. This condition is found to be about one-eighth to one-quarter of a wavelength behind the driven element. It is essential that the reflector be electrically one-half wavelength long to obtain the necessary positive reactance. Since the reflector has a purely reactive impedance it has little or no loss and re-radiates the energy falling on it to give increased directivity in the direction of the driven element. The increase in directivity or forward gain amounts to about 3 dB.

Sketch (*b*) shows the original dipole horizontal polar radiation pattern, together with that obtained with a dipole and reflector combination.

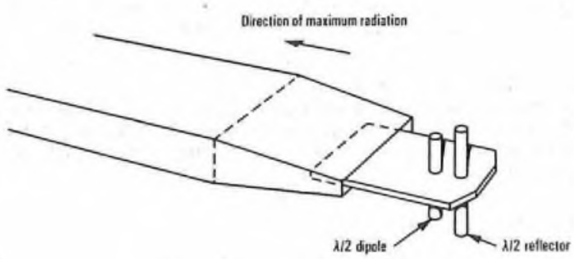


(*a*)



(*b*)

(*b*) In the feed radiator shown in sketch (*c*), most of the power is directed backwards, thus making this item a useful waveguide and dipole rear feed for parabolic-reflector type aerials.



(*c*)

Q4 *Fig. 1 represents the plan view and elevation of a cavity constructed from a 10 cm length of 6 × 3 cm waveguide short-circuited at both ends. A microwave source, connected by coaxial cable to the probe P, strongly excites resonance.*

(*a*) (*i*) *Name the mode of resonance and sketch the E-field and H-field patterns.*

(*ii*) *Calculate the resonant frequency.*

(*iii*) *State with reasons whether, with this probe position, strong excitation of the* TE_{102} *mode will be possible.*

(*b*) *Explain the effects, if any, on the cavity resonant frequency and Q-factor of increasing*

(*i*) *the a dimension, and*

(*ii*) *the b dimension.*

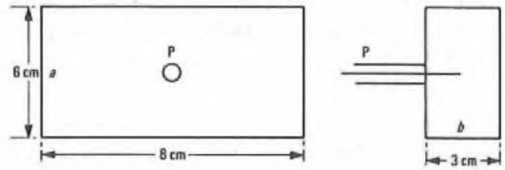
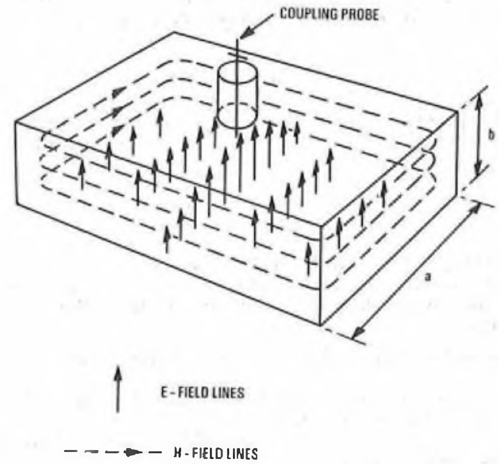


Fig. 1

A4 (*a*) (*i*) The mode of resonance is described as the TE_{101} mode and the E and H-field patterns are shown in the sketch.



(*ii*) The cavity is considered as a $\lambda_g/2$ long waveguide when operated in the above, lowest frequency, mode of resonance. The cut-off wavelength of the waveguide is given by

$$\lambda_c = 2 \times \text{dimension } a, = 2 \times 6 = 12 \text{ cm, and}$$
$$\lambda_g = 2 \times \text{length of cavity}, = 2 \times 8 = 16 \text{ cm.}$$
$$\frac{1}{\lambda^2} = \frac{1}{\lambda_c^2} + \frac{1}{\lambda_g^2} = \frac{1}{12 \times 12} + \frac{1}{16 \times 16} = \frac{1}{9{\cdot}6^2}.$$

Therefore, the lowest resonant frequency of the cavity

$$= \frac{3 \times 10^{10}}{9{\cdot}6} = 3{\cdot}125 \text{ GHz.}$$

(*iii*) It would not be possible to couple any power into the TE_{102} mode. This mode requires two concentrations of electric field intensity along the length of the resonator and this leaves a point of zero field intensity in the centre of the resonator, a point at which no power could be coupled by an electric probe.

(*b*) (*i*) Increasing the *a* dimension of the cavity would lower the frequency at which the cavity would resonate. The Q-factor of the cavity is proportional to the ratio of the cavity volume to the surface area. An increase in width of the cavity produces a greater increase in internal volume than the total wall surface area; therefore the Q-factor of the cavity will increase.

(*ii*) Increasing the *b* dimension of the cavity will not in general alter the resonant frequency of the cavity and, as in part (*i*), the volume-to-surface-area ratio will increase, thus raising the Q-factor of the cavity.

Q5 (*a*) *A frequency modulation (FM) system having a maximum deviation of 51 kHz is designed for inputs in the range 100–6800 Hz. A 4 V 2 kHz test signal produces a deviation of 16 kHz. Determine*

(*i*) *the modulation index in the test condition,*

(*ii*) *the deviation ratio of the system,*

(*iii*) *the deviation produced by a 6 V 2·5 kHz input, and*

(*iv*) *the 5 kHz test signal voltage required to produce maximum deviation.*

(*b*) *The output of a frequency modulator comprises a 1 MHz carrier with a pair of side frequencies separated from it by 5 kHz. This is applied to a frequency doubler. Describe the output from the frequency doubler.*

(*c*) *Explain how amplitude variations may be removed from a frequency-modulated signal.*

A5 (*a*) (*i*) If a 4 V 2 kHz test signal produces a frequency deviation of 16 kHz, the modulation index $= 16/2 = \underline{8}$.

(*ii*) The deviation ratio of the system is defined as

$$\text{deviation ratio} = \frac{\text{maximum deviation}}{\text{maximum modulating frequency}}.$$

In the system described, maximum deviation is 51 kHz and the maximum modulating frequency is 6800 Hz. Therefore the

$$\text{deviation ratio} = \frac{51}{6 \cdot 8} = \underline{7 \cdot 5}.$$

(*iii*) A 4 V 2 kHz test signal produces a deviation of 16 kHz; therefore a 6 V 2·5 kHz signal will produce a deviation of

$$16 \times \frac{6}{4} = \underline{24 \text{ kHz.}}$$

(*iv*) The specified frequency modulation system requires a 4 V test signal to produce 16 kHz deviation. Therefore, for the maximum deviation of 51 kHz, the signal voltage required will be

$$4 \times \frac{51}{16} = \underline{12 \cdot 75 \text{ V.}}$$

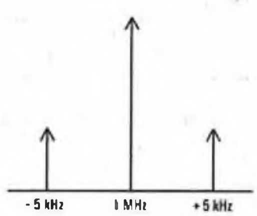
(*b*) The output spectrum from the frequency modulator is shown in sketch (*a*). When this signal is applied to a frequency doubler the output from the doubler circuit will contain the following frequency components:

(*i*) a component at twice the 1 MHz carrier frequency; that is, 2 MHz,

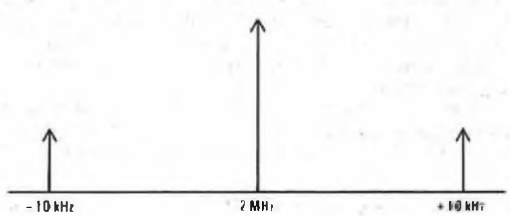
(*ii*) a component at 2 × (1000 − 5) kHz; that is, 1·99 MHz due to the lower side frequency, and

(*iii*) a component at 2 × (1000 + 5) kHz; that is, 2·01 MHz due to the upper side frequency.

The frequency spectrum of the output from the frequency doubler circuit is shown in sketch (*b*).

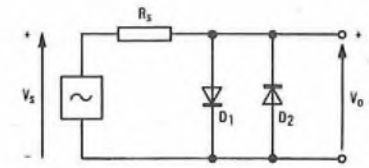


(*a*)



(*b*)

(*c*) Amplitude variations may be removed from a frequency-modulated signal by a limiter circuit. The circuit of an elementary diode limiter or clipper is shown in sketch (*c*). In the circuit shown, both diodes may be conventional silicon diodes and V_s is the varying-amplitude frequency-modulated input signal. Each diode has a forward



(*c*)

threshold voltage V_t and when the input signal is greater then $\pm V_t$ one or other of the diodes conducts and the output voltage V_o is restricted to a maximum of $\pm V_t$. To ensure adequate limiting at all applied input amplitude peaks, the input signal must of course be subjected to adequate linear amplification by, for example, the preceding intermediate-frequency amplifier stages.

Q6 (*a*) *For a 10 kΩ resistor at room temperature the components of noise voltage in a bandwidth of 2 MHz produce an RMS voltage of 17·8 μV.*

(*i*) *State why these noise components are produced.*

(*ii*) *For a bandwidth of 3 MHz, calculate the noise voltage across a 6 kΩ resistor at room temperature.*

(*b*) (*i*) *State what is meant by the input noise temperature of a receiver.*

(*ii*) *When an amplifier stage is screened from external noise it is found that the output signal-to-noise ratio is 6 dB below that at the input. Assuming an ambient temperature of 290 K, calculate the noise temperature of the stage.*

(*iii*) *In what circumstances is it preferable to quote noise temperature rather than noise factor?*

A6 (*a*) (*i*) A metallic resistor contains a number of free-electrons and ions. The ions vibrate randomly around their average positions to an extent determined by their temperature. The free electrons are also moving about continuously in the resistor, colliding with the ions. A continuous transfer of energy takes place between the electrons and ions and this gives rise to voltage fluctuations across the resistor. This voltage has a spectrum which is theoretically constant up to very high frequencies, and it was shown by Johnson and Nyquist that the thermally induced noise voltage, expressed as the mean square noise voltage, $\bar{v}_n^2$, at the resistor terminals, is related to temperature TK, resistance R ohms, and bandwidth B hertz, by

$$\bar{v}_n^2 = 4kTBR,$$

where k is Boltzmann's constant.

Therefore the RMS noise voltage, $\sqrt{\bar{v}_n^2}$, of 17·8 μV is produced by a 10 kΩ resistor at room temperature in a bandwidth of 2 MHz.

(*ii*) From part (*i*), the RMS noise voltage developed across a resistor is defined as

$$\sqrt{\bar{v}_n^2} = 2\sqrt{(kTBR)}.$$

For the 10 kΩ resistor,

$$kT = \frac{\bar{v}_n^2}{4BR} = \frac{17 \cdot 8 \times 17 \cdot 8 \times 10^{-12}}{4 \times 2 \times 10^6 \times 10 \times 10^3}$$

$$= 3 \cdot 96 \times 10^{-21} \text{ J.}$$

Therefore, for a 6 kΩ resistor at room temperature the RMS noise voltage measured in a bandwidth of 3 MHz will be given by

$$\sqrt{\bar{v}_n^2} = 2\sqrt{(3 \cdot 96 \times 10^{-21} \times 3 \times 10^6 \times 6 \times 10^3)} = 16 \cdot 9 \ \mu\text{V}.$$

(*b*) (*i*) The input noise temperature of a receiver is defined as the input termination noise temperature which, when the input termination is connected to a noise-free equivalent of the receiver, will result in the same output noise power as that of the actual receiver connected to a noise-free input termination. For a receiver the input noise temperature, T_e, is related to the noise factor by the equation

$$T_e = (F - 1)T_o,$$

where F is the noise factor and T_o is the ambient temperature, normally 290K.

(*ii*) The noise factor of the amplifier is defined as

$$\text{noise factor, } F, = \frac{\text{input signal-to-noise ratio}}{\text{output signal-to-noise ratio}},$$

or noise figure, F (dB), = input noise figure − output noise figure,

$$= F_{IP} - F_{OP}$$

The output noise figure is known to be 6 dB less than the input noise figure; that is,

$$\text{noise figure} = F_{IP} - (F_{IP} - 6) = 6 \text{ dB}.$$

Therefore, the noise factor of the amplifier is 4. From part (*b*) (*i*) it is known that the effective input noise temperature of an active stage is defined as

$$T_e = (F - 1)T_o.$$

Since the amplifier is shielded from external noise sources and the amplifier termination temperature T_o is at 290K,

$$T_e = (4 - 1) \times 290°K = 870K.$$

(*iii*) The effective noise temperature of an amplifier or receiver is a more meaningful measure than noise factor in the case of very low-noise devices such as parametric amplifiers or state-of-the-art GaAs field-effect-transistor amplifiers, where the noise factors of these units may not differ greatly from unity. A comparison of noise temperatures may more readily show more significant numerical differences.

Q7 (a) When the amplitude of a sinusoidal signal of frequency f is sampled at intervals T by a train of very narrow pulses, the resulting frequency spectrum contains the following components

Frequency	*Relative Amplitude*
f	*0·5*
$\left(\frac{1}{T} - f\right)$	*0·5*
$\frac{1}{T}$	*2·0*
$\left(\frac{1}{T} + f\right)$	*0·5*

Plot to scale the frequency spectra resulting from sampling a 4 kHz sinusoidal signal at intervals of

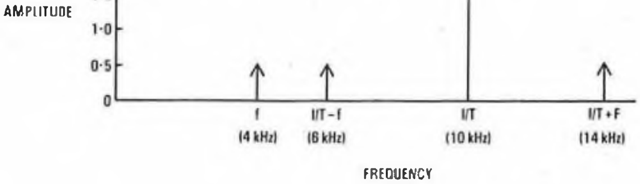
(*i*) *100 μs, and*
(*ii*) *160 μs.*

(*b*) *By use of the diagrams drawn for part (a) explain fully a proposition of fundamental importance in pulse modulation.*

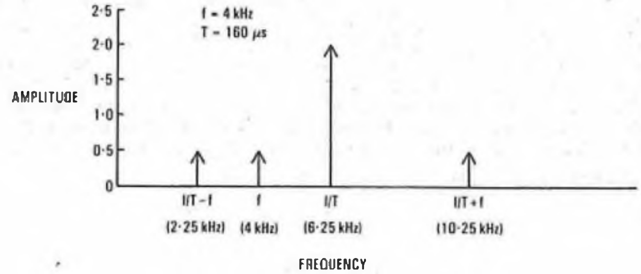
(*c*) *Explain briefly, how TWO channels can be multiplexed in a pulse-amplitude modulation system.*

A7 (*a*) With the 4 kHz sinusoidal signal being sampled at intervals of 100 μs this corresponds to a sampling frequency of 1/100 μs; that is, 10 kHz. Sketch (*a*) shows the resultant frequency spectrum. When the sampling period is changed to 160 μs, the sampling frequency is 6·25 kHz and the new frequency spectrum is shown in sketch (*b*).

(*b*) By studying sketches (*a*) and (*b*) it can be seen that there is a difference between the two frequency spectra resulting from the use of the two different sampling periods. In sketch (*b*) the sinusoidal signal at a frequency of 4 kHz falls inside the spectrum of the complete sampled signal. This will cause distortion of the received sampled waveform and indicates that the sampling frequency has been chosen to be too low. The condition that will cause the modulated signal component to lie just outside the spectrum of the complete sampled signal is when the sampling frequency is at least twice the highest frequency component to be sampled.

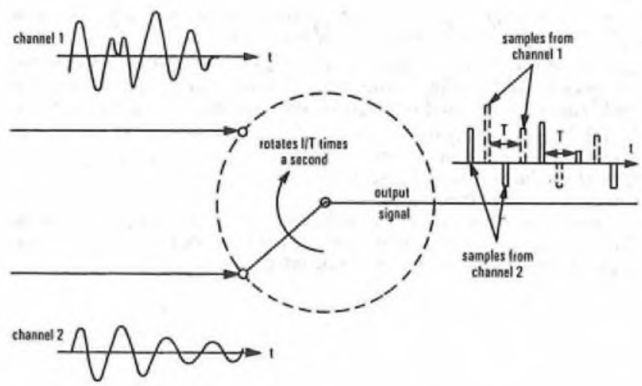


(*a*)



(*b*)

(*c*) Two channels of pulse-amplitude modulated information can be multiplexed using the principle of time-division multiplexing. In this process time-sampled pulse trains derived from the two input signals are interleaved to produce a single pulse train containing alternate samples of the information. Such a system is represented by the rotating switch shown in sketch (*c*). Assuming a maximum modulation frequency of 4 kHz, then in order that the samples of each waveform should be capable of being interleaved, each channel should be sampled at a rate of one sample every 62·5 μs; that is, the sampling switch shown in sketch (*c*) should rotate at 16 000 times a second, a sampling frequency of 16 kHz.



(*c*)

Q8 (a) The frequency changer of a microwave receiver often employs a balanced mixer when the local oscillator is a reflex klystron. Explain why a balanced mixer offers little advantage when a semiconductor local oscillator is used.

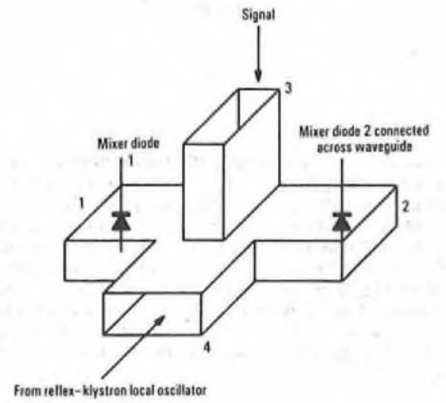
(*b*) *With the aid of a diagram, describe the construction of the frequency-changer stage of a microwave receiver in which the local oscillator is*
EITHER

(*i*) *a reflex klystron,*
OR
(*ii*) *a semiconductor device.*

A8 (*a*) In simple, single mixer-diode frequency changers employed in microwave receivers which use a reflex klystron as a local oscillator, the noise modulation sidebands generated by the klystron as a result of the shot effect have a very wide frequency spectrum and will therefore appear in the intermediate frequency (IF) circuits as additional noise. This undesirable noise can be largely eliminated by employing a balanced mixer, which is frequently based on the use of a pair of mixer diodes in a waveguide magic tee or equivalent coaxial or strip-line circuit.

However, this form of mixer circuit is required less with modern solid-state local oscillators as these have very much reduced noise modulation sidebands compared to the reflex klystron.

(*b*) The sketch shows the construction of a simplified waveguide balanced mixer based on the magic tee. The signal power is fed into the series arm (3 in the sketch), and divides between the two mixer arms with opposite phases. The local oscillator power, suitably attenuated, and its associated noise sidebands, enter at the shunt arm (4 in the sketch), and divide, reaching the mixer diodes in phase. The IF signals will appear at the IF circuitry out of phase, and by using a suitable transformer the signals can be re-phased, and at the same time the local oscillator noise sidebands entering the IF pass band will be reversed in phase and cancel, thus suppressing the noise contribution from the local oscillator.

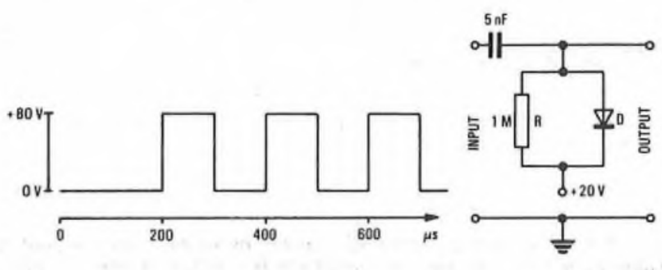


Q9 (*a*) *Explain, by referring to a typical example, why DC restoration may be necessary in the processing of a pulse train.*

(*b*) *The input to the circuit shown in Fig. 2 (b) is the group of three pulses represented by Fig. 2 (a). Draw to scale, indicating voltage levels in each case, time-related waveforms representing*

(*i*) *the input voltage v,*
(*ii*) *the voltage across C, v_C,*
(*iii*) *the voltage across R, v_R, and*
(*iv*) *the output voltage, v_{OUT}.*

Assuming the diode to have a resistance of 1 kΩ when forward biased and a resistance of 5 MΩ when reverse biased, label the output voltage waveform with appropriate time-constant values.



(*a*) (*b*)

Fig. 2

A9 (*a*) See A10 (*a*), Basic Microwave Communication C 1975, *Supplement*, Vol. 69, p. 88, Jan. 1977.

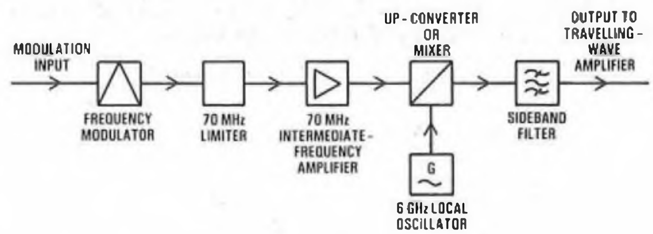
(*b*) See A10 (*b*), Basic Microwave Communication C 1974, *Supplement*, Vol. 68, p. 85, Jan. 1976.

Q10 (*a*) *A 6 GHz microwave transmitter uses a travelling-wave tube in the output stage. The output of the modulator, which is centred on 70 MHz, contains all the frequency components to be transmitted. Draw a labelled block diagram to show how the input for the travelling-wave tube is obtained.*

(*b*) (*i*) *Explain why, in a frequency-modulation (FM) system, pre-emphasis in excess of about 75 μs may be disadvantageous.*

(*ii*) *Pre-emphasis of 50 μs is applied to an FM signal. The de-emphasis network in the receiver is a combination of a 60 kΩ resistor and a 1200 pF capacitor. Explain the effect of this on the output from the receiver.*

A10 (*a*) The block diagram is shown in the sketch.



(*b*) (*i*) In a frequency-modulation (FM) system, pre-emphasis of 75 μs means that the signal is raised in level by 6 dB per octave, starting at a frequency of 2·1 kHz upwards before application to the modulator. When this is used with a matching de-emphasis network in a receiver, the net effect is to produce a constant signal-to-noise ratio in the receiver output. Applying a longer time constant pre-emphasis network will produce excessive modulation (that is, FM deviation) at high frequencies, which will cause an excessive bandwidth signal to be transmitted.

(*ii*) The effect of a de-emphasis network consisting of a series resistor of 60 kΩ and a parallel capacitor of value 1200 pF will be to produce a de-emphasis network suitable for a signal pre-emphasized by a network of 72 μs time constant. If the signal applied to this network had been pre-emphasized with a 50 μs time constant network, the result will be that the output signal will contain enhanced high frequency components of the original signal from 2·1 kHz upwards.

Students were expected to answer any 6 questions

Q1 (a) Write down an expression for the propagation coefficient of a uniform transmission line in terms of its primary coefficients and frequency.

(b) Hence derive an expression for the attenuation coefficient at high frequencies in terms of the primary coefficients only.

(c) Explain why, for high quality coaxial cable, the attenuation coefficient at high frequencies is substantially proportional to the square root of the frequency.

A1 (*a*) The propagation coefficient, γ, of a uniform transmission line is given by

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

where R is the loop resistance (ohms/kilometre), L is the loop inductance (henrys/kilometre), G is the loop leakance (siemens/kilometre), C is the loop capacitance (farads/kilometre), ω is the angular velocity (radians/second) and is equal to $2\pi f$ where f is the frequency (hertz). R, L, C and G are known as the primary coefficients.

γ is a complex quantity, of which the real part is the attenuation coefficient expressed in nepers/kilometre and the imaginary part is the phase-change coefficient expressed in radians/kilometre.

(*b*) Writing $\quad R + j\omega L = j\omega L\left(\dfrac{R}{j\omega L} + 1\right),$

and $\quad G + j\omega C = j\omega C\left(\dfrac{G}{j\omega C} + 1\right),$

gives $\quad \gamma^2 = (R + j\omega L)(G + j\omega C) =$

$$j\omega L\, j\omega C\left(1 + \frac{R}{j\omega L} + \frac{G}{j\omega C} - \frac{RG}{\omega^2 LC}\right).$$

$$\therefore \quad \gamma = j\omega\sqrt{(LC)}\left(1 + \frac{R}{j\omega L} + \frac{G}{j\omega C} - \frac{RG}{\omega^2 LC}\right)^{1/2}.$$

At high frequencies, the term containing ω^2 in the denominator may be neglected. Thus, using the binominal expansion, and neglecting any terms containing ω^2 and higher powers of ω in the denominator, gives

$$\gamma \approx j\omega\sqrt{(LC)}\left\{1 + \frac{1}{2}\left(\frac{R}{j\omega L} + \frac{G}{j\omega C}\right)\right\}$$

$$= j\omega\sqrt{(LC)} + \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{C}{L}}$$

As explained in part (*a*), the attenuation coefficient, α, is the real part of γ. Thus at high frequencies,

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

(*c*) As shown in part (*b*), the attenuation coefficient at high frequencies has two terms. The first term represents the series loss, and is proportional to R, whilst the second term represents the shunt loss, and is proportional to G. At high frequencies, the effective value of R is determined by the skin effect, and thus proportional to the square root of the frequency, while the shunt loss in a high-quality cable is very small and can be neglected.

Therefore, in a high-quality cable, the attenuation coefficient at high frequencies is substantially proportional to the square root of the frequency.

Q2 A 4-wire audio circuit is to be provided between two terminal repeater stations with one intermediate repeater station. The overall loss is to be 3 dB between 2-wire points. The cable pairs between consecutive stations have a loss of 25 dB at 3400 Hz and 16 dB at 300 Hz.

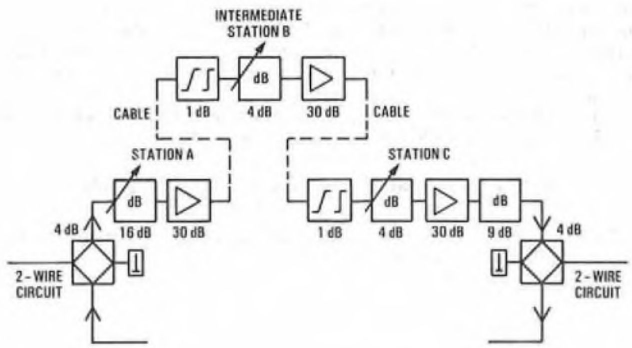
(a) Draw a block schematic diagram to show the arrangement of the various items of equipment which would be needed in the circuit.

(b) Explain the function of EACH item.

(c) Draw a level diagram for one direction of transmission.

(d) Explain the need for imposing upper and lower limits on the signal level at any point.

A2 (*a*) Sketch (*a*) shows a block diagram of the various items of equipment needed to provide the circuit in one direction; the other direction is similar, but is omitted for clarity.



(*a*)

(*b*) Hybrid transformers are used to separate and combine the GO and RETURN unidirectional 4-wire paths with the bi-directional 2-wire paths. Each hybrid transformer has a loss of about 4 dB, of which 3 dB is inherent in the combining/separating processes, and about 1 dB is attributable to practical iron and copper losses.

Each repeater comprises a 30 dB fixed-gain amplifier and an adjustable attenuator which is set to the value shown to give the required affective gain.

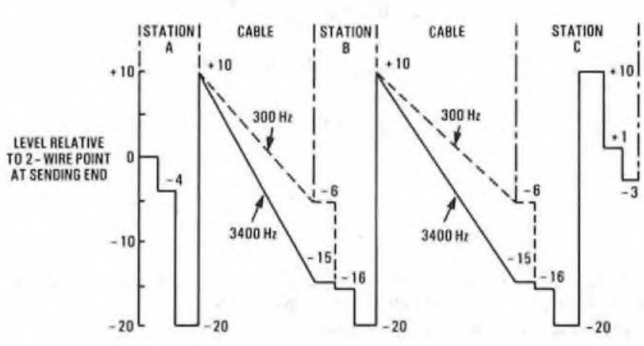
At stations B and C, the variation of attenuation with frequency of the preceding cable section is compensated by means of an equalizer. The equalizers introduce 9 dB more loss at 300 Hz than at 3400 Hz, so that with a basic equalizer loss of about 1 dB, each cable length and associated equalizer presents a loss of 26 dB which is independent of frequency.

The 9 dB attenuator between the final amplifier and hybrid transformer at station C is necessary to reduce the signal level at the receiving 2-wire point to the specified value.

(*c*) Sketch (*b*) shows a level diagram for one direction of transmission.

(*d*) It is necessary to impose limits on the signal level because, if the level is too high, crosstalk can occur between adjacent circuits, and if the signal level is too low, the signal-to-noise ratio becomes so low that the signal is distorted.

A representative range of signal levels is from +10 dB to −20 dB relative to the 2-wire point at the sending end.



(*b*)

Q3 (a) Define balance return loss.

(b) Explain how balance return loss can be measured.

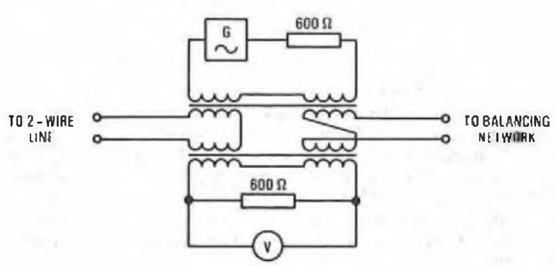
(c) Calculate the balance return loss between a line having a characteristic impedance of (600 − j100) and a balance network of impedance $600\angle -30°$.

A3 (*a*) A 2-wire–4-wire termination provides the transition between the two uni-directional 4-wire paths and the bidirectional 2-wire path. Ideally, none of the energy arriving at the 4-wire RECEIVE terminals should appear at the 4-wire TRANSMIT terminals, so that the 4-wire GO and RETURN paths are completely separate. The success of this separation depends on the degree of matching between the impedance presented by the 2-wire circuit and the balance impedance. Balance return loss is a measure of the accuracy of the matching, and is the value of return loss between the impedances. If the two impedances are perfectly matched, the balance return loss is infinite, and the 4-wire paths are completely separate.

For a balance impedance Z_B, and a 2-wire line having an impedance Z_0, the balance return loss is calculated as

$$\text{balance return loss} = 20 \log_{10} \frac{|Z_0 + Z_B|}{|Z_0 - Z_B|} \text{ decibels.}$$

(*b*) The sketch shows how a hybrid transformer can be used to measure balance return loss.



Two voltmeter readings are obtained: one with the balance network disconnected (V_1), and the second with the balance network connected (V_2). The balance return loss is given by

$$20 \log_{10} \left|\frac{V_1}{V_2}\right| \text{ decibels.}$$

(*c*) Using the equation given in part (*a*),

for $Z_0 = 600 - j100$,

and $Z_B = 600\angle -30° = 600 \cos(-30°) + j600 \sin(-30°)$,

$= 520 - j300$.

$$\text{Balance return loss} = 20 \log_{10} \left|\frac{600 + 520 - j(100 + 300)}{600 - 520 - j(100 - 300)}\right| \text{ dB},$$

$$= 20 \log_{10} \left|\frac{1120 - j400}{80 + j200}\right|,$$

$$= 20 \log_{10} \left|\frac{1189}{215}\right|,$$

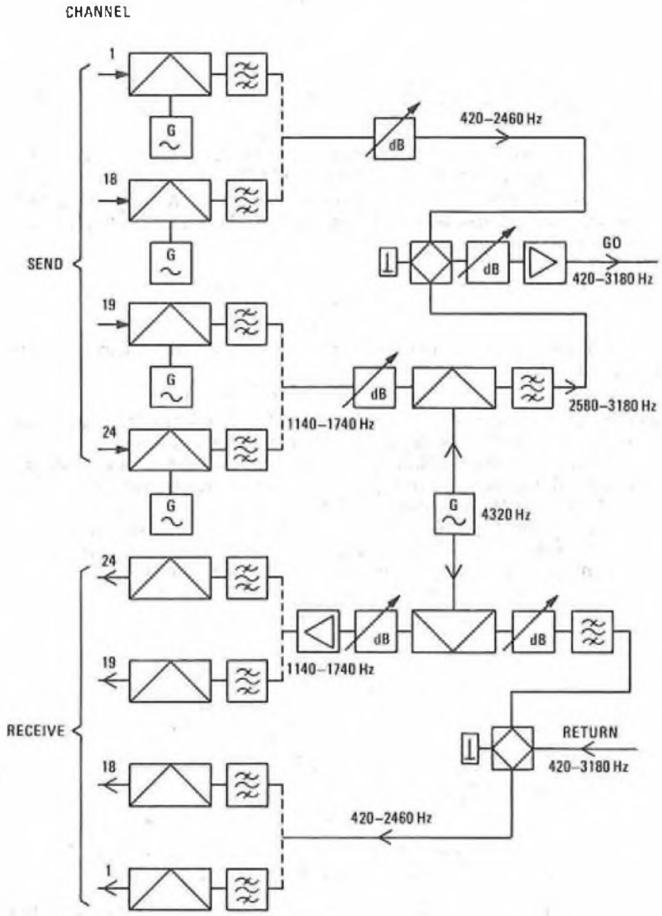
$$= \underline{14{\cdot}8 \text{ dB.}}$$

Q4 (a) Draw the block schematic diagram of the equipment for one end of a 24-channel voice-frequency telegraph system.

(b) Explain the function of EACH *block.*

(c) State how and why the requirements for the channel filters differ from those for a multi-channel carrier-telephone system.

A4 (*a*) The block diagram is shown in the sketch.



(*b*) A 24-channel voice-frequency telegraph system consists of an 18-channel system in the frequency range 420–2460 Hz together with a 6-channel system in the range 1140–1740 Hz which is then modulated into the frequency range 2580–3180 Hz to give line frequencies in the range 420–3180 Hz.

Transmit Each channel consists of a modulator (or static relay) where the channel carrier frequency is either shunted by a low impedance when the sending teleprinter is sending a SPACE element, or allowed to pass to line when a MARK element is being transmitted. A band-pass filter in each TRANSMIT channel provides separation between the channels.

For the 18-channel system, an attenuator then reduces the input to the terminating unit to a suitable level. For the 6-channel system, an attenuator reduces the level to a suitable value prior to it modulating a 4320 Hz carrier to produce a band in the range 2580–3180 Hz. The 6-channel system then passes through a band-pass filter which provides separation from the 18-channel system. The two systems are then combined in the terminating unit. The combined 24 channels then pass through an attenuator and the SEND amplifier to line.

Receive The received signals are separated by band-pass filters into the individual channels for channels 1–18 and fed through individual amplifiers for transmission to the respective teleprinter. However for channels 19–24 the signals are first demodulated as a block then attenuated and passed into their individual channels via band-pass filters and amplified to the receiving teleprinter.

(*c*) In any carrier system, it is essential to make the best use of the available bandwidth, and the cost of channel filters is an important consideration. For a telegraph system, the required capacity can be obtained by using fairly inexpensive channel filters with a gradual cut-off characteristic, because the bandwidth needed for each channel is comparatively small. For a carrier-telephony system, however, a much larger bandwidth is required for each channel, and it becomes economic to provide more expensive filters with a sharper cut-off characteristic, in order to make the best use of the frequency spectrum available.

Q5 (a) Explain how the attenuation of a transmission line can be reduced by the addition of loading coils.

(b) Why does the reduction occur only over a limited frequency range?

(c) Derive an expression for the cut-off frequency of a loaded line in terms of the inductance of the coils, the spacing between them and the capacitance of the line.

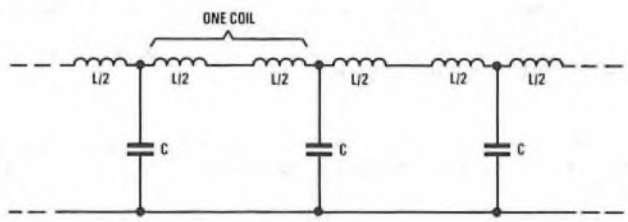
A5 (*a*) The two factors which cause dissipation of energy in a transmission line, and so cause attenuation, are the loop resistance, R, and the loop leakance, G. At any point in a uniform line carrying a current I at a voltage V, the series loss due to the loop resistance is I^2R, and the shunt loss due to the loop leakance is V^2G. In modern cables, the I^2R losses predominate because G is negligible over the frequency range used. It follows that the attenuation can be reduced if the current I is reduced.

In modern cables, the loop capacitance (C) is high, and the addition of series inductance in the form of loading coils tends to improve the power factor, so that Heaviside's distortionless condition ($LG = RC$) is approached more closely. It was shown in the answer to question 1 that at high frequencies the attenuation coefficient,

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

in which the first term represents the dominant series loss, and the second term represents the shunt loss which can be ignored. It can thus be seen that, since $\alpha \approx \frac{R}{2}\sqrt{\frac{C}{L}}$, an increase in the series inductance reduces the attenuation.

(*b*) Coil loading has the effect of converting the line into a number of tandem low-pass filter sections as shown in sketch (*a*):

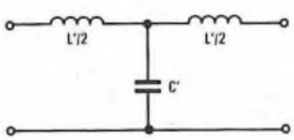


(*a*)

Note: The resistance and inductance of the line itself are ignored.

If such loading were continuous, the reduction in attenuation would not occur only over a limited frequency band; however, because in practice the loading is lumped at discrete intervals along the line, the combination of line and lumped loading exhibits the characteristics of a low pass filter.

(*c*) Consider one filter section as shown in sketch (*b*).



(*b*)

This filter has a low attenuation below its cut-off frequency, f_c, and a rapidly rising attenuation above it. Cut-off occurs at the resonance point, at which the reactance is zero, and

$$\frac{\omega L'}{4} = \frac{1}{\omega C'}$$

in which $\omega = 2\pi f_c$,

L' is the inductance of each loading coil, and

C' is the capacitance of the section of the line between loading coils.

Thus $$\omega^2 = \frac{1}{4L'C'},$$

and $$f_c = \frac{1}{\pi\sqrt{(L'C')}}.$$

Since C' is the total capacitance of the line between loading coils, $C' = Cd$, where C is the loop capacitance in farads/kilometre, and d is the separation between the coils in kilometres.

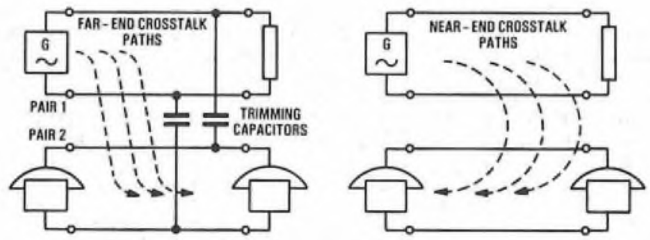
Thus $$f_c = \frac{1}{\pi\sqrt{(L_{COIL}\,Cd)}}.$$

In practice, the cut-off is not sharp, and the maximum usable frequency is about 80% of the theoretical cut-off frequency.

Q6 (a) Distinguish between near-end and far-end crosstalk in a transmission system.

(b) Explain what is meant by cable balancing. Show how it may be used to reduce far-end crosstalk to an acceptable level in a quad-type carrier cable.

A6 (*a*) Sketches (*a*) and (*b*) both illustrate 2 pairs in a cable, with a signal in pair 1 causing interference signals in pair 2. The interference signal appearing at the right-hand end of pair 2 is called *far-end crosstalk* (sketch (*a*)), and that appearing at the left-hand end is called *near-end crosstalk* (sketch (*b*)).



(*a*) (*b*)

(*b*) Crosstalk can be caused by

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

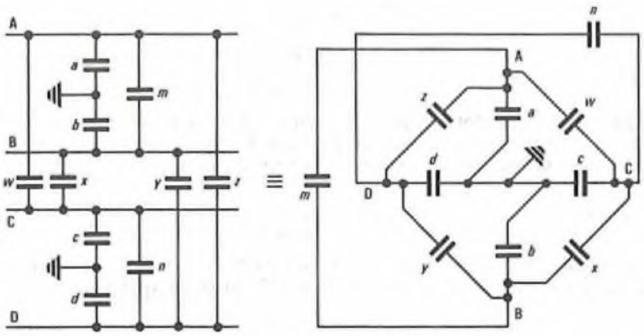
Some of these factors can be minimized by sound design, and by ensuring uniformity during manufacture; others by careful installation and maintenance procedures.

The 4 wires of each quad are taken from the same reel, thus eliminating any possible differences in diameter due to the drawing die, and all are insulated with paper ribbons cut from the same roll. The identification marks, printed on the paper in the form of 1, 2, 3 or 4 rings, are spaced in such a way as to use the same amount of ink in a given length for each wire of the quad, thus obviating any difference in insulation resistance attributable to the ink itself.

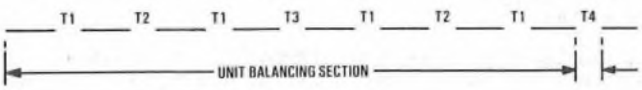
All quads have different lengths of lay, some being laid clockwise and some counter-clockwise. The cable is built up in layers of quads, each layer being laid helically over the one below it and in the opposite direction of lay. Thus, all quads in a layer have the same direction of lay, but adjacent layers have clockwise and counter-clockwise quads. This arrangement of layers and quads minimizes the possibility of inductive coupling between quads. The completed cable core is thoroughly dried in an oven before the sheathing is applied, and it is important that moisture should not be allowed to enter the core, and hence reduce the insulation resistance, during the subsequent jointing operations.

In practice, the main source of crosstalk is capacitance unbalance. It is necessary to make measurements on each cable length, and then to select quads for jointing in such a way as to even out and minimize the overall unbalance figures. Such joints are called *test-selected joints*.

Sketch (*c*) shows the various capacitances involved in a star quad with wires A, B, C and D. No crosstalk occurs within the quad if $w = x = y = z$, $a = b$, $c = d$, and $m = n$.



(*c*)



(*d*)

Sketch (*d*) shows a typical unit balancing section for a carrier cable system. It comprises 8 lengths of cable, each about 160 m long, and every joint is test selected. The cable lengths are first jointed in pairs (joints T1) and then in groups of four (joints T2). Selected joint T3 completes the unit balancing section. A number of such sections are then selectively jointed together (joints T4) to build up the cable to the required length. It is possible to carry out much of the measurement and selection work in the factory before the cable lengths are delivered to the site. However, it is usually necessary to test and select joints T3 and T4 *in situ*.

For a carrier system, it is usual to use separate cables for each direction of transmission. On completion of the balancing process, small trimming capacitors are added at the receiving end, as required, to compensate for any residual unbalance, as shown in sketch (*a*). For this purpose, a special balancing frame is provided at the terminal station. Although there are various paths over which far-end crosstalk can occur, they all have the same electrical length. Thus, all components of the total far-end crosstalk arrive in the same phase, and their effect can be substantially eliminated by the use of simple capacitors. This contrasts with the conditions for near-end crosstalk, where the various paths have different lengths (see sketch (*b*)); thus, the components of the total near-end crosstalk cover a range of phase relationships. Near-end crosstalk is avoided by making use of the screening effect of the lead sheaths.

Q7 (*a*) *Outline the various transmission tests which would be needed before a new design of telephone set could be accepted for installation in an existing network. Distinguish between subjective and objective tests.*

(*b*) *State what is meant by sidetone and explain its effect.*

A7 (*a*) 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 pure tones 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 responses 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 responses relative to the exchange termination.

Subjective tests 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 staff 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 addition, pairs of representative subjects are asked to use a telephone circuit and to give an opinion of 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.

For a description of the latest methods of testing telephone sets see p. 25 of the current issue of the *Journal*.

(*b*) 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 the telephone circuit design, but it depends to some extent on the degree of impedance matching between the telephone set and the exchange line. Sidetone has a reassuring effect on 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 voice 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.

Q8 (*a*) *Draw block diagrams to show the power supply arrangements for the transmission equipment in a repeater station. Explain how interruption is avoided in the event of a mains failure.*

(*b*) *Outline the steps taken to prevent the power supply from interfering with the transmission circuits.*

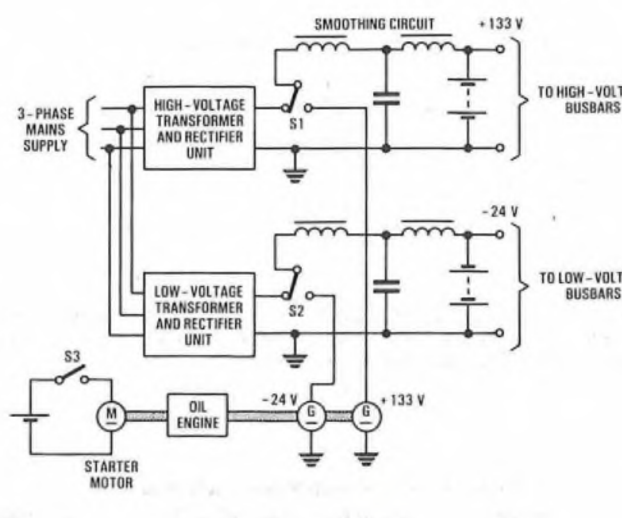
A8 (*a*) The primary source of power is usually the mains supply, and there are 2 main methods of using this for repeater-station equipment:

(*i*) by using batteries that are float-charged from the mains supply, and distributing to individual racks by busbars, or

(*ii*) by feeding a mains-type supply to small power units on individual racks.

The essential requirements are good regulation and absolute reliability. The former is fairly easy to achieve because the load is constant (in contrast to the load in a telephone exchange), but the latter requires some form of alternative supply in case of mains failure. The alternative supply is usually provided by a prime mover, such as an oil engine, coupled to an alternator or DC generators.

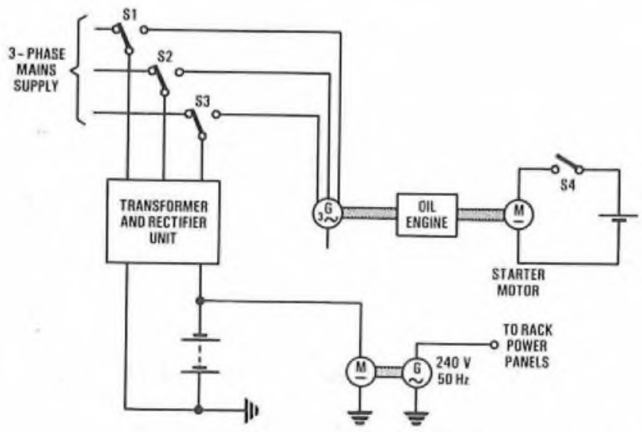
The example given in this answer pertain to a repeater station having tube-operated transmission equipment.



(*a*)

A typical arrangement for method (*i*) is shown in sketch (*a*). In the event of mains failure, switch S3 operates to start the oil engine. No interruption occurs because the batteries continue to feed the equipment while the engine runs up to speed. Switches S1 and S2 then transfer the loads to the generators.

Sketch (*b*) shows a typical arrangement for method (*ii*). The individual power units provide an AC heater supply and a high-voltage DC anode supply for the tubes on their racks. An alternator feeds the individual power units. This alternator is coupled to a DC motor driven by the rectified mains supply, backed up by a floated battery. In the event of mains failure switch S4 operates to start the oil engine, and when the correct running speed is reached, switches S1, S2 and S3 operate so that the 3-phase alternator takes over the supply. No interruption occurs because the DC motor continues to run from the battery while the engine runs up to speed.



(*b*)

(*c*) Noise can enter the transmission circuits from the power supply if the smoothing arrangements are inadequate. With floated batteries, the smoothing must be sufficient to suppress armature ripple from the generators. With individual power units, the smoothing must be sufficient to suppress mains hum from the rectifiers.

Mutual interference between transmission circuits can occur because of the common supply, and this can be suppressed only by making the internal impedance of the supply very low. This requires a very small impedance at signal frequencies, which is usually achieved by the use of decoupling capacitors.

Q9 *(a) Draw the circuit of a bridge suitable for measuring the impedance of an audio frequency cable pair.*

(b) Derive an expression for the modulus and angle of the cable impedance in terms of the bridge components.

(c) State the precautions necessary to ensure a reasonable degree of measurement accuracy.

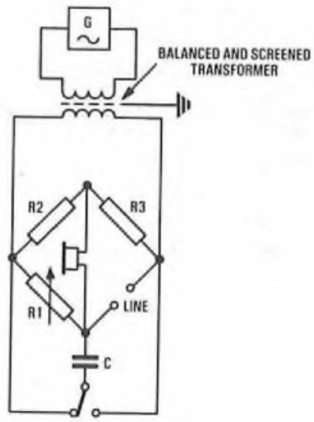
A9 (*a*) The sketch shows the circuit of a bridge suitable for measuring the impedance of an audio frequency cable pair.

R2 and R3 are fixed ratio-arms (typically 100 Ω each) and C is a variable capacitor which can be switched across the variable resistance R1 when the line impedance being measured has a small negative angle, or across the line itself when its impedance has a small positive angle. A high impedance telephone receiver acts as a detector, and the frequency source is connected to the bridge by means of a balanced and screened transformer.

(*b*) Assume that, in order to balance the bridge, C is switched as shown in the sketch. Writing:

Z_L as the impedance of the line being measured
Y as the impedance of C in parallel with R1

At balance $$R_2 Z_L = R_3 Y.$$



$$\therefore \quad Z_L = \frac{R_3}{R_2} Y.$$

Now, $$Y = \left(\frac{1}{R_1} + j\omega C\right)^{-1} = \frac{R_1}{1 + j\omega C R_1}.$$

Separating the real and imaginary parts,

$$Y = \frac{R_1}{1 + \omega^2 C^2 R_1^2} - j\frac{\omega C R_1^2}{1 + \omega^2 C^2 R_1^2},$$

or in polar form,

$$|Y| = \frac{R_1}{\sqrt{(1 + \omega^2 C^2 R_1^2)}}, \text{ and}$$

$$\arg Y = \tan^{-1}(-\omega C R_1).$$

Thus, $$|Z_L| = \frac{R_3}{R_2}\,\frac{R_1}{\sqrt{(1 + \omega^2 C^2 R_1^2)}}, \text{ and}$$

$$\arg Z_L = \tan^{-1}(-\omega C R_1).$$

In a similar way, it can be shown that if it is necessary for C to be switched across the line to balance the bridge,

$$Z_L = \frac{R_1 R_2 R_3}{R_2^2 + \omega^2 C^2 R_3^2 R_1^2} - j\frac{\omega C R_3^2 R_1^2}{R_2^2 + \omega^2 C^2 R_3^2 R_1^2}.$$

$$\therefore \quad |Z_L| = \frac{R_1 R_3}{\sqrt{(R_2^2 + \omega^2 C^2 R_3^2 R_1^2)}}, \text{ and}$$

$$\arg Z_L = \tan^{-1}\left(\frac{R_3}{R_2}\omega C R_1\right).$$

(*c*) To ensure accuracy of measurement, the following precautions are necessary:

(*i*) each component must be screened, and all the screens connected to a common earth point,

(*ii*) the oscillator must be connected to the bridge through a balanced screened transformer,

(*iii*) the oscillator must be free from harmonics; a suitable filter can be used in either the source or detector leads to assist with balancing the bridge if harmonics are present,

(*iv*) the detector must be of high quality, and

(*v*) all components must be accurately calibrated.

Q10 *(a) Write down the expression for the characteristic impedance of a uniform transmission line in terms of its primary coefficients.*

(b) Calculate the characteristic impedance of a uniform line at $\omega = 5000$ rad/s given that $R = 40$ Ω/km, $L = 2$ mH/km, $C = 0{\cdot}1$ μF/km and $G = 1$ μS/km.

(c) Explain how the characteristic impedance of a line can be obtained from two separate impedance measurements.

A10 (*a*) For a uniform transmission line, the characteristic impedance, Z_0, is given by the expression

$$Z_0 = \sqrt{\left\{\frac{R + j\omega L}{G + j\omega C}\right\}} \text{ ohms},$$

where R is the loop resistance in ohms/kilometre,
L is the loop inductance in henrys/kilometre,
G is the loop leakance in siemens/kilometre,
C is the capacitance in farads/kilometre, and
ω is the angular velocity in radians/second.

(*b*) Substituting the values given,

$$R + j\omega L = 40 + j5000 \times 2 \times 10^{-3},$$
$$= 40 + j10 = 41{\cdot}2 \angle 14{\cdot}0^\circ.$$
$$G + j\omega C = 10^{-6} + j5000 \times 0{\cdot}1 \times 10^{-6},$$
$$= 10^{-6}(1 + j500) = 500 \times 10^{-6} \angle 89{\cdot}9^\circ.$$

$$\therefore \quad |Z_0| = \sqrt{\left(\frac{41{\cdot}2}{500 \times 10^{-6}}\right)} = 287\ \Omega, \text{ and}$$

$$\arg Z_0 = \frac{14{\cdot}0 - 89{\cdot}9}{2} = -38{\cdot}0^\circ,$$

$$\therefore \quad \underline{Z_0 = 287 \angle -38^\circ.}$$

(*c*) The characteristic impedance of a transmission line can be calculated from bridge measurements of the input impedance of the line when the far-end is (*a*) open circuited (Z_{oc}) and (*b*) short circuited (Z_{sc}). Under these circumstances,

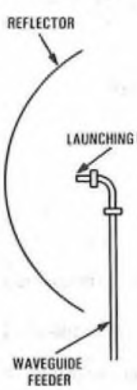
$$\underline{Z_o = \sqrt{(Z_{oc} \times Z_{sc})}.}$$

LINE PLANT PRACTICE B 1980

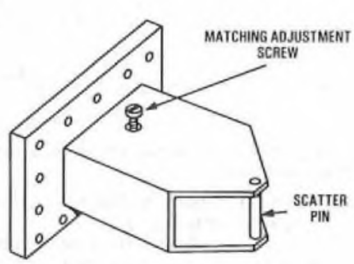
Students were expected to answer any 6 questions

Q1 Describe, with the aid of sketches of the equipment, an aerial and feeder suitable for use in a cross-country microwave-radio link.

A1 The type of aerial normally used in microwave systems is the parabolic-reflector type of aerial shown in sketch (*a*). The aerial consists essentially of a cast or spun-aluminium reflector (which may be between 2·1 and 3·6 m in diameter), a launching unit located at the focus of the parabolic reflecting surface, and a feeder connecting with the transmitter or receiver.



(*a*)



(*b*)

The launching unit consists of either a small waveguide horn or a cut-corner waveguide feed, the purpose of either being to broaden the beam radiated by the waveguide feeder. The most commonly used launcher or radiator is the cut-corner launching unit shown in sketch (*b*). A scatter pin is used to improve impedance matching.

The waveguide feeder assembly consists of straight lengths (each about 3 m long) and large-radius bends, fitted with jointing flanges at both ends. To prevent the ingress of moisture the feeder is sealed by means of a radio-transparent window at the junction with the launching unit. To prevent snow and ice remaining on the waveguide feed, and thereby causing loss and mismatch, a small electric heater is mounted at the end of the waveguide bend.

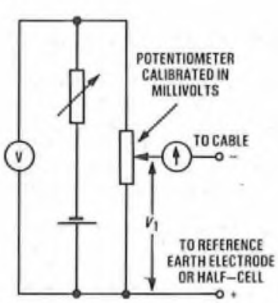
Q2 A lead-sheathed cable has corroded as a result of a stray sheath current.

(*a*) *Describe, with the aid of a diagram, a method of measuring the potential difference between the cable sheath and earth.*

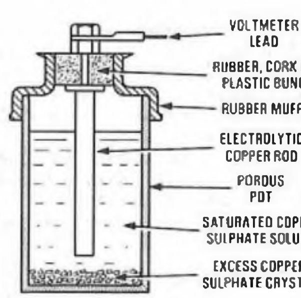
(*b*) *Describe, with the aid of a sketch,* ONE *type of earth electrode which would be used in the test.*

A2 (*a*) The potential difference between a lead cable sheath and earth can be measured by means of a high-resistance voltmeter and a reference earth electrode. A tester normally used for cable-sheath potential measurement contains a high-resistance moving-coil voltmeter with a coil resistance of 40 kΩ/V. Three testing ranges (of 1 V, 2 V and 10 V) are provided. A galvanometer, local battery and potential divider are required in addition to the voltmeter. These elements form a potentiometer which is used for measurements within the 1 V range. On the 2 V and 10 V ranges the voltmeter is used as a direct-reading instrument. The circuit diagram in sketch (*a*) shows the tester switched to the 1 V range.

If the cable-sheath potential is within the 1 V range, the voltage across the potential divider is adjusted using the variable resistor until the voltmeter registers exactly 1 V. The positive terminal of the tester is connected to the reference earth electrode and the negative terminal is connected to the cable sheath. The potential divider is adjusted until a zero reading is obtained on the galvanometer. The cable-sheath potential is then read directly from the position of the potentiometer, which is calibrated in millivolts.



(*a*)



(*b*)

(*b*) One form of earth electrode used in cable-sheath potential measurements is the copper-sulphate half-cell shown in sketch (*b*). The cell consists of a porous pot containing a rod of pure copper immersed in a copper-sulphate solution. The pot is placed on the floor of the jointing chamber (which must be wetted if necessary to ensure that intimate contact is made between the half-cell and the earth surrounding the cable sheath). The terminal fitted to the copper rod is connected by a lead to the voltmeter. A half-cell form of reference electrode enables contact to be made between one terminal of the tester and earth in such a manner that the junction potential between the electrode and earth is minimized.

Q3 (*a*) *Explain how an earth fault on a high-voltage overhead power line can cause an induced voltage in nearby overhead and underground telephone plant.*

(*b*) *State a formula for the magnitude of the induced voltage in* (*a*).

(*c*) *State briefly the effects, under fault conditions, arising from power systems in which the neutral point is earthed*

(*i*) *solidly,*

(*ii*) *via a resistance, and*

(*iii*) *via an arc-suppression coil.*

A3 (*a*) The neutral point of a high-voltage AC power transmission system is earthed at the transformer station. When the system is operating normally, the instantaneous value of the current in any conductor of a high-voltage power line is equal and opposite to the instantaneous sum of the currents in the other conductors, and induction effects from the power conductors therefore tend to cancel each other. If an earth fault occurs on a conductor, a heavy current will flow along the conductor and back via earth to the transformer station. This will upset the balance of the currents in the power conductors, resulting in a considerable alternating magnetic field along the route of the power line from the point of the fault back to the transformer station. A telephone line (either overhead or underground) which is parallel to the power line within this magnetic field for any considerable distance will have longitudinal voltages induced in it which could reach dangerous values.

(*b*) Induced voltage $= 2\pi flMI$ volts, where f is the frequency of the power system in hertz, l is the length of exposed line in metres, M is the mutual inductance per unit length between the two lines in henrys/metre, and I is the earth-fault current in amperes.

(*c*) (*i*) Where the neutral is solidly earthed, the power system will pass so much current through an earth fault that the circuit breakers will operate almost immediately and eliminate the dangerous conditions.

(*ii*) If the system is earthed through a resistance, the earth-fault current will be limited and the circuit breakers may not operate. Induced voltages will be limited but interference may still occur until the fault is cleared.

(*iii*) If the system is earthed through an arc-suppression coil, the reactance of the coil is adjusted so that under earth-fault conditions the phase voltage applied to the coil will produce a lagging current which is equal in magnitude and opposite in phase to the total capacitive current returning to the fault. The resultant current at fundamental frequency is therefore negligible. The circuit breakers will not operate in these circumstances and dangerous conditions will not exist.

Q4 (*a*) *What factors influence the transmission efficiency of a local line circuit?*

(*b*) *The cabling arrangement shown in Fig. 1 is planned to serve part of an exchange area. From the values given in the table, what distance would be possible between the cabinet and the DP using 0·5 mm aluminium conductors to keep the lines within transmission and signalling limits? Allow a 10% reduction for dropwire connexions.*

Conductor Diameter	*DC Loop Resistance Ω/km*	*Attenuation dB/km*
0·4 mm	270	*2·2*
0·5 mm	170	*1·7*
0·5 mm (Al)	280	*2·5*

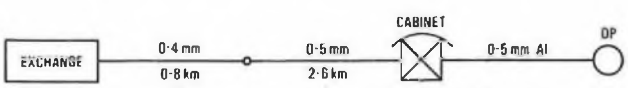


Fig. 1

A4 (*a*) The transmission efficiency of a local line circuit is influenced by the loop resistance and mutual capacitance of a pair. For a given loop resistance a cable pair with heavy-gauge conductors will cover a greater distance than a cable pair with lighter conductors but, because of its greater capacitance, gives an inferior transmission performance.

(*b*) For the 0·40 mm diameter conductor section, the loop resistance is $270 \times 0·8 = 216\ \Omega$ and the transmission loss is $2·2 \times 0·8 = 1·76$ dB.

Similarly, for the 0·5 mm diameter conductor section the loop resistance is 442 Ω and the transmission loss is 4·42 dB.

Thus the signalling resistance available for the 0·5 mm aluminium-conductor section is $1000 - (216 + 442) = 342\ \Omega$, giving a possible distance of $342/280 = 1·22$ km. Taking into account a 10% reduction adjusts this to $1·22 \times 90\% = 1·1$ km.

The maximum allowable transmission loss of the 0·5 mm aluminium-conductor section is $10 - (1·76 + 4·42) = 3·82$ dB, giving a permissible length of $3·82/2·5 = 1·53$ km. Taking account of the 10% reduction adjusts this to 1·38 km.

The signalling limit is thus the more stringent, and the possible distance is therefore 1·1 km.

Note: At the time of this examination, the signalling and transmission limits assumed in the answer were 1000 Ω and 10 dB respectively.

Q5 *An underground-plant project is to be undertaken.*

(*a*) *Describe the functions of planning and control.*

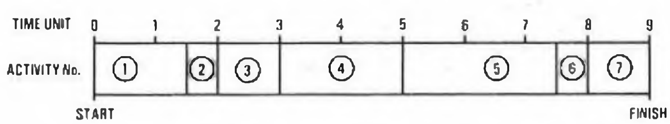
(*b*) *Describe briefly the* THREE *resource categories which are under the control of the planner.*

(*c*) *Describe a method of displaying the planned progress of the job.*

A5 (*a*) The function of project planning and control is to ensure that the project is completed as efficiently as possible within a specified time. The time available for a particular project will usually be dictated by factors such as the project's position in an overall plan or the need to meet a customer's requirement. The project planner has therefore to organize the available resources so that the most efficient use is made of them within the specified time.

(*b*) Three resource categories are under the control of the project planner. These are: components and materials consumed by the project (stores), tools and equipment used during the execution of the project (plant), and the workforce (labour). In order to forecast the quantity or degree needed of each category, and to forecast when each will be required on site, it is necessary to construct an accurate plan of the project presented, taking account of its chronology.

(*c*) When a project consists of a simple sequence of events, with each event starting after the completion of its predecessor, a simple bar-chart can be used to display the planned progress of a job. A project is divided into a numbered sequence of activities and a duration is assigned to each activity. The sketch shows how a sequence of 7 activities would be displayed in the form of a bar-chart.



Q6 (*a*) *With the aid of sketches, describe in detail the equipment which would be used to trace the route and depth of a buried armoured cable.*

(*b*) *Describe how the equipment in* (*a*) *would be used.*

A6 (*a*) It is important to be able to locate underground cables and cable tracks without the need to excavate, as this is both costly and time consuming. The principle of operation of the present equipment is that a 1 kHz signal is connected to the cable conductors, armouring or sheath, or to a metallic service, the signal setting up electric and magnetic fields along the route of the cable or service. At any point along the line of the route, the cable or service may be located and identified, or its depth determined, by detecting the alternating magnetic field with a search coil, the detected signal being amplified and fed to a headgear receiver.

The 1 kHz signal is produced from a transistor oscillator and the tone can be continuous or interrupted.

The search coil is a 180 mm long, rectangular coil, wound on a polystyrene former over a flat mu-metal core. A capacitor is connected in parallel with the coil to tune it to have maximum sensitivity at approximately 1 kHz.

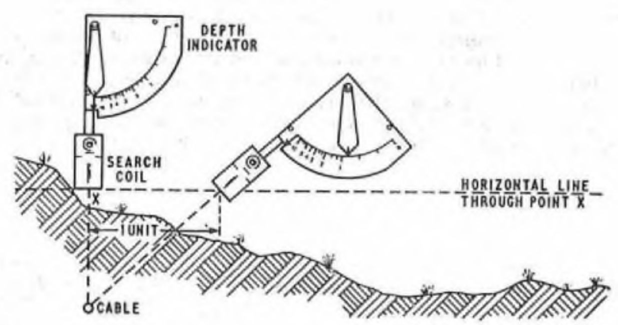
A transistor amplifier, with a variable gain of up to 70 dB, is used. The amplifier is connected to the search coil by a coaxial cable, and to a rocking-armature-type headgear receiver.

An aluminium plate, engraved with a quadrant depth-scale, with a freely moving pointer, serves as a depth-indicator.

(*b*) At a suitable point in the line of the route of the cable to be located, whether at the main distribution frame, cabinet, pillar, distribution point, or any other intermediate point, the cable pairs are bunched and connected to the CABLE terminal of the oscillator. If the buried cable is a working cable, the armouring may be used as a metallic path, provided that the armouring is bonded across any joints, and that the d.c. resistance is less than 100 ohms. The REMOTE EARTH terminal is connected to a low-resistance earth, either via an earth-spike driven into the ground to a depth of 0·75–1 m, approximately 8–9 m from the cable, or via existing earths. These existing earths may be used only if they have a low-enough resistance, preferably less than 100 ohms. The cable pairs, or armouring, at the distant end are also connected to a low-resistance earth, so that a strong magnetic field is produced by the current flowing in the cable.

The operator commences the search for the buried cable by moving the search coil from side to side across the line of the route of the cable. Initially, the search coil is operated on its side, with the marking, MAXIMUM, uppermost. In this position, the search coil produces a maximum-amplitude signal in the headgear receiver when it is directly above the cable. To obtain a more accurate location, the search coil is used upright, so that the marking, MINIMUM, is uppermost. In this position, the operator will not hear any signal when the search coil is directly above the cable; this is known as the *null point*. Where it is necessary to record the line of the route of the cable, marking pegs are placed in the ground directly above the cable, and measurements taken as necessary.

To ascertain the depth of the cable, the depth-indicator is fitted to the search coil, and the cable is accurately located using the null-point method, and its position marked on the ground. The depth-indicator is then held a chosen distance to one side of the cable and tilted, so that the flat base points in the general direction of the cable. The relative positions of the search coil should be in the same horizontal plane. The angle of tilt is adjusted until a further null point is found, and the pointer then indicates the depth of the cable, in the same units as the distance by which the coil was originally displaced. To increase the accuracy, measurement is made on the other side of the cable in a similar manner, and the average of the two readings taken. The sketch illustrates the procedure.



An alternative method is to tilt the coil at one side of the cable until the pointer reads unity, indicating an angle of 45°. The search coil is then moved horizontally until the null point is determined, the distance moved being, then, the same as the depth of the cable.

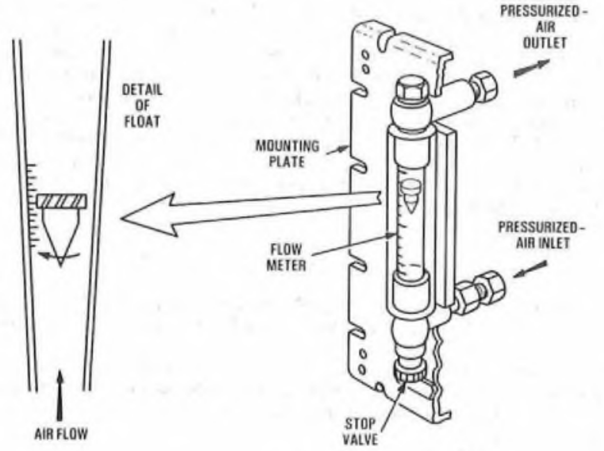
This apparatus, and method of operation, under good conditions, can detect cable for considerable distances, and at depths of up to 4·5 m.

Q7 Describe, in detail and with the aid of sketches, the following items of equipment used in cable pressurization:

(a) an air-flow indicator, and

(b) a pressure-alarm contact-gauge.

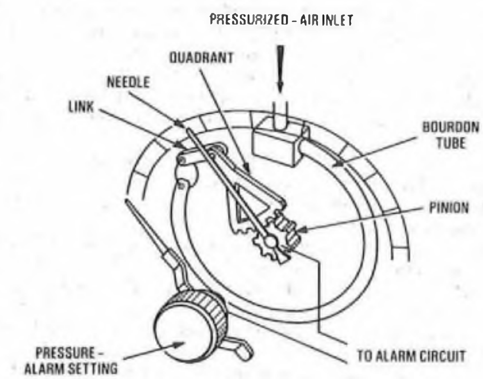
A7 (*a*) Air-flow indicators are used in conjunction with continuous-flow gas-pressure schemes to measure the rate of flow of air at cable inputs. The flow indicator shown in sketch (*a*) consists of a vertical tapered glass tube containing a cone-shaped float which acts as a metering element. The float is free to move in the tube and, when air is allowed to flow up the tube, the float is lifted. The float will move up the tube until a position of equilibrium is reached, determined by the



(*a*)

annular gap between the rim of the cap of the float and the side of the tube. If the rate of air flow is increased, the float will move further up the tube until the area between the cap and the side of the tube is again sufficiently large to give equilibrium. Similarly, if the flow rate is decreased, the float will drop to a narrower part of the tube. The glass tube is graduated to include the range 4–40 g/h and the top of the float indicates the rate of air flow. Oblique grooves in the cap of the float cause the air flow to rotate the float and prevent it from sticking to the glass.

(*b*) A contact-gauge is a pressure gauge of the Bourdon type, fitted with a pointer-operated electrical-contact assembly. The main features of a contact gauge are shown in sketch (*b*).



(*b*)

The Bourdon tube is made of copper alloy and is elliptical or flattened in cross-section and bent to form a coil. The sealed end of the tube is coupled by a link to a pivoted quadrant. Teeth on the quadrant mesh with the teeth of a pinion to which a pointer is attached. An increase in pressure in the tube causes the tube to uncoil slightly and thus turn the pointer. Usually, the case of the gauge is not sealed, so that the tube is subject to atmospheric pressure and the value of the measured pressure is the difference above atmospheric pressure.

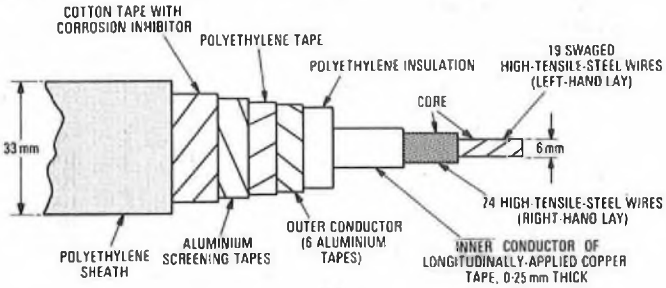
The pointer carries one contact, and a second contact is fitted to a spring-loaded manually-adjusted arm. The pointer travels over a calibrated scale of range 0–650 mbar, and the adjustable arm may be positioned at any point over the range. The contacts are connected to a spare cable pair which terminates on an alarm circuit. Thus the contact gauge may be set to operate an alarm circuit when the gauge pointer reaches any predetermined setting.

Q8 With the aid of sketches, describe

(a) the make-up of a lightweight deep-sea telephone cable, and

(b) the method of jointing the core and centre conductor only of a lightweight deep-sea cable.

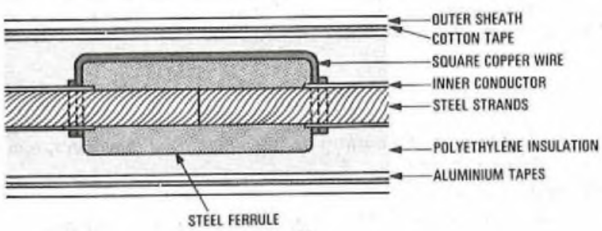
A8 (*a*) A section of lightweight deep-sea coaxial cable is shown in sketch (*a*).



(*a*)

The high-tensile-steel core is torsionally balanced so that it has no tendency to twist under tension. The copper inner conductor is applied tightly over the core and closed with a folded box-seam. Polyethylene insulation is applied to the inner conductor to increase the diameter to 25 mm. Six 0·45 mm thick aluminium tapes are then closely applied to the surface of the polyethylene to form the outer conductor. These are separated from aluminium screening tapes by a layer of polyethylene tape. The screening tapes are held in position by cotton tape, impregnated with barium chromate to prevent the aluminium tapes from corroding. A tightly-extruded sheath of polyethylene over the cotton tape then completes the cable.

(*b*) A joint in a lightweight deep-sea cable is made with an overlap of approximately 300 mm. The polyethylene sheath is removed for a distance of 200 mm, and the cotton and aluminium tapes are laid back to expose the copper inner conductor. Electrodes are connected about 100 mm apart on the copper tube, and a large current is applied until the tube and its enclosed steel strands are burnt through. The heat causes the steel strands to become welded together, which ensures that they all take a share of any tensile load applied to the completed joint. When the 2 cable-ends have been thus treated, they are allowed to cool before being cut back and filed to form stubs of circular cross-section, capable of being butted together.



(*b*)

The welded stubs are coated with a layer of silicon-carbide dust to provide extra grip, and are inserted into a 115 mm long steel ferrule. The bore of the ferrule is such that the ends of the copper tubes are just nipped by the ends of the steel ferrule. A press, capable of exerting a force of 2 MN on a 50 mm length of ferrule, is used to swage the ferrule down on to the steel strands; the central section of the ferrule is pressed first, and then the ends. After pressing, the ferrule is approximately 130 mm long.

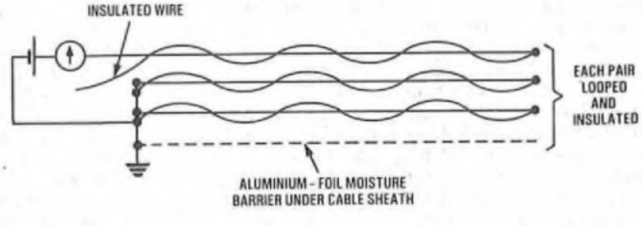
A groove in the steel ferrule accommodates a 0·315 mm square-section copper wire, which is soldered to each copper tube with high-melting-point solder to maintain the continuity of the inner conductor. The joint is illustrated in sketch (*b*).

***Q9** Describe FIVE tests that would be carried out after jointing a 0·8 km section of audio-type paper-insulated cable which does not require balancing.*

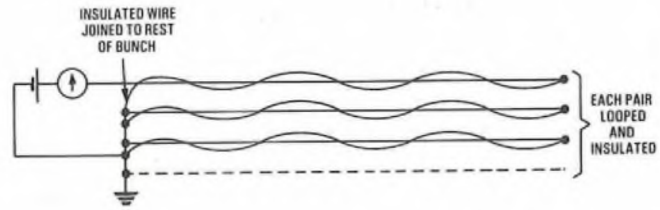
A9 The 5 tests which would be carried out on a 0·8 km section of audio-type paper-insulated cable are:

(*i*) continuity of the conductors,
(*ii*) absence of wire crosses between pairs,
(*iii*) freedom from wire-to-wire and wire-to-earth contacts,
(*iv*) insulation resistance, and
(*v*) absence of air-leakage under pressure.

Tests for the first three conditions can be carried out in one sequence. At the far end of the cable under test, each pair is looped and insulated. At the testing end, all the conductors are connected together with tinned copper wire and earthed to the aluminium-foil moisture barrier



(*a*)



(*b*)

under the cable sheath. One pair is taken at a time from the earthed bunch, and one wire is connected as shown in the sketch (*a*) while the other is insulated. No deflexion on the detector indicates that the pair is clear of wire or earth contacts and crosses. The insulated wire is returned to the earthed bunch as shown in sketch (*b*), and a deflexion on the detector indicates continuity of the pair. The tested pair is then placed on one side and the remaining pairs are tested in a similar manner. If a deflexion larger than normal is noted on any pair, then a contact between the wires of the pair under test is indicated. Any faulty pairs must be reconnected to the earthed bunch before further pairs are tested.

For the insulation test, the conductors at the distant end are insulated from each other and from earth. At the testing end the A-wires are bunched together, as are the B-wires. If a quad cable is involved the C and D-wires are also respectively bunched. The insulation resistance between each group of wires and the remaining groups, joined to the cable earth, is measured using a 500 V ohmmeter. The insulation resistance of the cable per kilometre is obtained by multiplying together the reading of the ohmmeter, the number of wires in the group and the length of the cable.

A pressure test is carried out on each joint to ensure the effectiveness of the joint between the sleeve and the cable sheath. Dried air is passed into the air valve fitted to the jointing sleeve. When a pressure of about 65 kPa has been obtained, the epoxy-resin joints are coated with a bubble-forming solution and carefully examined for any leakage (indicated by the presence of bubbles).

***Q10** (a) Explain what is meant by the fineness modulus of an aggregate used in a concrete mix.*

(b) Describe, in detail, how to measure and tabulate the fineness modulus of an aggregate.

A10 (*a*) The fineness modulus is a method of representing the grading of an aggregate by means of a numerical value. Its significance in the design of concrete mixes is that it enables some measure of control to be exercised over the grading of the aggregate used. The grading, or proportion of the various sizes of particles in an aggregate, is of importance because it affects the workability of the concrete. The maximum suitable size of an aggregate is controlled by the nature of the work; for example, in thin slabs or walls, the largest size of aggregate should not exceed about 0·2–0·25 times the thickness of the concrete. For reinforced-concrete work, the maximum size of coarse aggregate in normal situations is 19 mm, although larger aggregates are frequently used in road construction.

The fineness modulus is a useful means of comparing the gradings of 2 aggregates, so long as they are of a somewhat similar type. Thus, it can be effectively used to record any variation in the grading of aggregates used for a particular application. If, however, the sizes of particles in a sample of aggregate are very dissimilar, the fineness modulus is not a good guide to the concrete-making properties of the aggregate.

(*b*) To measure the fineness modulus of an aggregate, it is first necessary to obtain a representative sample of the aggregate, and the method of *quartering* is best employed. The aggregate is spread out evenly on a clean surface to a depth of about 75 mm. It is then divided into 4 equal parts, and 2 diagonal quarters are discarded. The remainder is mixed and again quartered. This procedure is continued until a sample of about 10 kg remains. The sample is then carefully dried, cooled and weighed, after which it is passed successively through 9 sieves having decreasing aperture sizes. The material retained on each sieve, together with any material cleaned from the mesh, is weighed and recorded, as shown in the table below, which gives the results for a typical 9·6 kg sample. The weight of aggregate passing each sieve, and its relationship to the sample, is then calculated.

The fineness modulus is calculated by adding up the cumulative percentages of the aggregates retained on each sieve and dividing by 100. In the case of the aggregate sample in the table,

$$\text{fineness modulus} = \frac{658}{100} = \underline{6\cdot 58.}$$

Sieve-Mesh Size (mm)	Weight Retained on Sieve (kg)	Weight Passing Sieve (kg)	Percentage Passing Sieve (%)	Percentage Retained on Sieve (%)
38	0	9·6	100	0
19	0·24	9·36	97	3
9·5	5·52	3·84	40	60
4·8	3·36	0·48	5	95
2·4	0·48	—	0	100
1·2	—	—	0	100
0·6	—	—	0	100
0·3	—	—	0	100
0·15	—	—	0	100
			Total:	658

COMPUTERS B 1980

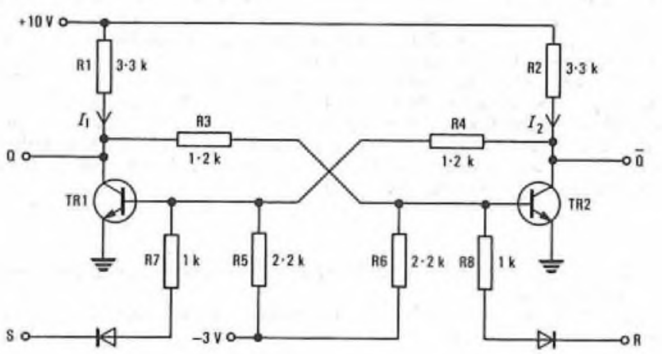
Students were expected to answer any 6 questions

Q1 *(a) Draw a practical circuit diagram of a bistable using n p n transistors and explain how it works. Give typical component values.*

(b) State what is meant by

(i) an astable device,
(ii) a bistable device, and
(iii) a monostable device.

A1 (*a*) A typical bistable circuit is shown in the sketch, with typical component values indicated.



Assume that at the instant of switching on $I_1 = I_2$ and that a fluctuation then causes I_1 to increase by a small amount. The voltage across R1 increases by a corresponding amount, which reduces the base voltage of TR2. This in turn decreases the collector current of TR2 (I_2) which reduces the voltage across R2. The base voltage of TR1 is thus increased, further increasing I_1 and reinforcing the original change. The combined gain of TR1 and TR2 is large and, therefore, this regenerative action proceeds very fast. TR1 becomes saturated (switches to the ON state) and TR2 is driven to the OFF state. This is a stable state, but the circuit could equally have become stable with TR2 switched ON and TR1 switched OFF. Thus, the circuit has two stable states. Each stable state is held by the circuit until some external stimulus causes the state to change. Hence, the circuit has memory.

If TR1 is switched ON and TR2 is switched OFF, a change of state may be initiated by applying a negative pulse to the S input. The negative pulse will turn TR1 OFF and its collector voltage is increased by R1. The increased base voltage at TR2 switches TR2 ON and the collector voltage of TR2 is reduced. This holds TR1 in the OFF state. The bistable circuit has therefore changed state. Applying a negative pulse to input R will cause the circuit to change to the original state.

(*b*) (*i*) An astable device is one which has two quasi-stable states, one following the other. Thus the device alternates between the two states because neither can be held indefinitely. The output of the device is a square wave and the device is therefore a digital oscillator, the frequency of which is determined by the values of the components.

(*ii*) A bistable device has two completely stable states. It can be changed from one state to the other by the application of pulses to its inputs. Thus a bistable device can be used as a digital memory.

(*iii*) A monostable device is asymmetrical, having one stable state and one quasi-stable state. The idle state of the device is obviously the stable state. If a pulse is applied to produce the quasi-stable state, the output changes for a defined length of time and then reverts to the stable state without any further stimulus.

Q2 *(a) Explain the relationships between high level languages, machine code and microprograms.*

(b) What is a compiler?

(c) Give TWO *ways of holding a microprogram in a computer.*

A2 (*a*) A computer performs its various functions by means of a program of instructions. In the form in which they are actually operated on by the computer's processor, these instructions consist of a series of numbers or coded patterns of digits. In this form, the instructions are said to be in machine code.

Programming in machine code is extremely tedious and error prone. The various numerical operation codes have no obvious relationship to their functions. Addresses of store locations used by the program must be carefully noted and their numerical values used when the location is referred to. Any changes to the program which alter the numerical address of any location require thorough checking and alteration of other references to these locations. The set of operation codes is designed for use with the hardware of a specific computer.

High-level languages have been developed to make programming easier by making greater use of the computer. High-level languages are problem oriented, rather than machine oriented. They are designed not with a particular machine code in mind, but to simplify the solution of a particular type of problem.

The instructions in a high-level language take the form of fairly complex statements which must be translated into machine code by special translating programs which have previously been entered into the computer store. The operation code-set comprises words that are in common use in the English language, and normally a single operation code is equivalent to a number of machine operation codes.

High-level language programs are designed to be very flexible and can be used on a variety of different machines. However, the object program is usually longer than a machine-code program as the programmer has less control over the specific machine-code instructions which his program generates.

The concept of microprogramming has been developed to provide a means whereby the programmer may, within limits, design the machine-code instruction-set to suit his requirements. Microprograms translate the machine-code instructions into a greater number of elemental operations to achieve the required effect.

In a microprogrammed machine, the functions used by the programmer are interpreted and performed by an underlying program, the *microprogram*, which uses a very primitive order code implemented in electronic logic. The feature common to all forms of microprogramming is that some method is provided to specify the individual control signals to be actuated by the computer. A micro-operation is an operation controlled by one of these individual control signals. In the most elementary form of microprogramming, the instruction format, particularly the operation part, is so designed that the individual bits in the instruction pertain to individual control signals. In the more general form of microprogramming, the bit combinations in the operation part of each instruction are used to represent a set of instructions.

(*b*) A compiler is a complex program which converts a source program written in a high-level language into machine code. The resulting object program can then be read and executed by the computer. In order to produce the object program, the compiler translates each high-level language statement into its machine code equivalent. It also incorporates into the object program any library subroutines requested by the programmer and supplies the interconnecting links between the parts of the program. A compiler generates several machine-code instructions for each high-level language statement.

The compiler allocates absolute addresses to all data items and program instructions, and provides error detection by rejecting any incorrectly formed statements or any labels which would cause logical errors. Normally, the compiler produces a listing of the source program, together with the resulting object program, and a list of storage areas used and errors detected.

(*c*) Two ways of holding microprograms in a computer are by using a diode matrix, and by software using a READ/WRITE store. In the first case, the program is wired-in, and the control circuits are arranged so that any given bit-combination in the operation part of an instruction initiates a desired set of micro-instructions. The control signals are generated through arrays of diode matrices.

In the second case, microprograms are stored in a storage unit (for example, a core store) and the execution of each instruction is somewhat analogous to the functioning of a subroutine in conventional programming.

Q3 Write a program using a simple machine code of your own choice to perform the following calculation:

$$x = a^2 + b^2 + 4ac.$$

Include a test for a positive or negative result and give a key to the code.

A3 A machine code program to calculate the function

$$x = a^2 + b^2 + 4ac$$

is given below. It is assumed that the values of a, b, c, 4, 1 and 0 have already been loaded into store. The functions of the machine-code instructions are given in Table 1. The contents of the relevant store locations are given in Table 2.

Label	Program Instruction	Comments
	STA	Start the program
	LDA 10	Load the accumulator with a
	MULT 10	Multiply the contents of the accumulator by a
	STO 14	Store a^2 in location 14
	LDA 11	Load the accumulator with b
	MULT 11	Multiply the contents of the accumulator by b
	STO 15	Store b^2 in location 15
	LDA 10	Load the accumulator with a
	MULT 12	Multiply the contents of the accumulator by c
	MULT 13	Multiply the contents of the accumulator by 4
	ADD 14	Add a^2 to the contents of the accumulator
	ADD 15	Add b^2 to the contents of the accumulator
	STO 19	Store the result, x, in location 19
	JGT L1	If the contents of the accumulator are positive jump to label L1
	LDA 17	Load the accumulator with 0
	JMP L2	Jump to label L2
L1	LDA 16	Load the accumulator with 1
L2	STO 18	Store the contents of the accumulator in location 18 (Flag)
	STP	Stop the program

The sign of the result, x, can now be determined by examining the contents of store location 18.

Table 1

Operation code	Function
STA	Start the program
LDA Z	Load the accumulator with the contents of location Z
MULT Z	Multiply the contents of the accumulator by the contents of location Z
STO Z	Store the contents of the accumulator in location Z
ADD Z	Add the contents of location Z to the contents of the accumulator
JGT L	Jump to label L if the contents of the accumulator are positive
JMP L	Jump to label L
STP	Stop the program

Table 2

Store location	Contents
10	a
11	b
12	c
13	4
14	a^2
15	b^2
16	1
17	0
18	Flag
19	x

Q4 (a) Draw the truth table for a binary full adder.

(b) Draw a logic diagram of a binary full adder using NAND *elements only.*

(c) Explain ONE *way of performing a subtract function using only adder logic.*

A4 (*a*) A full adder is a device that adds together two binary numbers, one digit at a time, together with the carry digit from the previous stage of addition. The output is a SUM digit and a CARRY digit. In the truth table below, A and B are the two binary numbers to be added, C is the CARRY digit from the previous stage, S is the SUM output and C_0 is the CARRY output.

A	B	C	S	C_0
0	0	0	0	0
0	0	1	1	0
0	1	0	1	0
0	1	1	0	1
1	0	0	1	0
1	0	1	0	1
1	1	0	0	1
1	1	1	1	1

(*b*) From the truth table given in part (*a*), the Boolean expressions for the sum and carry outputs are

$$S = \bar{A}.\bar{B}.C + \bar{A}.B.\bar{C} + A.\bar{B}.\bar{C} + A.B.C,$$

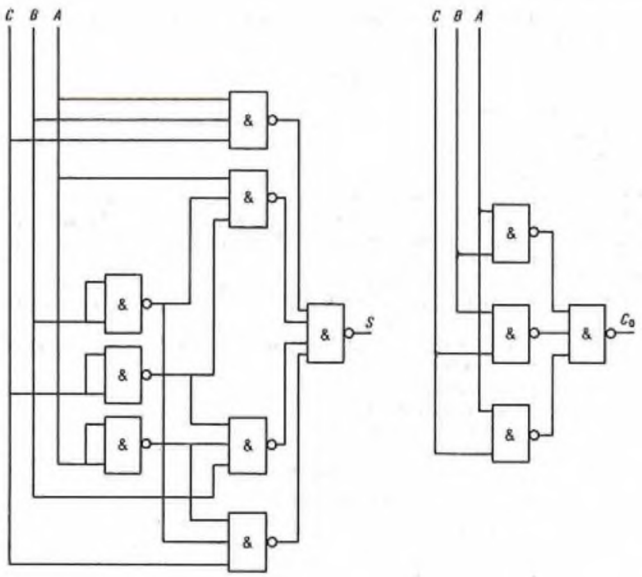
$$= \overline{(\overline{\bar{A}.\bar{B}.C}).(\overline{\bar{A}.B.\bar{C}}).(\overline{A.\bar{B}.\bar{C}}).(\overline{A.B.C})}. \text{ (By de Morgan's theorem)}$$

$$C_0 = B.C + A.C + A.B,$$

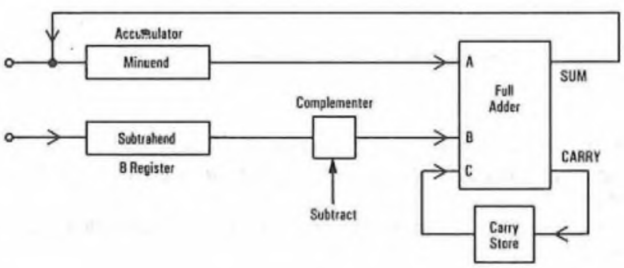
$$= \overline{(\overline{B.C}).(\overline{A.C}).(\overline{A.B})}.$$

A logic diagram of a binary full adder using NAND elements only is shown in sketch (*a*).

(*c*) In subtraction, the subtrahend is subtracted from the minuend to give the difference. Subtraction can be performed by adding the two's complement of the subtrahend to the minuend (see A7 of this paper). In this way it is possible to use the existing adder logic in the computer in conjunction with a complementer which automatically finds the complement of a binary number. The arrangement of the arithmetic unit for serial subtraction by complementary addition is shown in sketch (*b*).



(*a*)



(*b*)

Initially, the minuend is placed in the accumulator and the subtrahend in the B register. The SUBTRACT signal from the control unit switches the complementer into circuit. Before any arithmetic is performed, the carry digit is initially set to 1 in order to add 1 to the one's complement of the subtrahend to produce the two's complement. The digits of the two operands are then shifted sequentially into the full adder. The one's complement of the subtrahend arrives at the B input of the full adder. After the required number of shift pulses have been applied, the difference will be contained in the accumulator in two's-complement form.

Q5 (a) Draw a block diagram of a typical modern digital computer showing the main registers and their interconnexions.

(b) Using the diagram drawn for part (a) explain the sequence of operations and data flow for a non-equivalent operation.

Q6 (a) Draw a diagram showing how six bistable circuits could be connected together to form a six-bit shift register. Use a timing diagram to explain its operation. Explain TWO *methods by which the shift register may be loaded.*

(b) Explain how a shift register could be used to convert serial data to parallel.

Q7 (a) State what is meant by the term "two's complement" and explain why it is used in digital computers.

(b) Using two's complement arithmetic, perform the following calculations in binary and convert the result to denary. Show all working.

(i) $56_{10} - 40_{10}$,
(ii) $12_{10} - 76_{10}$, *and*
(iii) $43_8 - 24_8$.

A7 (*a*) The two's complement of a binary number is that number which, when added to the original number, will result in an all zeros answer and a carry from the left-most bit. The two's complement is obtained by finding the one's complement and adding 1; the one's complement is obtained by inverting each bit of the original number.

Two's complement arithmetic is used in digital computers because it simplifies the process of subtraction. The two's complement of a number is a convention used for representing negative numbers. During calculations using two's-complement working, it is not necessary to keep check of the sign of partial results; if the answer is negative, its two's complement is taken as the magnitude.

In two's complement, each number is assigned a sign bit: 0 for positive and 1 for negative. To obtain a negative number, the positive number, including the sign bit, is two's complemented. In two's-complement arithmetic, subtraction is performed by addition.

(*b*) (*i*) Converting the denary numbers to binary by successive division by 2 and noting the remainder gives

Quotient	Remainder
2)56	
28	0
14	0
7	0
3	1
1	1
0	1

Quotient	Remainder
2)40	
20	0
10	0
5	0
2	1
1	0
0	1

The binary number is obtained by writing down the remainder digits in reverse order.

$\therefore \quad 56_{10} = 111\,000_2$ and $40_{10} = 101\,000_2$.

The subtraction is performed using two's complement arithmetic.

	Sign bit	
40_{10}	0	101 000
One's complement of 40_{10}	1	010 111
Two's complement of 40_{10}	1	011 000
56_{10}	0	111 000+
	0	010 000

$$\therefore \quad 56_{10} - 40_{10} = +010000_2,$$
$$= 1 \times 2^4,$$
$$= \underline{16_{10}}.$$

(*ii*) Converting 12_{10} and 76_{10} to binary in a similar manner to part (*i*), gives

$$12_{10} = 0\,001\,100_2 \text{ and } 76_{10} = 1\,001\,100_2.$$

Subtraction by two's complement arithmetic:

	Sign bit	
76_{10}	0	1 001 100
One's complement of 76_{10}	1	0 110 011
Two's complement of 76_{10}	1	0 110 100
12_{10}	0	0 001 100+
	1	1 000 000

The fact that the sign bit is a 1 indicates a negative number and, hence, the result must be two's complemented again.

$$\therefore \quad 12_{10} - 76_{10} = -1\,000\,000_2,$$
$$= -1 \times 2^6,$$
$$= \underline{-64_{10}}.$$

(*iii*) Octal numbers may be converted to binary by successively converting each octal digit to a binary number. Each octal digit converts to 3 binary numbers.

$\therefore \quad 43_8 = 100\,011_2$ and $24_8 = 010\,100_2$.

	Sign bit	
24_8	0	010 100
One's complement of 24_8	1	101 011
Two's complement of 24_8	1	101 100
43_8	0	100 011 +
	0	001 111

$$\therefore \quad 43_8 - 24_8 = +001\ 111_2,$$
$$= 1 \times 2^0 + 1 \times 2^1 + 1 \times 2^2 + 1 \times 2^3,$$
$$= \underline{15_{10}}.$$

Q8 (*a*) *Explain the following terms:*

(*i*) *transfer rate,*
(*ii*) *access time, and*
(*iii*) *packing density.*

(*b*) *Compare the characteristics of magnetic tape and disc storage with reference to part* (*a*).

(*c*) *For what application would the following storage methods be used:*

(*i*) *magnetic tape, and*
(*ii*) *disc?*

A8 (*a*) (*i*) The transfer rate of a memory device is the rate at which it is capable of transferring data from its memory to another device and/or reading data from another device.

(*ii*) Access time is defined as the delay between the instant when a request is made for the contents of a memory location and the instant when data appears at the output of the memory device.

(*iii*) Packing density is defined as the quantity of information which can be stored per unit size which the storage medium occupies.

(*b*) The transfer rate of information to and from magnetic tape is dependent on the speed of movement of the tape relative to the READ/WRITE head. Typically, a slow moving tape runs at 635 mm/s, which creates a transfer rate of 40 kbytes/s when information is stored at 63 bit/mm. Faster tapes move at 5·08 m/s; that is, a transfer rate of 320 kbytes/s.

The transfer rate of disc storage again depends on the speed at which the disc passes the READ/WRITE head. A rate of 400 kbytes/s is typical. A typical recording density on magnetic discs is 160 bit/mm on any one track, with a track density of up to 160 tracks/cm.

Information is stored on magnetic tape linearly along its length. Consequently, tape is a slow medium when a particular piece of information is to be retrieved, because the tape must be wound to the required position. Access time can be anything from a few seconds, if the tape is by chance positioned exactly as required, to several minutes if the tape has first to be located, loaded onto the tape-deck mechanism and then wound to the required position.

Information is stored on magnetic disc on a series of concentric tracks. An actuating arm moves a READ/WRITE head to the desired track on the rotating disc. In this form of storage, the disc can easily be removed in a protective plastic pack. The pack and its disc, known collectively as a *cartridge*, can be replaced in the disc unit when required. Access time is dependent on the mechanical movement of the head to the required track, and the waiting time until the continously revolving disc is in the appropriate position relative to the head. Typical head-movement delay is from zero (if the head is positioned as required) to 60 ms, with an average delay of 20 ms due to the rotation of the disc.

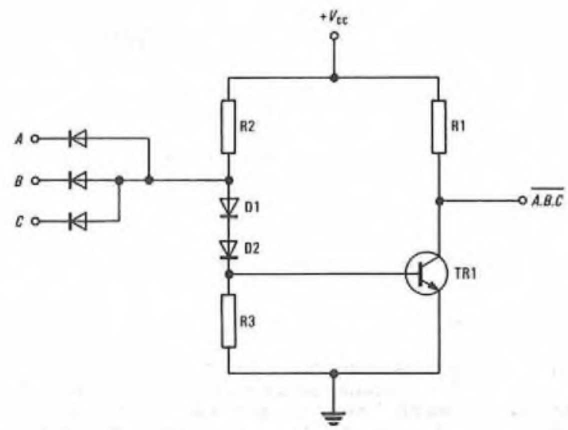
(*c*) (*i*) Magnetic tape is used for the semi-permanent recording of bulk information; that is, information that can be erased when no longer required. A typical example is the retention of historical files of information such as personnel records and monetary transactions. In telecommunications, magnetic tape is used to record the charging of calls in stored-program controlled switching systems. Tapes are periodically removed so that information stored on them can be processed to produce statistics and customers' bills.

(*ii*) The ease with which exchangeable magnetic discs can be removed and replaced makes them an ideal medium for increasing the security of a computer system. The disc, containing a copy of vital information about the state of the system, is periodically removed and replaced by another disc. Should the information be lost or corrupted within the system, it is possible to replace the disc and renew the information. Exchangeable discs are also used as a convenient means of transferring software (programs and data) to a system, particularly during the starting-up of the system.

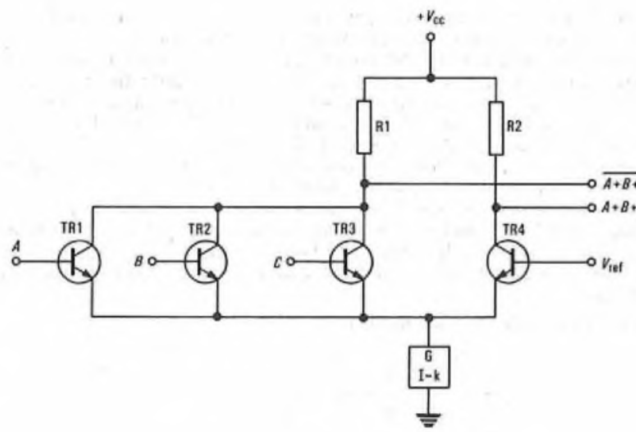
Q9 (*a*) *Compare with respect to speed and fan-out the following logic systems DTL, ECL and TTL. What factors must be taken into consideration when calculating the fan-out for* EACH *system?*

(*b*) *Draw a circuit diagram of a 3-input logic TTL* NAND *element and explain its operation.*

A9 (*a*) Sketch (*a*) shows a typical diode-transistor logic (DTL) circuit. The speed of the circuit is low, a typical propagation delay being 30 ns. Typical fan-out for DTL is 10, the same as for transistor-transistor logic (TTL).



(*a*)



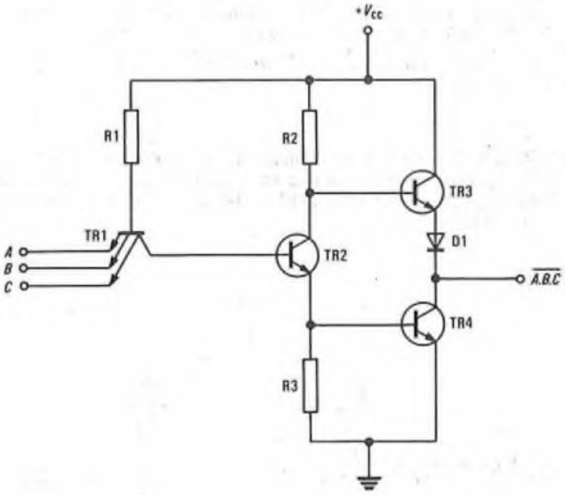
(*b*)

Sketch (*b*) shows a typical emitter-coupled logic (ECL) circuit. The switching speed of this circuit is very fast (propagation delays in the order of 1 ns are possible). Fan-out is higher than for TTL and DTL, 25 being a typical value.

Sketch (*c*) shows a typical TTL circuit. The switching speed is moderately fast (typical propagation delay is 10 ns). Typical fan-out is 10.

When calculating the fan-out of DTL, it can be seen that the fan-out is limited by the ability of the transistor to sink the whole of the current which would otherwise flow through D1 and D2 of every circuit which it is required to drive. The fan-out is therefore dependent on the output resistance of TR1 in the saturated state, the magnitude of the current through R2 and the voltage to which the output can be allowed to rise without it becoming a logic 1. All these factors must be subjected to worst-case tolerancing and allowances must be made for temperature drift and ageing.

When calculating the fan-out of ECL, it can be seen that the fan-out is limited by the ability of the output resistor (R1 or R2) to supply sufficient base current to every input which it is driving when the output is in the logic 1 state. Fan-out is therefore dependent on the collector-resistor value (R1 or R2), the base current required to take the current from the other half of the long-tailed pair and the voltage to which the output may sink without the output becoming a logic 0. All these factors must again be subjected to worst-case tolerancing



(*c*)

and allowances made for temperature drift and ageing.

The fan-out of TTL is limited by the ability of the transistor TR4 to sink the current from all the inputs connected to it when the output is in the low state. Therefore, the fan-out is dependent on the output resistance of TR4 when it is turned on and by the current produced by the emitters of all the input transistors and the value of R1. As in the previous cases worst-case tolerancing, temperature drift and ageing must be considered.

(*b*) Sketch (*c*) shows a typical 3-input TTL NAND gate. TR1 is a multi-emitter transistor which is switched ON if any of the base-emitter junctions is switched ON. When all inputs are at logic 1, the base-emitter junctions are all reverse biased. In this state the transistor behaves as two separate diodes, with current flowing through the forward-biased base-collector junction with TR2 switched ON. The current flow through the collector resistor of TR2 causes TR3 to be switched OFF. Emitter current from TR2 raises the base potential of TR4, which switches ON. The output is at logic 0 (low).

If A or B or C is taken to logic state 0, TR1 switches on and the base voltage of TR2 is held low. TR2 switches OFF. TR4 is thus deprived of base current and switches OFF. TR3 base potential is pulled up by the collector resistor of TR2, and TR3 switches ON. The output state is then HIGH.

The truth table for the circuit is:

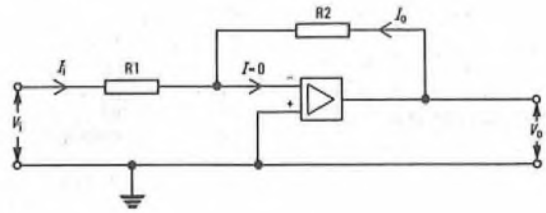
A	B	C	Output
0	0	0	1
0	0	1	1
0	1	0	1
0	1	1	1
1	0	0	1
1	0	1	1
1	1	0	1
1	1	1	0

Q10 *(a) Describe, with the aid of a suitable diagram, how the following may be obtained using a high-gain DC amplifier:*

(i) multiplication by a constant coefficient, and
(ii) integration.

(b) Give THREE *major sources of error in analogue computing.*

A10 (*a*) (*i*) A circuit for multiplication by a constant coefficient is shown in sketch (*a*). The circuit contains an operational amplifier together with input and feedback resistors R1 and R2 respectively. An ideal operational amplifier has infinite gain and takes no current; that is, $I = 0$. Therefore, providing the output voltage, V_o, of the amplifier is of a reasonably finite value, point S is virtually at earth potential. This type of amplifier is often referred to as a *virtual-earth amplifier*. The virtual-earth concept is applied to the circuit in the following derivation, in which V_i is the input voltage, I_i is the input current and I_o is the output current.



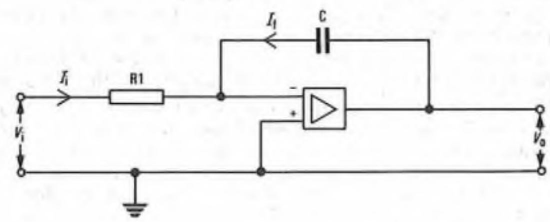
(*a*)

$$V_i = I_i R_1. \qquad \ldots\ldots (1)$$
$$I_i = -I_o$$
$$V_o = I_o R_2,$$
$$= -I_i R_2. \qquad \ldots\ldots (2)$$

Dividing equation (2) by equation (1) gives

$$\frac{V_o}{V_i} = -\frac{I_i R_2}{I_i R_1},$$
$$= -\frac{R_2}{R_1}.$$
$$\therefore \quad V_o = -\frac{R_2}{R_1} V_i.$$

By selecting values of resistors in the required ratio, the input voltage can be multiplied by any desired coefficient.



(*b*)

(*ii*) An operational amplifier connected as an integrator is shown in sketch (*b*). The feedback component is a capacitor, so that current flows only when the capacitor is charging or discharging. As the capacitor is charging, the current flow is large initially, and reduces exponentially. However, the reversing property of an operational amplifier has the effect of keeping the rate of charge constant, thus maintaining a constant-current flow in the capacitor. (As the voltage applied to the plate on the input side of the capacitor increases, so the opposite voltage at the output of the operational amplifier is applied to the other plate, thus keeping the rate of charge constant.)

The operation can be explained mathematically as follows.

If q is the charge on the capacitor at any instant, then the feedback current is

$$I_f = \frac{dq}{dt}.$$

But $q = V_o C$, assuming the virtual-earth principle.

$$\therefore \quad I_f = C\frac{dV_o}{dt}.$$

By Kirchhoff's first law,

$$I = -I_f.$$
$$\therefore \quad \frac{V_i}{R_i} = -C\frac{dV_o}{dt}.$$
$$\therefore \quad \frac{dV_o}{dt} = -\frac{V_i}{CR_i}. \qquad \ldots\ldots (3)$$

The value of V_o at any instant is found by integrating equation (3) with respect to time.

Thus, $$V_o = -\frac{1}{CR_i}\int V_i\, dt.$$

This shows that the output voltage is proportional to the time integral of the input voltage.

(*b*) Three major sources of error in analogue computing are due to component tolerance, drift and noise.

The accuracy of analogue computing is limited partly by the accuracy of the values of the components, which can be made only within certain tolerances. In particular, resistors and capacitors must have values with very small variations resulting from manufacturing processes.

Drift is a DC phenomenon which causes a DC output when both inputs are zero. This varies with time and temperature. It is a particular problem with integrators, where it manifests itself as an error which increases with time.

Noise may limit accuracy when very low-level signals are used. Noise is generated by components which have resistance and is amplified by successive stages of the analogue computer. The total noise produced depends on the noise generated by the amplifier, the source resistance and the bandwidth.

TECHNICIAN EDUCATION COUNCIL

Certificate Programme in Telecommunications

Sets of model questions and answers for TEC units are given below. They have been designed following analysis of assessment test papers actually set during the 1978–79 session by a number of colleges all over the country. The model questions and answers reflect the types and standard of question set and answer expected, and include the styles of both in-course and end-of-unit assessments.

The model questions and answers therefore illustrate the assessment procedures that students will encounter, and are useful as practice material for the skills learned during the course.

The use of calculators is permitted except where otherwise indicated.

Representative time limits or proportion of marks are shown for each question (or group of questions), and care has been taken to give model answers that reflect these limits. Where additional text is given for educational purposes, it is shown within square brackets [ ] to distinguish it from the information expected of students under examination conditions.

We would like to emphasize that, because the model questions are based on work at a number of colleges, they are not representative of questions set by any particular college.

As a general rule, questions are given in italic type and answers in upright type. Answers are sometimes shown in bold upright type; this is because, for some objective questions, it is convenient to place the questions and answers side by side, and bold type enhances the distinction in such cases. Where possible, answers have been positioned such that they may be covered up if desired.

LINES 2 1979–80

Students are advised to read the notes above

Note: Following suggestions made by students, the layout of the questions and answers for this subject is different from that used previously. For the short-answer questions, the answers have been printed at the end of the set of questions. In this way, students may, if they wish, use the questions as self-assessment questions without being distracted by the answers. The editors would be pleased to receive comments on this form of layout, which is experimental.

Questions (for answers see p. 28)

Q1 *Define the term "system".* (2 marks)

Q2 *List the corporate objectives of a typical national telecommunications system.* (3 marks)

Q3 *Define the term "gap".* (2 marks)

Q4 *Draw a simple schematic diagram to show the relationship of a telephone system with a telecommunications system.* (2 marks)

Q5 *Explain how the system parameter "quality of transmission" may be measured.* (2 marks)

Q6 *The net assets (fixed assets less depreciation at historic cost) for a telecommunications business are £1200M at the end of a financial year. Calculate the return on capital as a percentage (ROC%) if the return (after the addition of interest and supplementary depreciation) is £100M.* (5 marks)

Q7 *What are the disadvantages of the manual (operator) method of interconnecting links compared with the automatic method?* (3 marks)

Q8 *Draw a straight-line diagram to illustrate a national transmission plan.* (5 marks)

Q9 *Explain the basic differences between the local-lines and junction-cable forecasting methods.* (2 marks)

Q10 *Define the term "rolling forecast".* (4 marks)

Q11 *If the future number of tenancies is 65 and the future penetration factor is 0·6, what is the value of the future number of connexions?* (2 marks)

Q12 *List FIVE reasons why it is difficult to obtain accurate forecasts.* (5 marks)

Q13 *Define the term "inductance".* (3 marks)

Q14 *Draw a network representing a short length of uniform transmission line. Show four resistances each of value R/4 and four inductances each of value L/4.* (5 marks)

Q15 *If the length of a cable is increased by additional cable, what effect will this have on the measured value of the conductance?* (2 marks)

Q16 *Define the term "characteristic impedance".* (3 marks)

Q17 *What effect has lumped loading on the bandwidth of a transmission line?* (2 marks)

Q18 *Why is capacitance unbalance reduced in value in star-quad cables?* (2 marks)

Q19 *Why is it necessary to ensure that poles of an adequate strength are used when stays are applied?* (2 marks)

Q20 *If the pull on a pole is 500 N and the height of the pole above the ground is 10 m, what is the value of the bending moment if no stay is applied?* (4 marks)

Q21 *Find the value of the wire tension in newtons where a suspended wire of 65 m span length weighs 0·560 N/m and has a dip of 450 mm at the centre of the span.* (4 marks)

Answers to questions on p. 27.

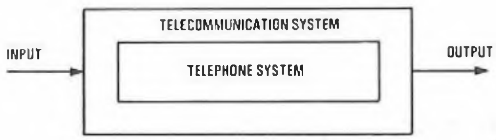
A1 A system is defined as a set of component parts working together to achieve an objective.

A2 The corporate objectives of a typical telecommunications system are:
(*a*) provision of a sound and efficiently engineered system,
(*b*) achievement of a suitable return on the capital employed, and
(*c*) achievement of good staff relations with employees.

A3 *Gap* is defined as the difference between the Board's corporate target (F_T) and the departmental forecasts (F_O) at an early stage; that is,

$$Gap = F_T - F_O.$$

A4 The schematic diagram is shown in the sketch.

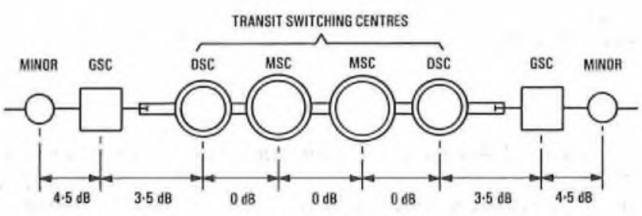


A5 Quality of transmission is measured by determining the proportion, as a percentage or a fraction, of the number of speech components correctly received over the system to the number transmitted. The type of speech component employed is known as *logatom articulation;* that is, meaningless syllables spoken in a random sequence.

A6
$$\text{Return on capital} = \frac{\text{Return}}{\text{Net assets}} \times 100\%,$$
$$= \frac{100 \times 10^6}{1200 \times 10^6} \times 100\%,$$
$$= \underline{8 \cdot 3\%}.$$

A7 The disadvantages of a manual method of interconnecting links compared with an automatic system are:
(*a*) it is a labour intensive system,
(*b*) there is a need to provide staff for 24 h/d, and
(*c*) it takes longer to switch calls.

A8 A typical transmission plan is shown in the sketch.



GSC: group switching centre
DSC: district switching centre
MSC: main switching centre

A9 Local-line forecasting is based on recording and projecting the number of *customers*, whereas junction-cable forecasting depends on recording and projecting *telecommunications traffic* figures.

A10 Rolling forecasting is a technique covering a selected length of time (say 20 years) and having fixed and variable bases with provision for regularly updating the forecast information as the calendar coverage increases.

A11 Number of connexions = penetration factor × number of tenancies,

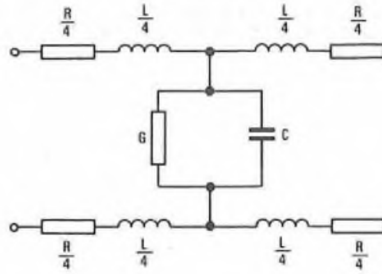
$$= 0 \cdot 6 \times 65 = \underline{39}.$$

A12 Any FIVE of the following:

(*a*) Last minute changes in the use of planned development and re-development.
(*b*) Unforeseen changes in the locations of building sites because of technical difficulties.
(*c*) The unexpected closures of leasehold property.
(*d*) Companies going bankrupt and into liquidation.
(*e*) Unexpected resistance by customers.
(*f*) Damage by fires and storms.
(*g*) Unforeseen national and international economic crises.
(*h*) Unforeseen political or technical developments.

A13 Inductance is defined as that property of an element or circuit which when carrying a current is characterized by the formation of a magnetic field and the storage of magnetic energy.

A14 The network is shown in the sketch.



A15 The value of the conductance will be increased.

A16 The characteristic impedance of a uniform transmission line is the impedance measured at the sending-end terminals of the line when the line is infinitely long or is terminated in an impedance equal to the characteristic impedance.

A17 Lumped loading lowers the cut-off frequency and causes the line to act like a low-pass filter.

A18 Because of the symmetry of the conductors.

A19 To prevent the pole buckling because of the downward thrust.

A20 The bending moment = pull on the pole × height above ground,

$$= 500 \times 10 = \underline{5000 \text{ N m}}.$$

A21
$$T = \frac{WL^2}{8d} \text{ newtons},$$

where T is the tension in newtons, W is the weight of the wire in newtons/metre, and d is the dip at the centre of the span in metres.

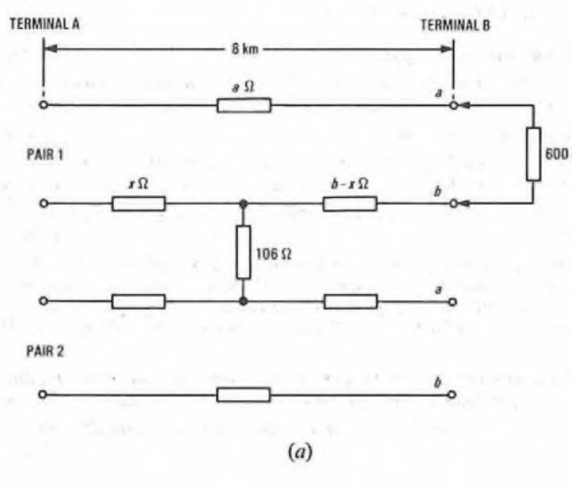
$$\therefore \quad T = \frac{0 \cdot 560 \times 65^2}{8 \times 450 \times 10^{-3}},$$
$$= \underline{657 \text{ N}}.$$

Q22 A 106 Ω contact fault is discovered between 2 cable pairs (pairs 1 and 2) in the same cable. During Varley-type fault-location measurements at terminal A, a 600 Ω loop is inadvertently applied at terminal B.

The measured loop reading is 1048·2 Ω with 1:1 ratio arms. Three-wire tests show that there is 0·0 Ω resistance unbalance between the pair conductors.

Calculate the distance to the fault from terminal B if the cable distance between A and B is 8 km, and the value of the variable resistance at Varley balance is 182·7 Ω with the bridge ratio arms set at 3:1. (*20 marks*)

A22 The line conditions may be represented as shown in sketch (*a*).



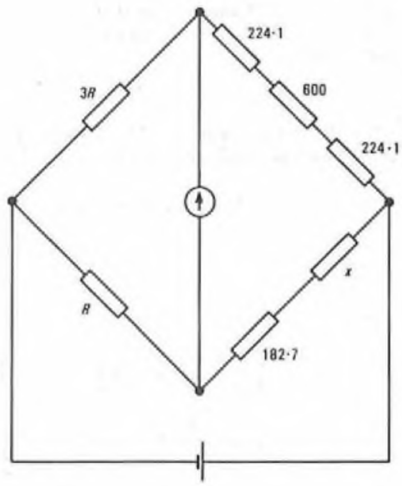
(*a*)

If the loop conductor resistance reading of pair 1 = 1048·2 Ω, and the additional resistance = 600·0 Ω, the actual loop conductor resistance = 448·2 Ω.

A 3-wire test showed that $a = b = \dfrac{448 \cdot 2}{2} = 224 \cdot 1\ \Omega$.

$$\therefore \quad b - x = 224 \cdot 1 - x\ \Omega.$$

The bridge conditions at Varley balance can be represented by sketch (*b*).



(*b*)

At balance,

$$3R(182 \cdot 7 + x) = R\{224 \cdot 1 + 600 + (224 \cdot 1 - x)\}.$$
$$\therefore \quad 3(182 \cdot 7 + x) = 224 \cdot 1 + 600 + (224 \cdot 1 - x).$$
$$\therefore \quad 548 \cdot 1 + 3x = 1048 \cdot 2 - x.$$
$$\therefore \quad 4x = 1048 \cdot 2 - 548 \cdot 1.$$
$$\therefore \quad x = \frac{500 \cdot 1}{4}.$$
$$= 125 \cdot 0\ \Omega.$$

Therefore the distance to the fault from terminal A

$$= \frac{125 \cdot 0}{224 \cdot 1} \times 8 \cdot 0,$$
$$= 4 \cdot 462\ \text{km}.$$

Therefore the distance to the fault from terminal B

$$= 8 - 4 \cdot 462,$$
$$= \underline{3 \cdot 538\ \text{km}}.$$

Q23 (*a*) *List 5 objectives of systems comparison.*

(*b*) *Three simple local telephone systems are each operated by small companies (A, B and C).*

Company A has modern automatic switching plant, but a restricted amount of out-of-date line plant. The quality-of-switching-and-signalling factor (QSF) is 40·0, but the time-to-install-plant factor (TPF) is only 5·0 and the quality-of-maintenance factor (QMF) is 30·0.

Company B has a restricted amount of out-of-date automatic switching plant but a sufficient amount of modern line plant. The QSF is 5·0, but the TPF is 80·0 and the QMF is 30·0.

Company C has sufficient automatic switching and line plant, but a high fault liability on line and switching plant. The QSF is 10·0, the TPF is 80·0 and the QMF is 5·0.

If the respective and agreed weighting factors are 0·3, 0·1 and 0·1 for QSF, TPF and QMF, and all the other performance and weighting factors are the same for each company, which company has the best overall system performance?

(*c*) *If company A is to maintain its overall system performance, what action is required at an early date?*

(*d*) *If, at this stage, the department forecasts (F_O) do not allow for the action determined in part (c), how will this be revealed by the Company Board?* (*25 marks*)

A23 (*a*) The objects of carrying out system comparisons are

(*i*) to assist in determining the value of performance targets and how system performance may be improved,
(*ii*) to monitor the performance of business competitors,
(*iii*) to allow analysis of different operational and management techniques,
(*iv*) to illustrate how the contribution of technicians in the field can lead to greater efficiency and productivity, and
(*v*) to assist in the establishment of standard performances.

(*b*) As each company has the same net-asset value, the overall system performance can be calculated as a percentage from the given weighting factors as follows:

Company A

QSF = 40·0; weighting factor = 0·3.
∴ weighted QSF = 12·0.
TPF = 5·0; weighting factor = 0·1.
∴ weighted TPF = 0·5.
QMF = 30·0; weighting factor = 0·1.
∴ weighted QMF = 3·0.

If the value of the remaining performance factors is x (for all companies), the overall system performance is x + 15·5%.

Company B

QSF = 5·0; weighting factor = 0·3.
∴ weighted QSF = 1·5.
TPF = 80·0; weighting factor = 0·1.
∴ weighted TPF = 8·0.
QMF = 30·0; weighting factor = 0·1.
∴ weighted QMF = 3·0.

Therefore, the overall system performance is x + 12·5%.

Company C

QSF = 10·0; weighting factor = 0·3.
∴ weighted QSF = 3·0.
TPF = 80·0; weighting factor = 0·1.
∴ weighted TPF = 8·0.
QMF = 5·0; weighting factor = 0·1.
∴ weighted QMF = 0·5.

Therefore, the overall system performance is x + 11·5%.

Therefore, company A has the best overall system performance at the date of measurement.

(*c*) It is necessary for company A to provide sufficient, and up-to-date line plant at an early date if it is to maintain its overall system performance.

(*d*) This will be revealed by the company's corporate planning process. The difference between the Board's corporate targets (F_T) and the departmental forecasts (F_O) will be revealed by the gap.

$$F_T - F_O = \text{gap}.$$

Q24 (*a*) *Define what is meant by the term cable.*

(*b*) *List the line primary coefficients.*

(*c*) *Draw a T network to represent a short length of uniform transmission line in terms of its primary coefficients.*

(*d*) *Primary-coefficient measurements are made on various lengths of one-pair cable (copper conductor, paper and air insulated with polyethylene sheath) which are uniformly manufactured. Assuming that there are no conductor-resistance unbalances, calculate the primary coefficient values per kilometre, if all measurements are made at the same cable temperature, from the following data:*

(*i*) *mutual capacitance = 0·01 μF absolute on a 100 m length,*
(*ii*) *single-wire resistance = 15 Ω absolute on a 545 m length,*
(*iii*) *inductance = 0·1 mH absolute on a 100 m length, and*
(*iv*) *insulation resistance = 2 kΩ absolute on a 500 m length.*

(*e*) *Define the term characteristic impedance (Z_0).*

(*f*) *Given that the cable pair in part (d) has primary and secondary coefficients uniformly distributed and that the characteristic impedance can be expressed as follows,*

$$Z_0 = \sqrt{\frac{\textit{series impedance of line per unit length}}{\textit{shunt admittance of line per unit length}}},$$

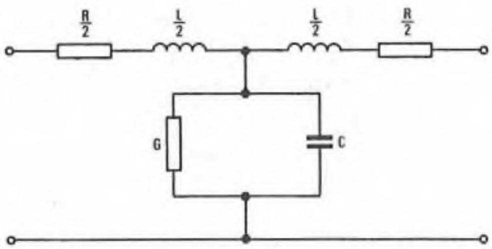
calculate the value of the characteristic impedance using the values of the primary coefficients per kilometre if ω = 5000 rad/s. (*25 marks*)

A24 (*a*) A cable is defined as an assembly of conductors insulated from one another and enclosed in a common binding or sheathing, and having some degree of flexibility.

(*b*) The primary coefficients are:

(*i*) resistance (R),
(*ii*) inductance (L),
(*iii*) conductance (G), and
(*iv*) capacitance (C).

(*c*) The network representing a short length of transmission line is shown in the sketch.



(*d*) (*i*) Capacitance per kilometre = $0{\cdot}01 \times 10 = \underline{0{\cdot}1\ \mu\text{F}}$.

(*ii*) Resistance per kilometre = $30 \times \dfrac{1000}{545} = \underline{55{\cdot}05\ \Omega}$.

(*iii*) Inductance per kilometre = $0{\cdot}1 \times 10 = \underline{1{\cdot}0\ \text{mH}}$.

(*iv*) Insulation resistance per kilometre = $2000 \times \dfrac{500}{1000} = 1000\ \Omega$.

$\therefore$ conductance per kilometre = $\dfrac{1}{1000} = \underline{1 \times 10^{-3}\ \text{S}}$.

(*e*) The characteristic impedance of a uniform transmission line is defined as the impedance measured at the sending end terminals of the line when the line is infinitely long or is terminated in an impedance equal to the characteristic impedance.

(*f*)

$$Z_0 = \sqrt{\left\{\frac{\sqrt{(R^2+\omega^2L^2)}\angle \tan^{-1}\dfrac{\omega L}{R}}{\sqrt{(G^2+\omega^2C^2)}\angle \tan^{-1}\dfrac{\omega C}{G}}\right\}},$$

$$= \sqrt{\left\{\sqrt{\left(\frac{R^2+\omega^2L^2}{G^2+\omega^2C^2}\right)}\angle\left(\tan^{-1}\frac{\omega L}{R} - \tan^{-1}\frac{\omega C}{G}\right)\right\}},$$

$$= \sqrt[4]{\left(\frac{R^2+\omega^2L^2}{G^2+\omega^2C^2}\right)}\angle \tfrac{1}{2}\left(\tan^{-1}\frac{\omega L}{R} - \tan^{-1}\frac{\omega C}{G}\right),$$

$$= \sqrt[4]{\left(\frac{55{\cdot}05^2 + 25\times10^6\times10^{-6}}{10^{-6}+25\times10^6\times10^{-12}\times0{\cdot}01}\right)}\angle\tfrac{1}{2}$$

$$\left(\tan^{-1}\frac{5\times10^3\times10^{-3}}{55{\cdot}05} - \tan^{-1}\frac{5\times10^3\times10^{-7}}{10^{-3}}\right),$$

$$= \sqrt[4]{\left(\frac{3055{\cdot}5\times10^6}{1{\cdot}25}\right)}\angle\tfrac{1}{2}(5^\circ\,11' - 26^\circ\,34'),$$

$$= 22{\cdot}24\times10\angle -10^\circ\,42',$$

$$= \underline{222{\cdot}4\angle -10^\circ\,42'\ \Omega}.$$

Q25 *(a) Define the term critical path.*

(b) You are given the task of controlling the time spent on the provision of telecommunications plant facilities at a large new housing estate. The planning work starts with the receipt of site plans, a map survey in the office and then a meeting with the housing architect to agree on the facilities required and the type of distribution.

At present, the site footways and roads are unmade, but surfacing work commences on day No. 61.

From the list of activities given and estimated times for completion, provide the following information:

(i) show by means of a suitable flow diagram the network of activities in the correct sequence of events,
(ii) list the activities included in the critical path,
(iii) calculate the latest date for end event,
(iv) calculate the earliest date for end event, and
(v) if the necessary wayleaves cannot be obtained in less than 39 d can the project be completed before surfacing work commences?

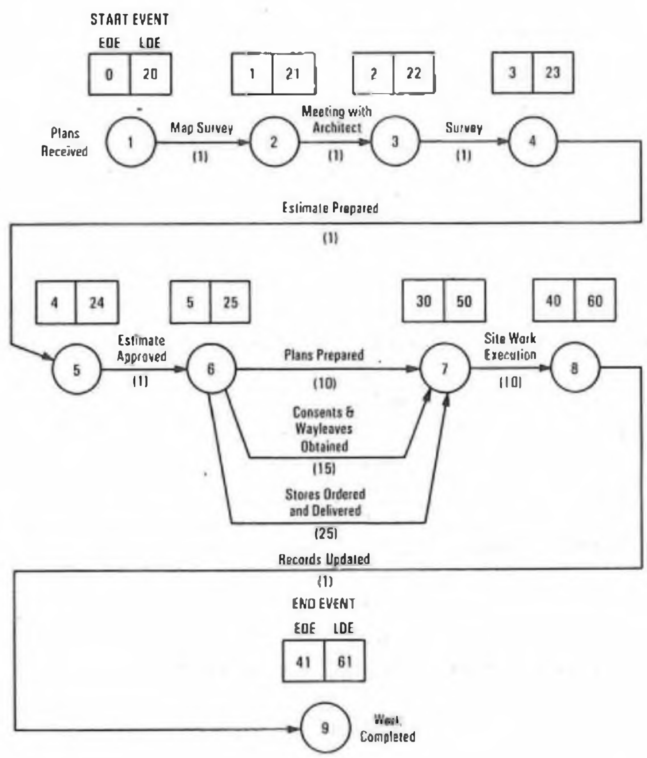
List of activities

Estimate approved (1 d)	*Plant work executed (10 d)*
Records updated (1 d)	*Meeting with architect (1 d)*
Map survey (1 d)	*Estimate prepared (1 d)*
Stores ordered and delivered (25 d)	*Plans prepared (1 d)*
Consents and wayleaves obtained (15 d)	*Site survey (1 d)*

(20 marks)

A25 (*a*) Critical path is defined as a path from a start event to an end event, the total duration of which is not less than that of any other path between the two events.

(*b*) (*i*) The flow diagram is shown in the sketch.



(*ii*) The activities included in the critical path are:
1–2, 2–3, 3–4, 4–5, 5–6, stores ordered and delivered, 7–8 and 8–9.

(*iii*) The latest date for end event (LDE) is day 61.

(*iv*) The earliest date for end event (EDE) is day 41.

(*v*) On a forward pass with the duration for obtaining wayleaves increased to 39 d, EDE for event 6 is day 5.

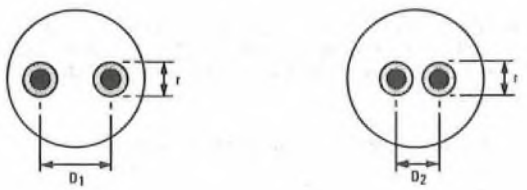
Therefore, the EDE for event 7 is day 44, the EDE for event 8 is day 54 and the EDE for end event is day 55.

Therefore, since the surfacing work is not to commence until day 61, the project can still be completed.

Q26 *If, during the manufacture of the cable referred to in Q24 (d), a length of cable is produced in which the conductors are located closer to each other (but not touching), and the other dimensions remain the same, explain what effect this will have on the cable's primary coefficients.*

(10 marks)

A26 The sketch shows the two conditions, where D_1 is greater than D_2.



(*i*) *The effect on resistance* (*R*) *per unit length*
Since the resistance is proportional to the length of the cable and to the cross-sectional area of the cable and neither of these are changed, the effect on the resistance per unit length is nil.

(*ii*) *The effect on the capacitance* (*C*) *per unit length*
The value of the capacitance per unit length will increase as the distance between the conductors (D) is reduced and

$$C = \frac{\text{a constant}}{\log_{10} \frac{D}{r}},$$

where r is the radius of the conductors.

(*iii*) *The effect on inductance* (*L*) *per unit length*
The value of the inductance per unit length will decrease as the distance between the conductors (D) is reduced and

$$L = \text{a constant} \times \log_{10} \frac{D}{r},$$

where r is the radius of the conductors.

(*iv*) *The effect on conductance* (*G*) *per unit length*
The value of the conductance will increase as the value of D is decreased and the value of the insulation resistance is decreased.

TEXT BOOKS FOR TECHNICIAN EDUCATION COUNCIL COURSES

The following textbooks which cover aspects of the Technician Education Council Courses have been reviewed in the *Journal*.

Mathematics Level 1

Technician Mathematics 1. M. G. Page. Cassell and Co. Ltd. (Cassell's TEC Series). xiv + 402 pp. 129 ills. £2·75. Reviewed Vol. 71, p. 207, Oct. 1978.

General Mathematics for Technicians. H. G. Davis and G. A. Hicks. McGraw-Hill. 294 pp. 249 ills. £3·35. Reviewed Vol. 71, p. 246, Jan. 1979.

Mathematics Level 2

Mathematics Level II. J. Morris. Van Nostrand Reinhold. xii + 119 pp. 143 ills. Cloth cover: £8·00; paperback £3·50. Reviewed Vol. 73, p. 153, Oct. 1980.

Technician Mathematics 2. M. G. Page. Cassell and Co. Ltd. v + 346 pp. 122 ills. £4·00. Reviewed Vol. 73, p. 262, Jan. 1981.

Mathematics Level 3

Mathematics Level III. J. Morris. Van Nostrand Reinhold. ix + 94 pp. 95 ills. Cloth cover: £6·50; paperback: £3·00. Reviewed Vol. 73, p. 153, Oct. 1980.

Electrical Drawing Level 1

Electrical Drawing for Technicians 1. F. Linsley. Newnes-Butterworth. x + 83 pp. 135 ills. £3·75. Reviewed Vol. 73, p. 53, Apr. 1980.

Electrical Principles Level 2

Electrical Principles for Technicians 2. S. A. Knight, Newnes-Butterworth. 128 pp. 139 ills. £3·75. Reviewed Vol. 72, p. 179, Oct. 1979.

Transmission Systems Level 2

Transmission System, TEC Level II, D. C. Green, Pitman. 138 pp. 120 ills. £3·20. Reviewed Vol. 72, p. 122, July 1979.

Radio System Level 2

Radio Systems II. D. C. Green. Pitman. 116 pp. 101 ills. £3·00. Reviewed Vol. 71, p. 272, Jan. 1979.

Science Level 1

Physical Science for Technicians. W. Bolton. McGraw Hill. 135 pp. 135 ills. £2·25. Reviewed Vol. 70, p. 177, Oct. 1977.

Technician Physical Science I. R. G. Meadows. Cassell and Co. Ltd. (Cassells TEC Series). 253 pp. 138 ills. £2·75. Reviewed Vol. 71, p. 66, Apr. 1978.

Digital Techniques Level 3 and Transmission System Level 3

Digital Techniques and Systems. D. C. Green. Pitman. x + 182 pp. 223 ills. £4·95. Reviewed Vol. 73, p. 267, Jan. 1981.

Materials and Workshop Practices Level 1

Technician Workshop Processes and Materials 1. C. R. Shotbolt, Cassell and Co Ltd. (Cassells TEC Series). 206 pp. 179 ills. £3·00. Reviewed Vol. 72, p. 42, Apr. 1979.

CGLI, TEC AND SCOTEC: GUIDANCE FOR STUDENTS

Back numbers of the *POEEJ* are available, complete with *Supplements* containing model answers to past examinations of the City and Guilds of London Institute Telecommunications Technician Courses and model questions and answers for students following the syllabi of Technician Education Council and the Scottish Technical Education Council courses for telecommunications technicians.

The price of back numbers is £1·30 each, including the *Supplement* and postage and packaging. (*Supplements* are not sold separately).

Apr.	1973	PM^{Q3-10}, ETP, ES, TPA, TpB, TPB^{Q1-6} (1972)	☐
Oct.	1973	TPC^{Q8-10}, CRC, TgC, MB, CB, BMCC, TpC^{Q1-3} (1972)	☐
Apr.	1974	LTC^{Q2-10}, LPPC (1972); ETP, PM, ES, RLTA, TPA (1973)	☐
July	1974	MA, TTA, LPPA, CA, MB, TpB, TgB^{Q1-6} (1973)	☐
Oct.	1974	TgB^{Q7-10}, LPPB, CB, MC, CRC, LTC, TpC^{Q1-9} (1973)	☐
Jan.	1975	TpC^{Q10}, TPC, TgC, LPPC, BMCC (1973); PM, ES, MA^{Q1-9} (1974)	☐
July	1975	TgB, TpB, LPPB, CB^{Q1-9} (1974)	☐
Jan.	1976	LPPC^{Q8-10}, BMCC (1974); ETP, PM, ES (1975)	☐
July	1976	TTA, RLTB, TpB, TgB (1975)	☐
Oct.	1976	LPPB, CB, TPC, MC, CRC, TgC, TpC, LTC (1975)	☐
Jan.	1977	LPPC, BMCC (1975); PM, RLTA (1976)	☐
Apr.	1977	ES, ETP, TPA, MA, LPPA, CA, MB, TPB (1976)	☐
July	1977	TpB, TgB, CB, LPPB (1976)	☐
Oct.	1977	TTA, TPC, MC, TpC, TgC, BMCC, LPPC, LTC (1976)	☐
Jan.	1978	[M1, PS1, TS1]; CRC (1976); PM, ES, RLTA (1977)	☐
Apr.	1978	[LCA1, M2, EP2, E2]; ETP, TPA, MA, LPPA, CA, MB, RLTB, CB (1977)	☐
July	1978	[TSS2, XS2, RS2]; RLTB (1976); TTA, TpB, MC (1977)	☐
Oct.	1978	TgB, LPPB, TPB, TPC, TpC, TgC, LPPC, CRC, BMCC, LTC (1977)	☐
Jan.	1979	TS1, PS1, LCA1, M1; SEEP1, SITS1, SM1/2	☐
Apr.	1979	PM, ES, TPA, RLTA, LPPA, CA, TPB, MB, TpC, MA (1978)	☐
July	1979	ETP, CB, BMC, TgC (1978)	☐
Oct.	1979	MC, LTC, TPC, TgB (1978)	☐
Jan.	1980	TS1, PS1, TSS2, EP2; MA (1979), TpB (1978), TPC, RLTA, TTA (1979)	☐
Apr.	1980	TPA, LPPA, CA, TPB, MB, CB, TgC (1979); M1, M2	☐
July	1980	MC, TgB, TpC, BMCC (1979); XS2	☐
Oct.	1980	TTA, TpB, LTC (1979); E2	☐
Jan.	1981	LPPB, LPPC (1979), MB, TPB, TgB, TPC, TpC, TgC (1980); EP2	☐

CGLI

BMCC: Basic Microwave Communication C
CA, CB: Computers A, B
CRC: Communication Radio C
ES: Engineering Science
ETP: Elementary Telecommunication Practice
LPPA, LPPB, LPPC: Line Plant Practice A, B, C
LTC: Line Transmission C
MA, MB, MC: Mathematics A, B, C
PM: Practical Mathematics
RLTA, RLTB: Radio and Line Transmission A, B
TgB, TgC: Telegraphy B, C
TpB, TpC: Telephony B, C
TPA, TPB, TPC: Telecommunication Principles A, B, C
TTA: Telephony and Telegraphy A

TEC

[ ] Denotes table of references
M1, M2: Mathematics 1, 2
PS1: Physical Science 1
LCA1: Line and Customer Apparatus 1
TS1: Telecommunication Systems 1
EP2: Electrical Principles 2
E2: Electronics 2
TSS2: Telephone Switching Systems 2
XS2: Transmission Systems 2
RS2: Radio Systems 2

SCOTEC

SEEP1: Electrical and Engineering Principles 1
SITS1: Introduction to Telecommunications Systems 1
SM1/2: Mathematics 1 and 2

Model-Answer Books A series of model-answer books is also available, in which selected answers from past CGLI examinations have been collected together to cover the following subjects.

Elementary Telecommunication Practice	60p	☐
Telecommunication Principles A	£1·00	☐
Line Plant Practice A	60p	☐
Radio and Line Transmission A	85p	☐
Telephony and Telegraphy A	85p	☐
Telecommunication Principles B	£1·00	☐

Orders, by post only, should be addressed to *The Post Office Electrical Engineers' Journal* (Sales), 2–12 Gresham Street, London EC2V 7AG. This panel, or a photocopy, may be used as an order form by marking the appropriate boxes. Cheques and postal orders, payable to "*The POEE Journal*", should be crossed "& Co." and enclosed with the order. Cash should not be sent through the post.

Published by *The Post Office Electrical Engineers' Journal*, 2–12 Gresham Street, London EC2V 7AG, and printed in Great Britain by Unwin Brothers Limited, Old Woking, Surrey