Radio Diary 1970

COLLINS

LONDON & GLASGOW

LAST YEAR 1969

	January	February	March
S	5 12 19 26	2 9 16 23	2 9 16 23 30
M	6 13 20 27	3 10 17 24	3 10 17 24 31
T	7 14 21 28	4 11 18 25	4 11 18 25
W	1 8 15 22 29	5 12 19 26	5 12 19 26
T	2 9 16 23 30	6 13 20 27	6 13 20 27
F	3 10 17 24 31	7 14 21 28	7 14 21 28
S	4 11 18 25	1 8 15 22	1 8 15 22 29

	April	May	June
S	6 13 20 27	4 11 18 25	1 8 15 22 29
M	7 14 21 28	5 12 19 26	2 9 16 23 30
T	1 8 15 22 29	6 13 20 27	3 10 17 24
W	2 9 16 23 30	7 14 21 28	4 11 18 25
T	3 10 17 24	1 8 15 22 29	5 12 19 26
F	4 11 18 25	2 9 16 23 30	6 13 20 27
S	5 12 19 26	3 10 17 24 31	7 14 21 28

	July	August	September
S	6 13 20 27	3 10 17 24 31	7 14 21 28
M	7 14 21 28	4 11 18 25	1 8 15 22 29
T	1 8 15 22 29	5 12 19 26	2 9 16 23 30
W	2 9 16 23 30	6 13 20 27	3 10 17 24
T	3 10 17 24 31	7 14 21 28	4 11 18 25
F	4 11 18 25	1 8 15 22 29	5 12 19 26
S	5 12 19 26	2 9 16 23 30	6 13 20 27

	October	November	December
S	5 12 19 26	2 9 16 23 30	7 14 21 28
M	6 13 20 27	3 10 17 24	1 8 15 22 29
T	7 14 21 28	4 11 18 25	2 9 16 23 30
W	1 8 15 22 29	5 12 19 26	3 10 17 24 31
T	2 9 16 23 30	6 13 20 27	4 11 18 25
F	3 10 17 24 31	7 14 21 28	5 12 19 26
S	4 11 18 25	1 8 15 22 29	6 13 20 27

Easter Day, April 6

THIS YEAR 1970

	January	February	March
S	4 11 18 25	1 8 15 22	1 8 15 22 29
M	5 12 19 26	2 9 16 23	2 9 16 23 30
T	6 13 20 27	3 10 17 24	3 10 17 24 31
W	7 14 21 28	4 11 18 25	4 11 18 25
T	1 8 15 22 29	5 12 19 26	5 12 19 26
F	2 9 16 23 30	6 13 20 27	6 13 20 27
S	3 10 17 24 31	7 14 21 28	7 14 21 28

	April	May	June
S	5 12 19 26	3 10 17 24 31	7 14 21 28
M	6 13 20 27	4 11 18 25	1 8 15 22 29
T	7 14 21 28	5 12 19 26	2 9 16 23 30
W	1 8 15 22 29	6 13 20 27	3 10 17 24
T	2 9 16 23 30	7 14 21 28	4 11 18 25
F	3 10 17 24	1 8 15 22 29	5 12 19 26
S	4 11 18 25	2 9 16 23 30	6 13 20 27

	July	August	September
S	5 12 19 26	2 9 16 23 30	6 13 20 27
M	6 13 20 27	3 10 17 24 31	7 14 21 28
T	7 14 21 28	4 11 18 25	1 8 15 22 29
W	1 8 15 22 29	5 12 19 26	2 9 16 23 30
T	2 9 16 23 30	6 13 20 27	3 10 17 24
F	3 10 17 24 31	7 14 21 28	4 11 18 25
S	4 11 18 25	1 8 15 22 29	5 12 19 26

	October	November	December
S	4 11 18 25	1 8 15 22 29	6 13 20 27
M	5 12 19 26	2 9 16 23 30	7 14 21 28
T	6 13 20 27	3 10 17 24	1 8 15 22 29
W	7 14 21 28	4 11 18 25	2 9 16 23 30
T	1 8 15 22 29	5 12 19 26	3 10 17 24 31
F	2 9 16 23 30	6 13 20 27	4 11 18 25
S	3 10 17 24 31	7 14 21 28	5 12 19 26

Easter Day, March 29

NEXT YEAR 1971

	January	February	March
S	3 10 17 24 31	7 14 21 28	7 14 21 28
M	4 11 18 25	1 8 15 22	1 8 15 22 29
T	5 12 19 26	2 9 16 23	2 9 16 23 30
W	6 13 20 27	3 10 17 24	3 10 17 24 31
T	7 14 21 28	4 11 18 25	4 11 18 25
F	1 8 15 22 29	5 12 19 26	5 12 19 26
S	2 9 16 23 30	6 13 20 27	6 13 20 27

	April	May	June
S	4 11 18 25	2 9 16 23 30	6 13 20 27
M	5 12 19 26	3 10 17 24 31	7 14 21 28
T	6 13 20 27	4 11 18 25	1 8 15 22 29
W	7 14 21 28	5 12 19 26	2 9 16 23 30
T	1 8 15 22 29	6 13 20 27	3 10 17 24
F	2 9 16 23 30	7 14 21 28	4 11 18 25
S	3 10 17 24	1 8 15 22 29	5 12 19 26

	July	August	September
S	4 11 18 25	1 8 15 22 29	5 12 19 26
M	5 12 19 26	2 9 16 23 30	6 13 20 27
T	6 13 20 27	3 10 17 24 31	7 14 21 28
W	7 14 21 28	4 11 18 25	1 8 15 22 29
T	1 8 15 22 29	5 12 19 26	2 9 16 23 30
F	2 9 16 23 30	6 13 20 27	3 10 17 24
S	3 10 17 24 31	7 14 21 28	4 11 18 25

	October	November	December
S	3 10 17 24 31	7 14 21 28	5 12 19 26
M	4 11 18 25	1 8 15 22 29	6 13 20 27
T	5 12 19 26	2 9 16 23 30	7 14 21 28
W	6 13 20 27	3 10 17 24	1 8 15 22 29
T	7 14 21 28	4 11 18 25	2 9 16 23 30
F	1 8 15 22 29	5 12 19 26	3 10 17 24 31
S	2 9 16 23 30	6 13 20 27	4 11 18 25

Easter Day, April 11

BANK & PUBLIC HOLIDAYS 1970

England, N. Ireland, Wales

St. Patrick's Day (Ireland)	March 17
Good Friday	March 27
Easter Monday	March 30
Spring Bank Holiday	May 25
Orangeman's Day Holiday (N. Ireland)	July 13
Summer Bank Holiday	August 31
Christmas Day (Friday)	December 25
Boxing Day	December 26

Scotland

New Year's Day	January 1
Good Friday	March 27
Spring Bank Holiday	May 25
Summer Bank Holiday	August 3
Christmas Day (Friday)	December 25

QUARTER DAYS

England, Ireland, Wales

Lady Day	March 25
Midsummer	June 24
Michaelmas	Sept 29
Christmas	Dec 25

Scotland

Candlemas	Feb 2
Whitsunday	May 15
Lammas	Aug 1
Martinmas	Nov 11

LAW

Sittings

Hilary	Jan 11–Mar 25
Easter	Apr 7–May 15
Trinity	May 26–Jul 31
Mich'lmas	Oct 1–Dec 21

Dining Terms

Hilary	Jan 21–Feb 12
Easter	Apr 8–Apr 30
Trinity	Jun 24–Jul 16
Mich'lmas	Nov 4–Nov 26

UNIVERSITY FULL TERMS

Oxford

Hilary	Jan 18–Mar 14
Trinity	Apr 26–Jun 20
Mich'lmas	*Oct 11–Dec 5

Cambridge

Lent	Jan 13–Mar 13
Easter	Apr 21–Jun 12
Mich'lmas	Oct 6–Dec 4

**Provisional*

Radio Diary 1970

COLLINS

LONDON & GLASGOW

Contents

Useful tables	8-19
Resistance	20-22
Inductance	23-26
Capacitance	27-28
Time constant	29
Alternating current	30-35
Sound	36-40
Useful networks	41-48
Power supplies	49-53
Audio frequency	54-73
Electro-magnetic waves	74-81
Transmission	82-87
Reception	88-93
Microwaves	94-96
Television	97-103
Colour television	104-115
Feeders and aerials	116-123
Ultra-high frequencies	124-126
Semi-conductor devices	127-142
Receiving valves	143-158
Index	159-160

Useful tables

WEIGHTS AND MEASURES

Linear measure

4 ins.	1 hand = 10·16 cms.	6 feet	1 fathom	= 1·828 metre
9 ins.	1 span = 22·86 cms.	22 yards		= 1 chain
12 ins.	1 foot = 30·48 cms.	1 metre		= 39·37 inches
3 feet	1 yard = 0·914 metre	10 chains		= 1 furlong
5 feet	1 pace = 1·524 metre	8 furlongs		= 1 mile

Cubic or solid measure

Cubic foot = 1728 cub. inches × 16·387 = 28317 cub. centimetres.
Cubic yard = 27 cubic feet = 21·033 bushels = 0·7645 cubic metre.
Shipping ton = 40 cubic feet of merchandise = 1·13 cubic metre.
Shipping ton = 42 cubic feet of timber = 1·18 cubic metre.
One ton or load = 50 cubic feet of hewn timber = 1·42 cubic metre.
Ton of displacement of a ship = 35 cubic feet = 1·02 cubic metre

Square or land measure

144	square inches	= 1 square foot
9	square feet	= 1 square yard
1210	square yards	= 1 rood
4	roods	= 1 acre (0·407 hectares)
640	acres	= 1 square mile
1	square link	= 62¼ square inches (approx.)
1	square chain	= 10,000 square links = 484 square yards.
33	square yards	= 1 rod of building = 27·6 square metre.
100	square feet	= Square of flooring or roofing = 9·3 sq. metre.
272¼	square feet	= Rod of bricklayer's work = 25·4 sq. metre.

Avoirdupois weight

16 drams	= 1 oz. (437·5 gr.).	28 lbs.	= 1 qr.
16 ounces	= 1 pound (lb.).	112 lbs.	= 1 cwt.
14 pounds	= 1 stone	20 cwts.	= 1 ton

Fluid memoranda

1 cubic foot of water = 6¼ gals. (approx.) = 62½ lb. = 7·48 U.S. gal.
1 U.S. gal. = 231 cub. in. = 0·1337 cub. ft.
1 lb. water at 62°F. = 0·016 cub. ft.
1 B.I. gal. = 277·418 cub. in. 1 cwt. of water = 1·8 cu. ft. = 11·2 gal.
1 British = 1·2009 U.S. gal. 1 ton of water = 35·9 cu. ft. = 224 gal.
1 inch of rainfall = 22,622 gals. per acre = 100 tons (approx.).

	lb./gal.		lb./gal.
Acetic acid	10·49	Petrol	7·5
Alcohol	8	Sperm oil	8·8
Hydrochloric acid	12·0	Sulphuric acid 98 %	18·35
Mercury	135·9	Turpentine	8·7
Milk	10·3	Water (distilled)	10·0

METRIC SYSTEM

The unit of length (or lin. measure) is the metre
The unit of surface (or sq. measure) is the area (100 sq. metres)
The unit of capacity is the litre (1000 cu. cm.)
The unit of mass is the gramme or gram (mass of 1 cc. of water)

Multiples and sub-multiples are denoted by the following prefixes:

deca-	= × 10,	thus 1 decalitre	= 10 litres
hecto-	= × 100,	thus 1 hectogram	= 100 grams
kilo-	= × 1,000,	thus 1 kilometre	= 1,000 m.
mega-	= × 1,000,000,	thus 1 megahertz	= 1 million Hz
deci-	= ÷ 10,	thus 1 decimetre	= 1/10th m.
centi-	= ÷ 100,	thus 1 centimetre	= 1/100th m.
milli-	= ÷ 1,000,	thus 1 milligram	= 1/1000th g.
micro-	= ÷ 1,000,000,	thus 1 microvolt	= 1/1,000,000th of a volt
nano	= ÷ 10^9,	thus 1 nanosecond	= 1/10^9 of a second

Micron is 1/1,000,000th of a metre
Tonne is 1,000,000 grams (1,000 Kg.).

Linear measure

1 centimetre	= 0·3937 ins.	1 inch	= 2·54 centimetres
1 metre	= 39·3708 ins.	1 yard	= 0·914 metres
1 kilometre	= 0·6214 miles	1 mile	= 1·6093 kilometres

Square measure

1 sq. cm.	= 0·155 sq. in.	1 sq. in.	= 6·45 sq. cm.
1 sq. metre	= 1·196 sq. yds.	1 sq. yd.	= 0·836 sq. metres
1 hectare	= 2·471 acres	1 acre	= 0·4047 hectare

(*N.B.*—1 hectare = 10·000 sq. metres)

Measure of capacity

1 cu. cm.	= 0·061 cu. in.	1 cu. in.	= 16·39 cu. cm.
1 litre	= 0·0353 cu. ft.	1 cu. ft.	= 28·3 litres
1 litre	= 0·22 gallons	1 gallon	= 4·546 litres

Measure of weight

1 milligram	= 0·015 grain	1 grain	= 64·8 milligrams
1 gramme	= 0·0352 ounce	1 ounce	= 28·35 grammes
1 kilogram	= 2·2046 lbs.	1 pound	= 0·4536 kilograms
1 tonne	= 0·984 tons	1 ton	= 1·016 tonnes

General Conversions

	To obtain	*From*	*Multiply by*
Multiply by	*To convert*	*To*	
2·54	inches	centimetres	·3937
30·48	feet	centimetres	·0328
·914	yards	metres	1·094
1,609·3	miles	metres	·000621
1,853·27	nautical miles	metres	·000539
6·45	square inches	sq. cms.	·155
·093	square feet	sq. metres	10·764
·836	square yards	sq. metres	1·196
16·39	cubic inches	cub. cms.	·061
28·3	cubic feet	litres	·0353
6·24	cubic feet	gallons	·1602
·765	cubic yards	cub. metres	1·308
·3732	pounds (troy)	kilogrammes	2·68
31·10	ounces (troy)	grammes	·03216
·4536	pounds (avoir.)	kilogrammes	2·2045
7,000	pounds (avoir.)	grains (troy)	·000143
28·35	ounces (avoir)	grammes	·0352
·065	grains	grammes	·1538
50·8	cwt.	kilogrammes	·01968
1,016·0	tons	kilogrammes	·000984
4·546	gallons	litres	·22
10	gallons of water	pounds	·1
·454	pounds of water	litres	2·202
70·3	lb. per sq. in.	gm./sq. cm.	·0142
2·3	lb. per sq. in.	head of water(ft.)	·434
·7	lb. per sq. in.	head of water (M)	1·4285
·068	lb. per sq. in.	atmospheres	14·7
1·575	tons per sq. in.	kgm./sq. mm.	·635
4·883	lb. per sq. ft.	kgm./sq. metre	·205
·593	lb. per cub. yd.	kgm./cub. metre	1·686
16·02	lb. per cub. ft.	kgm./cub. metre	·0624
·0998	lb. per gallon	kgm./litre	10·02
·138	foot-lb.	k'grammetres	7·23
·33	foot-tons	tonne-metres	3
1·014	horse-power	force de cheval	·9861
746	horse-power	watts	·00134
33,000	horse-power	ft.-lb./min.	1/33000
76	horse-power	kg.-m/sec.	·01316
44	watts	ft.-lb./min.	·0227
0·1	watts	kg.-m./sec.	10
0·252	B.Th.U.	kg. calories	3·97
14·7	atmospheres	lb./sq. inch	·068
0·90	German candles	English candles	1·1111
9·55	carcels	candles	·1047
·737	joules	ft.-lb.	1·357
88	miles/hour	ft./min.	·01136
197	metres/sec.	ft./min.	·00508
1·8	C.H.U.	B.Th.U.	·5555
·0000208	centipoise	lb. force sec./sq. ft.	48,000

SQUARES, CUBES, SQUARE ROOTS AND CUBE ROOTS

No.	Square	Cube	Square root	Cube root
⅛	·015	·0019	·353	·5
¼	·062	·0156	·500	·629
⅜	·140	·0527	·612	·721
½	·250	·1250	·707	·793
⅝	·390	·244	·790	·855
¾	·562	·421	·866	·908
⅞	·765	·670	·935	·956
1	1	1	1	1
1⅛	1·265	1·423	1·060	1·04
1¼	1·562	1·953	1·118	1·07
1⅜	1·890	2·599	1·172	1·11
1½	2·250	3·375	1·224	1·14
1⅝	2·641	4·291	1·274	1·17
1¾	3·062	5·359	1·322	1·20
1⅞	3·515	6·591	1·369	1·23
2	4	8	1·414	1·26
2⅛	4·515	9·595	1·457	1·28
2¼	5·062	11·390	1·500	1·30
2⅜	5·640	13·396	1·541	1·33
2½	6·250	15·625	1·581	1·35
2⅝	6·890	18·088	1·620	1·37
2¾	7·562	20·796	1·658	1·40
2⅞	8·265	23·763	1·695	1·42
3	9	27	1·732	1·44
3⅛	9·765	30·517	1·767	1·46
3¼	10·562	34·328	1·802	1·48

No.	Square	Cube	Square root	Cube root
3⅜	11·390	38·443	1·837	1·50
3½	12·250	42·875	1·870	1·51
3⅝	13·140	47·634	1·903	1·53
3¾	14·062	52·734	1·936	1·55
3⅞	15·015	58·185	1·968	1·57
4	16	64	2	1·58
4¼	18·062	76·765	2·061	1·61
4½	20·250	91·125	2·121	1·65
4¾	22·562	107·171	2·179	1·68
5	25	125	2·236	1·71
5¼	27·562	144·703	2·291	1·73
5½	30·250	166·375	2·345	1·76
5¾	33·062	190·109	2·397	1·79
6	36	216	2·449	1·81
6¼	39·062	244·140	2·500	1·84
6½	42·250	274·625	2·549	1·86
6¾	45·562	307·546	2·598	1·88
7	49	343	2·645	1·91
7¼	52·562	381·078	2·692	1·93
7½	56·250	421·875	2·738	1·95
7¾	60·062	465·484	2·783	1·97
8	64	512	2·828	2
8¼	68·062	561·515	2·872	2·02

No.	Square	Cube	Square root	Cube root
8½	72·250	614·125	2·915	2·04
8¾	76·562	669·921	2·958	2·06
9	81	729	3	2·08
9¼	85·562	791·453	3·041	2·09
9½	90·25	857·375	3·082	2·11
9¾	95·062	926·859	3·122	2·13
10	100	1000	3·162	2·15
10¼	105·062	1076·89	3·201	2·17
10½	110·250	1157·625	3·240	2·18
10¾	115·562	1242·296	3·278	2·20
11	121	1331	3·316	2·22
11¼	126·562	1423·828	3·354	2·24
11½	132·250	1520·875	3·391	2·25
11¾	138·062	1622·234	3·427	2·27
12	144	1728	3·464	2·29
12¼	150·062	1838·265	3·500	2·31
12½	156·250	1953·125	3·535	2·32
12¾	162·562	2072·672	3·572	2·34
13	169	2197	3·606	2·35
13¼	175·562	2326·203	3·640	2·36
13½	182·250	2460·375	3·675	2·38
13¾	189·062	2599·609	3·710	2·39
14	196	2744	3·742	2·41

Trigonometrical Ratios

Angle Deg.	Angle Rad.	Sine	Tangent	Co-tangent	Cosine		
0°	0	0	0	∞	1	1·5708	90°
·5	·0087	·0087	·0087	114·6	1	1·5621	89·5
1	·0175	·0175	·0175	57·2900	·9998	1·5533	89
1·5	·0262	·0262	·0262	38·19	·9997	1·5446	88·5
2	·0349	·0349	·0349	28·6363	·9994	1·5359	88
2·5	·0436	·0436	·0437	22·90	·9990	1·5272	87·5
3	·0524	·0523	·0524	19·0811	·9986	1·5184	87
3·5	·0611	·0610	·0612	16·35	·9981	1·5097	86·5
4	·0698	·0698	·0699	14·3006	·9976	1·5010	86
4·5	·0785	·0785	·0787	12·71	·9969	1·4923	85·5
5	·0873	·0872	·0875	11·4301	·9962	1·4835	85
5·5	·0960	·0958	·0963	10·39	·9954	1·4748	84·5
6	·1047	·1045	·1051	9·5144	·9945	1·4661	84
6·5	·1134	·1132	·1139	8·7769	·9936	1·4573	83·5
7	·1222	·1219	·1228	8·1443	·9925	1·4486	83
7·5	·1309	·1305	·1317	7·5958	·9914	1·4399	82·5
8	·1396	·1392	·1405	7·1154	·9903	1·4312	82
8·5	·1484	·1478	·1495	6·6912	·9890	1·4224	81·5
9	·1571	·1564	·1584	6·3138	·9877	1·4137	81
9·5	·1658	·1650	·1673	5·9758	·9863	1·4050	80·5
10	·1745	·1736	·1763	5·6713	·9848	1·3963	80
10·5	·1833	·1822	·1853	5·3955	·9833	1·3875	79·5
11	·1920	·1908	·1944	5·1446	·9816	1·3788	79
11·5	·2007	·1994	·2035	4·9152	·9799	1·3701	78·5
12	·2094	·2079	·2126	4·7046	·9781	1·3614	78
12·5	·2182	·2164	·2217	4·5107	·9763	1·3526	77·5
13	·2269	·2250	·2309	4·3315	·9744	1·3439	77
13·5	·2356	·2334	·2401	4·1653	·9724	1·3352	76·5
14	·2443	·2419	·2493	4·0108	·9703	1·3265	76
14·5	·2531	·2504	·2586	3·8667	·9681	1·3177	75·5
15	·2618	·2588	·2679	3·7321	·9659	1·3090	75
15·5	·2705	·2672	·2773	3·6059	·9636	1·3003	74·5
16	·2793	·2756	·2867	3·4874	·9613	1·2915	74
16·5	·2880	·2840	·2962	3·3759	·9588	1·2828	73·5
17	·2967	·2924	·3057	3·2709	·9563	1·2741	73
17·5	·3054	·3007	·3153	3·1716	·9537	1·2654	72·5
18	·3142	·3090	·3249	3·0777	·9511	1·2566	72
18·5	·3229	·3173	·3346	2·9887	·9483	1·2479	71·5
19	·3316	·3256	·3443	2·9042	·9455	1·2392	71
19·5	·3403	·3338	·3541	2·8239	·9426	1·2305	70·5
20	·3491	·3420	·3640	2·7475	·9397	1·2217	7
20·5	·3578	·3502	·3739	2·6746	·9367	1·2130	69·5
21	·3665	·3584	·3839	2·6051	·9336	1·2043	69
21·5	·3752	·3665	·3939	2·5386	·9304	1·1956	68·5
22	·3840	·3746	·4040	2·4751	·9272	1·1868	68
		Cosine	Co-tangent	Tangent	Sine	Angle Rad.	Angle Deg.

Trigonometrical Ratios

Angle		Sine	Tangent	Co-tangent	Cosine		
Deg.	Rad.						
22·5°	·3927	·3827	·4142	2·4142	·9239	1·1781	67·5
23	·4014	·3907	·4245	2·3559	·9205	1·1694	67
23·5	·4102	·3987	·4348	2·2998	·9171	1·1606	66·5
24	·4189	·4067	·4452	2·2460	·9135	1·1509	66
24·5	·4276	·4147	·4557	2·1943	·9100	1·1432	65·5
25	·4363	·4226	·4663	2·1445	·9063	1·1345	65
25·5	·4451	·4305	·4770	2·0965	·9026	1·1257	64·5
26	·4538	·4384	·4877	2·0503	·8988	1·1170	64
26·5	·4625	·4462	·4986	2·0057	·8949	1·1083	63·5
27	·4712	·4540	·5095	1·9626	·8910	1·0996	63
27·5	·4800	·4617	·5206	1·9210	·8870	1·0908	62·5
28	·4887	·4695	·5317	1·8807	·8829	1·0821	62
28·5	·4974	·4772	·5430	1·8418	·8788	1·0734	61·5
29	·5061	·4848	·5543	1·8040	·8746	1·0647	61
29·5	·5149	·4924	·5658	1·7675	·8704	1·0559	60·5
30	·5236	·5000	·5774	1·7321	·8660	1·0472	60
30·5	·5323	·5075	·5890	1·6977	·8616	1·0385	59·5
31	·5411	·5150	·6009	1·6643	·8572	1·0297	59
31·5	·5498	·5225	·6128	1·6319	·8526	1·0210	58·5
32	·5585	·5299	·6249	1·6003	·8480	1·0123	58
32·5	·5672	·5373	·6371	1·5697	·8434	1·0036	57·5
33	·5760	·5446	·6494	1·5399	·8387	·9948	57
33·5	·5847	·5519	·6619	1·5108	·8339	·9861	56·5
34	·5934	·5592	·6745	1·4826	·8290	·9774	56
34·5	·6021	·5664	·6873	1·4550	·8241	·9687	55·5
35	·6109	·5736	·7002	1·4281	·8192	·9599	55
35·5	·6196	·5807	·7133	1·4019	·8141	·9512	54·5
36	·6283	·5878	·7265	1·3764	·8090	·9425	54
36·5	·6370	·5948	·7400	1·3514	·8039	·9338	53·5
37	·6458	·6018	·7536	1·3270	·7986	·9250	53
37·5	·6545	·6088	·7673	1·3032	·7934	·9163	52·5
38	·6632	·6157	·7813	1·2799	·7880	·9076	52
38·5	·6720	·6225	·7954	1·2572	·7826	·8988	51·5
39	·6807	·6293	·8098	1·2349	·7771	·8901	51
39·5	·6894	·6361	·8243	1·2131	·7716	·8814	50·5
40	·6981	·6428	·8391	1·1918	·7660	·8727	50
40·5	·7069	·6494	·8541	1·1708	·7604	·8639	49·5
41	·7156	·6561	·8693	1·1504	·7547	·8552	49
41·5	·7243	·6626	·8847	1·1303	·7490	·8465	48·5
42	·7330	·6691	·9004	1·1106	·7431	·8378	48
42·5	·7418	·6756	·9163	1·0913	·7373	·8290	47·5
43	·7505	·6820	·9325	1·0724	·7314	·8203	47
43·5	·7592	·6884	·9490	1·0538	·7254	·8116	46·5
44	·7679	·6947	·9657	1·0355	·7193	·8029	46
44·5	·7767	·7009	·9827	1·0176	·7133	·7941	45·5
45	·7854	·7071	1·0000	1 0000	·7071	·7854	45
		Cosine	Co-tangent	Tangent	Sine	Rad.	Deg.
						Angle	

LOGARITHMS

	0	1	2	3	4	5	6	7	8	9	1	3	5	7	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	4	12	21	29	37
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	4	11	19	26	34
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	3	10	17	24	31
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	3	10	16	23	29
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	3	9	15	21	27
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	3	8	14	20	25
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	3	8	13	18	24
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	2	7	12	17	22
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	2	7	12	16	21
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	2	7	11	16	20
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	2	6	11	15	19
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	2	6	10	14	18
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	2	6	10	14	17
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	2	6	9	13	17
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	2	5	9	12	16
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	2	5	9	12	15
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	2	5	8	11	15
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	2	5	8	11	14
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	2	5	8	11	14
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	1	4	7	10	13
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	1	4	7	10	13
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	1	4	7	10	12
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	1	4	7	9	12
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	1	4	6	9	12
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	1	4	6	9	11
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	1	4	6	9	11
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	1	4	6	8	11
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	1	3	6	8	10
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	1	3	5	8	10
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	1	3	5	8	10
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	1	3	5	8	10
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	1	3	5	7	9
42	6232	6243	6263	6263	6274	6284	6294	6304	6314	6325	1	3	5	7	9
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	1	3	5	7	9
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	1	3	5	7	9
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	1	3	5	7	9
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	1	3	5	7	8
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	1	3	5	6	8
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	1	3	4	6	8
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	1	3	4	6	8
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	1	3	4	6	8
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	1	3	4	6	8
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	1	2	4	6	7
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	1	2	4	6	7
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	1	2	4	6	7

Note: **Differences 2, 4, 6, 8 obtained by interpolation.**

Logarithms

	0	*1*	*2*	*3*	*4*	*5*	*6*	*7*	*8*	*9*	*1*	*3*	*5*	*7*	*9*
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	1	2	4	5	7
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	1	2	4	5	7
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	1	2	4	5	7
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	1	2	4	5	7
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	1	2	4	5	7
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	1	2	4	5	6
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	1	2	4	5	6
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	1	2	3	5	6
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	1	2	3	5	6
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	1	2	3	5	6
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	1	2	3	5	6
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	1	2	3	5	6
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	1	2	3	5	6
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	1	2	3	4	6
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	1	2	3	4	6
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	1	2	3	4	5
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	1	2	3	4	5
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	1	2	3	4	5
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	1	2	3	4	5
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	1	2	3	4	5
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	1	2	3	4	5
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	1	2	3	4	5
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	1	2	3	4	5
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	1	2	3	4	5
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	1	2	3	4	5
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	1	2	3	4	5
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	1	2	3	4	5
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	1	2	3	4	5
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	1	2	3	4	5
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	1	2	3	4	5
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340	1	2	3	4	5
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	1	2	3	4	5
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	0	2	2	3	4
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	0	1	2	3	4
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	0	1	2	3	4
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	0	1	2	3	4
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	0	1	2	3	4
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	0	1	2	3	4
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	0	1	2	3	4
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	0	1	2	3	4
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	0	1	2	3	4
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	0	1	2	3	4
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	0	1	2	3	4
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	0	1	2	3	4
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996	0	1	2	3	4

Note: Common logarithms = hyperperbolic logarithms × 0·43429.

ANTILOGARITHMS

	0	*1*	*2*	*3*	*4*	*5*	*6*	*7*	*8*	*9*	*1*	*3*	*5*	*7*	*9*
·00	1000	1002	1005	1007	1009	1012	1014	1016	1019	1021	0	1	1	2	2
·01	1023	1026	1028	1030	1033	1035	1038	1040	1042	1045	0	1	1	2	2
·02	1047	1050	1052	1054	1057	1059	1062	1064	1067	1069	0	1	1	2	2
·03	1072	1074	1076	1079	1081	1084	1086	1089	1091	1094	0	1	1	2	2
·04	1096	1099	1102	1104	1107	1109	1112	1114	1117	1119	0	1	1	2	2
·05	1122	1125	1127	1130	1132	1135	1138	1140	1143	1146	0	1	1	2	2
·06	1148	1151	1153	1156	1159	1161	1164	1167	1169	1172	0	1	1	2	2
·07	1175	1178	1180	1183	1186	1189	1191	1194	1197	1199	0	1	1	2	2
·08	1202	1205	1208	1211	1213	1216	1219	1222	1225	1227	0	1	1	2	3
·09	1230	1233	1236	1239	1242	1245	1247	1250	1253	1256	0	1	1	2	3
·10	1259	1262	1265	1268	1271	1274	1276	1279	1282	1285	0	1	1	2	3
·11	1288	1291	1294	1297	1300	1303	1306	1309	1312	1315	0	1	2	2	3
·12	1318	1321	1324	1327	1330	1334	1337	1340	1343	1346	0	1	2	2	3
·13	1349	1352	1355	1358	1361	1365	1368	1371	1374	1377	0	1	2	2	3
·14	1380	1384	1387	1390	1393	1396	1400	1403	1406	1409	0	1	2	2	3
·15	1413	1416	1419	1422	1426	1429	1432	1435	1439	1442	0	1	2	2	3
·16	1445	1449	1452	1455	1459	1462	1466	1469	1472	1476	0	1	2	2	3
·17	1479	1483	1486	1489	1493	1496	1500	1503	1507	1510	0	1	2	2	3
·18	1514	1517	1521	1524	1528	1531	1535	1538	1542	1545	0	1	2	2	3
·19	1549	1552	1556	1560	1563	1567	1570	1574	1578	1581	0	1	2	3	3
·20	1585	1589	1592	1596	1600	1603	1607	1611	1614	1618	0	1	2	3	3
·21	1622	1626	1629	1633	1637	1641	1644	1648	1652	1656	0	1	2	3	3
·22	1660	1663	1667	1671	1675	1679	1683	1687	1690	1694	0	1	2	3	3
·23	1698	1702	1706	1710	1714	1718	1722	1726	1730	1734	0	1	2	3	4
·24	1738	1742	1746	1750	1754	1758	1762	1766	1770	1774	0	1	2	3	4
·25	1778	1782	1786	1791	1795	1799	1803	1807	1811	1816	0	1	2	3	4
·26	1820	1824	1828	1832	1837	1841	1845	1849	1854	1858	0	1	2	3	4
·27	1862	1866	1871	1875	1879	1884	1888	1892	1897	1901	0	1	2	3	4
·28	1905	1910	1914	1919	1923	1928	1932	1936	1941	1945	0	1	2	3	4
·29	1950	1954	1959	1963	1968	1972	1977	1982	1986	1991	0	1	2	3	4
·30	1995	2000	2004	2009	2014	2018	2023	2028	2032	2037	0	1	2	3	4
·31	2042	2046	2051	2056	2061	2065	2070	2075	2080	2084	0	1	2	3	4
·32	2089	2094	2099	2104	2109	2113	2118	2123	2128	2133	0	1	2	3	4
·33	2138	2143	2148	2153	2158	2163	2168	2173	2178	2183	0	1	2	3	4
·34	2188	2193	2198	2203	2208	2213	2218	2223	2228	2234	1	2	3	4	5
·35	2239	2244	2249	2254	2259	2265	2270	2275	2280	2286	1	2	3	4	5
·36	2291	2296	2301	2307	2312	2317	2323	2328	2333	2339	1	2	3	4	5
·37	2344	2350	2355	2360	2366	2371	2377	2382	2388	2393	1	2	3	4	5
·38	2399	2404	2410	2415	2421	2427	2432	2438	2443	2449	1	2	3	4	5
·39	2455	2460	2466	2472	2477	2483	2489	2495	2500	2506	1	2	3	4	5
·40	2512	2518	2523	2529	2535	2541	2547	2553	2559	2564	1	2	3	4	5
·41	2570	2576	2582	2588	2594	2600	2606	2612	2618	2624	1	2	3	4	5
·42	2630	2636	2642	2649	2655	2661	2667	2673	2679	2685	1	2	3	4	6
·43	2692	2698	2704	2710	2716	2723	2729	2735	2742	2748	1	2	3	4	6
·44	2754	2761	2767	2773	2780	2786	2793	2799	2805	2812	1	2	3	4	6
·45	2818	2825	2831	2838	2844	2851	2858	2864	2871	2877	1	2	3	5	6
·46	2884	2891	2897	2904	2911	2917	2924	2931	2938	2944	1	2	3	5	6
·47	2951	2958	2965	2972	2979	2985	2992	2999	3006	3013	1	2	3	5	6
·48	3020	3027	3034	3041	3048	3055	3062	3069	3076	3083	1	2	4	5	6
·49	3090	3097	3105	3112	3119	3126	3133	3141	3148	3155	1	2	4	5	6

Note: **Differences, 2, 4, 6, 8 obtained by interpolation**

ANTILOGARITHMS

	0	1	2	3	4	5	6	7	8	9	1	3	5	7	9
·50	3162	3170	3177	3184	3192	3199	3206	3214	3221	3228	1	2	4	5	7
·51	3236	3243	3251	3258	3266	3273	3281	3289	3296	3304	1	2	4	5	7
·52	3311	3319	3327	3334	3342	3350	3357	3365	3373	3381	1	2	4	5	7
·53	3388	3396	3404	3412	3420	3428	3436	3443	3451	3459	1	2	4	6	7
·54	3467	3475	3483	3491	3499	3508	3516	3524	3532	3540	1	2	4	6	7
·55	3548	3556	3565	3573	3581	3589	3597	3606	3614	3622	1	2	4	6	7
·56	3631	3639	3648	3656	3664	3673	3681	3690	3698	3707	1	3	4	6	8
·57	3715	3724	3733	3741	3750	3758	3767	3776	3784	3793	1	3	4	6	8
·58	3802	3811	3819	3828	3837	3846	3855	3864	3873	3882	1	3	4	6	8
·59	3890	3899	3908	3917	3926	3936	3945	3954	3963	3972	1	3	5	6	8
·60	3981	3990	3999	4009	4018	4027	4036	4046	4055	4064	1	3	5	6	8
·61	4074	4083	4093	4102	4111	4121	4130	4140	4150	4159	1	3	5	7	9
·62	4169	4178	4188	4198	4207	4217	4227	4236	4246	4256	1	3	5	7	9
·63	4266	4276	4285	4295	4305	4315	4325	4335	4345	4355	1	3	5	7	9
·64	4365	4375	4385	4395	4406	4416	4426	4436	4446	4457	1	3	5	7	9
·65	4467	4477	4487	4498	4508	4519	4529	4539	4550	4560	1	3	5	7	9
·66	4571	4581	4592	4603	4613	4624	4634	4645	4656	4667	1	3	5	7	10
·67	4677	4688	4699	4710	4721	4732	4742	4753	4764	4775	1	3	5	8	10
·68	4786	4797	4808	4819	4831	4842	4853	4864	4875	4887	1	3	6	8	10
·69	4898	4909	4920	4932	4943	4955	4966	4977	4989	5000	1	3	6	8	10
·70	5012	5023	5035	5047	5058	5070	5082	5093	5105	5117	1	4	6	8	11
·71	5129	5140	5152	5164	5176	5188	5200	5212	5224	5236	1	4	6	8	11
·72	5248	5260	5272	5284	5297	5309	5321	5333	5346	5358	1	4	6	9	11
·73	5370	5383	5395	5408	5420	5433	5445	5458	5470	5483	1	4	6	9	11
·74	5495	5508	5521	5534	5546	5559	5572	5585	5598	5610	1	4	6	9	12
·75	5623	5636	5649	5662	5675	5689	5702	5715	5728	5741	1	4	7	9	12
·76	5754	5768	5781	5794	5808	5821	5834	5848	5861	5875	1	4	7	9	12
·77	5888	5902	5916	5929	5943	5957	5970	5984	5998	6012	1	4	7	10	12
·78	6026	6039	6053	6067	6081	6095	6109	6124	6138	6152	1	4	7	10	13
·79	6166	6180	6194	6209	6223	6237	6252	6266	6281	6295	1	4	7	10	13
·80	6310	6324	6339	6353	6368	6383	6397	6412	6427	6442	1	4	7	10	13
·81	6457	6471	6486	6501	6516	6531	6546	6561	6577	6592	2	5	8	11	14
·82	6607	6622	6637	6653	6668	6683	6699	6714	6730	6745	2	5	8	11	14
·83	6761	6776	6792	6808	6823	6839	6855	6871	6887	6902	2	5	8	11	14
·84	6918	6934	6950	6966	6982	6998	7015	7031	7047	7063	2	5	8	11	15
·85	7079	7096	7112	7129	7145	7161	7178	7194	7211	7228	2	5	8	12	15
·86	7244	7261	7278	7295	7311	7328	7345	7362	7379	7396	2	5	8	12	15
·87	7413	7430	7447	7464	7482	7499	7516	7534	7551	7568	2	5	9	12	16
·88	7586	7603	7621	7638	7656	7674	7691	7709	7727	7745	2	5	9	12	16
·89	7762	7780	7798	7816	7834	7852	7870	7889	7907	7925	2	5	9	13	16
·90	7943	7962	7980	7998	8017	8035	8054	8072	8091	8110	2	6	9	13	17
·91	8128	8147	8166	8185	8204	8222	8241	8260	8279	8299	2	6	9	13	17
·92	8318	8337	8356	8375	8395	8414	8433	8453	8472	8492	2	6	10	14	17
·93	8511	8531	8551	8570	8590	8610	8630	8650	8670	8690	2	6	10	14	18
·94	8710	8730	8750	8770	8790	8810	8831	8851	8872	8892	2	6	10	14	18
·95	8913	8933	8954	8974	8995	9016	9036	9057	9078	9099	2	6	10	15	19
·96	9120	9141	9162	9183	9204	9226	9247	9268	9290	9311	2	6	11	15	19
·97	9333	9354	9376	9397	9419	9441	9462	9484	9506	9528	2	7	11	15	20
·98	9550	9572	9594	9616	9638	9661	9683	9705	9727	9750	2	7	11	16	20
·99	9772	9795	9817	9840	9863	9886	9908	9931	9954	9977	2	7	11	16	20

Note: **Hyperbolic logarithms = 2·3026 × common logarithms**

COPPER WIRE TABLE

S.W.G.	Size—bare: Diameter in.	Diameter mm.	Section area in.²	Section area mm.²	Weight (lbs.) per 1000 yds.: Enam.	Enam. S.S.C.	Lewmex M.
10	·128	3·2512	·01287	8·3020	151	—	—
12	·104	2·6416	·00850	5·4805	100	—	—
14	·080	2·0320	·00503	3·2430	59·00	—	59·00
16	·064	1·6256	·00322	2·0755	37·69	38·07	37·73
18	·048	1·2192	·00181	1·1675	21·22	21·43	21·30
20	·036	·9144	·001018	·6567	11·96	12·10	12·03
21	·032	·8128	·000804	·5189	9·467	9·591	9·518
22	·028	·7112	·000616	·3973	7·257	7·329	7·303
23	·024	·6096	·000452	·2919	5·341	5·406	5·382
24	·022	·5588	·000380	·2453	4·494	4·560	4·531
25	·020	·5080	·000314	·2027	3·719	3·755	3·775
26	·018	·4572	·000254	·1642	3·016	3·069	3·049
27	·0164	·4166	·000211	·1363	2·504	2·554	2·536
28	·0148	·3759	·000172	·1110	2·043	2·089	2·086
29	·0136	·3454	·000145	·0937	1·726	1·763	1·753
30	·0124	·3150	·000121	·0779	1·436	1·474	1·459
32	·0108	·2743	·0000916	·0591	1·090	1·222	1·111
34	·0092	·2337	·0000665	·0429	·792	·805	·809
36	·0076	·1930	·0000454	·0293	·543	·565	·554
38	·0060	·1524	·0000283	·0182	·339	·358	·347
40	·0048	·1219	·0000181	·0117	·217	·232	·222

Length/ohm		*Working current*	A.W.G. +
British	*Metric*		
535 yds.	489 m.	12·87 Amp.	8
353	324	8·50	10
209	191	5·03	12
134	122·5	3·22	14
75·3	68·72	1·81	17
42·4	38·75	1·02	19
33·4	30·53	804 mA.	20
25·6	23·40	616	21
18·8	17·18	452	22
15·8	14·44	380	23
13·1	11·97	314	24
10·6	9·688	254	25
9·08	8·339	211	26
7·18	6·563	172	27
6·05	5·540	145	28
5·08	4·643	121	28
3·82	3·491	92	29
2·77	2·532	67	31
1·89	1·728	45	33
1·18	93·05 cm.	28	34
27·0 in.	68·58	18	36

Wire Tables and Resistance Wires

	‡*Turns/inch*		‡*Turns/cm*		*Resistance wire (ohms/yd)*		
S.W.G.	*Enam. or lewmex*	*Enam. S.S.C. or D.S.C.*	*Enam. or lewmex*	*Enam. S.S.C. or D.S.C.*	*Eureka @ 15·5°C.*	*Ni-chrome @ 500°C.*	*Man-gain @ 15·5°C.*
10	7·5	7·4	2·95	2·91	·054	—	—
12	9·2	9·1	3·62	3·58	·082	—	—
14	12·0	11·8	4·72	4·64	·138	·31	·123
16	14·7	14·5	5·79	5·71	·216	·50	·198
18	19·5	19·1	7·68	7·52	·384	·89	·311
20	25·7	24·8	10·12	9·76	·682	1·59	·605
21	28·8	27·6	11·34	10·87	·863	1·98	·736
22	32·5	31·2	12·78	12·28	1·13	2·61	1·03
23	37·5	36	14·77	14·77	1·53	3·55	1·40
24	41	39	16·14	15·35	1·83	4·22	1·62
25	45	42	17·72	16·54	2·21	5·13	1·96
26	50	47	19·68	18·51	2·73	6·30	2·48
27	54	51	21·25	20·07	3·29	7·62	3·00
28	60	56	23·62	22·04	4·04	9·32	3·53
29	65	60	25·59	23·62	4·78	11·0	4·17
30	70	66	27·66	25·98	5·75	13·3	4·85
32	80	74	31·50	29·13	7·58	17·5	6·73
34	93	85	36·61	33·47	10·4	24·2	9·27
36	112	99	44·10	38·97	15·3	35·4	14·08
38	140	117	55·12	46·07	24·6	57·7	21·80
40	172	137	67·71	53·94	38·4	88·5	34·45
42	206	155	81·00	61·02	55·3	130	48·68

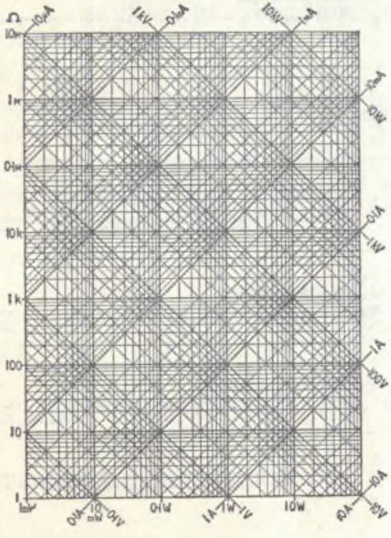
‡Average only: variations occur in size of wire and thickness of insulation

Resistance

Ohms law

The current in a d.c. circuit is directly proportional to the applied voltage and inversely proportional to the resistance of the circuit. $I = E/R$

Power, in watts, in d.c. circuit is given by the product of voltage and current. $W = E \times I$. Combining this formula with Ohm's law, gives also: $W = E^2/R$, or $W = I^2R$ The chart combines the quantities R, E, I & W, so that, if any two are known, the remaining two can be found.



CIRCUIT FORMULÆ

Resistors in series, $R_T = +R_1 + R_2 + R_3 \quad + \ldots$ etc.
Resistors in parallel, $1/R_T = 1/R_1 + 1R_2 + 1/R_3 + \ldots$ etc.

Star or delta transformation

If a network has three terminals, then, no matter how complicated, it will resolve into a star or delta.

Star to delta	Delta to star
$Ra = R_1 + R_3 + R_1R_3/R_2$	$R_1 = RaRc/(Ra + Rb + Rc)$
$Rb = R_2 + R_3 + R_2R_3/R_1$	$R_2 = RbRc/(Ra + Rb + Rc)$
$Rc = R_1 + R_2 + R_1R_2/R_3$	$R_3 = RaRb/(Ra + Rb + Rc)$

Resistance of materials

$R = \rho l/A$, ρ specific resistance, l length, A cross sectional area of conductor.

Material	*Resistivity, ρ, at 0° C.*		*Resistivity relative to copper*	*Temperature coefficient*
	Microhms /cm.cub	*Ohms/ circ. mil. ft.*		
Aluminium	2·62	15·75	1·65	·0042
Constantan	49·0	294·0	30·8	·00002
Copper (standard)	1·59	9·56	1·0	·0043
Copper (hard drawn)	1·60	9·62	1·02	·0041
Eureka	48·0	288·0	30·0	·00004
Gold	2·20	13·23	1·38	·0037
Lead	19·8	118·8	12·5	·0041
Mercury	94·1	565·2	59·2	·00086
Nickel (drawn wire)	9·9	59·5	6·30	·0039
Nichrome	109·0	657·0	68·5	·00015
Platinum (drawn)	11·0	66·2	6·92	·0037
Silver	1·47	8·84	·924	·0040
Steel (hard)	45·6	274·0	28·7	·0016
Tungsten (drawn)	5·42	32·6	3·41	·0051
Zinc	5·38	32·3	3·38	·0040

Resistance varies with temperature within 110°/–85°C. according to $Rt = Ro(1 + \alpha(\tau - 25))$. Rt is resistance at T°C., Ro reference resistance at 25°C., α the temperature coeff. $= (\Delta R/Ro)/\Delta T$.

Carbon resistors

α is —ve, i.e. R falls as T rises; –·006 to –·00012/°C.

R falls with applied voltage about –0·5/%100 volts.

Skin effect at H.F.

Due to changing magnetic field when passing a.c. current crowds to the conductor surfaces.

Skin depth of 36·8% ($1/\varepsilon$) of surface current at $d = 5033\sqrt{\rho/\mu f}$ mc.; ρ in ohms/cm. cube, μ permeability, f in Hz. $R_{HF} = \rho/dP\ \Omega$/cm, where P is perimeter in cm. In copper, $d = 6{\cdot}62\sqrt{f}$, $R = 261\sqrt{f}10^{-9}/P$.

Resistor colour code

Resistors are marked by a colour code painted on in either of two methods. (i) The first colour (A) occupies the body, the second colour (B) forms the tip or end, the third colour (C) is in the form of either a dot or band around the centre.

Colour	no.
Black	0
Brown	1
Red	2
Orange	3
Yellow	4
Green	5
Blue	6
Violet	7
Grey	7
White	9

(ii) Three bands of the appropriate colour may be painted as three successive rings placed to one end of the resistor.

A fourth band (D) of gold or silver denotes tolerance of 5% or 10% respectively. If no fourth band, the tolerance is –20%, e.g. an all brown resistor would be 110 Ω, a blue/grey/red 6·8 K, a red/black/green 2M Ω.

Preferred values

These values follow a roughly logarithmic scale at 20% increments thus: 10, 12, 15, 18, 22, 27, 33, 39, 47, 56, 68, 82 —repeating in multiples of 10. e.g., 470, 4,700, 47,000.

Inductance

When current in a conductor changes, the magnetic field associated with it also changes creating a *back e.m.f.* which opposes the original change of current. The value of the back e.m.f. created by unit rate of change of current is a measure of the self-inductance, or inductance (L) of the circuit. The unit is the henry, which is the inductance of a circuit producing a back e.m.f. of 1 volt when the current changes by 1 ampere in 1 second.

Mutual inductance

The changing magnetic field in one circuit may induce an e.m.f. into a neighbouring circuit; the amount of e.m.f. produced depends on the mutual inductance (M) between the circuits. It is defined similarly to self-inductance, viz., two circuits have a mutual inductance of 1 henry when a change of current of 1 ampere per second in one circuit induces an e.m.f. of 1 volt into the other.

Induced voltage expressions

Self inductance, $e = -L(di/dt)$ volts.
mutual inductance $e_2 = -M/di/dt)$ volts.

$M = K\sqrt{L_1L_2}$, where K the coupling coeff. is (total flux due to i_1)/(flux producing e_2).

Energy stored in magnetic fields is $T = LI^2/2$ joules.

Inductances in series

Inductors connected in series without their magnetic fields affecting each other have effective inductance, $Leff. = L_1 + L_2$. If their fields do interact, they possess a mutual inductance, M, the fields may either *aid* or *oppose* each other. When the fields aid, the total inductance, $Leff. = L_1 + L_2 + 2M$. Where they oppose, $Leff. = L_1 + L_2 + 2M$.

Measurement of mutual inductance

From the two previous equations, if L_A and L_O = total inductance when aiding and opposing respectively, it follows that $M = (L_A - L_O)/4$, and this provides a convenient method of finding M. The coils are first connected up in one way and their total inductance measured; then the connection to one coil is reversed, and the measurement made again. This gives L_A and L_O, and hence M.

Inductance in circuits

In circuits containing L and R, the current does not rise instantaneously to Ohms law value of E/R amps because of the back e.m.f. e. The current found by solving $(E-L\,di/dt) = Ri$, viz. $i = (1-\varepsilon^{Rt/L})$. E/R amps. At $t = L/R$ secs., $i = 63\%\ E/R$. If E volts a.c. at frequency f cycles/sec. is applied, the back e.m.f. opposes E, this opposition to a.c. flow is termed *inductive reactance* (X_L) If $i = I \sin 2\pi ft$. then $e = -L(di/dt) = -2\pi f LI \cos 2\pi ft$. volts. Thus, $X_L = (e/i) = 2\pi fL/90°$, where /90° indicates i lags V by 90° phase, by virtue of which power cannot be dissipated in a pure inductor.

CALCULATION OF INDUCTANCE

The inductance of straight wires and various coil formations may be calculated from a number of formulae, all of which demand some recourse to constants which have been derived empirically. Few formulae give accurate results for coils used on vhf, due to self-capacitance and skin effects.

In the formulae given below (for air-cored coils) the dimensions are in *centimetres* and the inductance is in *microhenrys*.

μ is the permeability of the conductor, $=1$ (except for iron.)

δ is a factor, between 0 and 0·25 which is dependent upon frequency and wire diameter. Its value may be deduced approximately from the following table, where $x = 0{\cdot}1d - f$.

x	0	2	5	10	20	50	100	∞
δ	·25	·24	·14	·07	·035	·014	·007	0

Other symbols used are: d = diam. of wire: l = length of wire (or coil); D = distance between wires; r = radius of coil. In all cases of straight wires, the formulae have been simplified by assuming l/d very large.

Round straight wires. $L = 0{\cdot}002l[2{\cdot}303 \log_{10}(4l/d) - 1 + \mu\,\delta]$.

Two parallel wires. Round section, $L = 0{\cdot}004l[2{\cdot}303 \log_{10}(2D/d) - (D/l) + \mu\,\delta]$. If the two wires are not of the same dimensions, each must be calculated separately, and combined by the formula, $L = L_1 + L_2 \pm 2M$.

Single circular turn of round wire. $L = 0{\cdot}0126r(2{\cdot}303 \log_{10}(16r/d) - 2 + \mu\,\rho)$, provided $d/2r << 0{\cdot}2$.

Single square turn of round wire. $L = 0{\cdot}008s[2{\cdot}303 \log_{10}(2s/d) - 0{\cdot}75 + \mu\,\rho]$, (s = side of square).

Single-layer coil of round wire (solenoid). $L = 4\pi^2r^2N^2K/l \times 10^3 = 0{\cdot}0395r^2N^2K/l$ where N = number of turns, and K is a constant dependent upon the ratio $2r/l$ and given in the table below.

$2r/l$	K	$2r/l$	K	$2r/l$	K	$2r/l$	K
·00	1·0000	·32	·8767	1·00	·6884	2·5	·4719
·02	·9916	·34	·8699	1·05	·6777	3·0	·4292
·04	·9832	·36	·8632	1·10	·6673	3·5	·3944
·06	·9750	·38	·8565	1·15	·6573	4·0	·3654
·08	·9668	·40	·8499	1·20	·6475	4·5	·3409
·10	·9588	·45	·8337	1·25	·6381	5·0	·3198
·12	·9509	·50	·8181	1·30	·6290	6·0	·2854
·14	·9430	·55	·8031	1·35	·6201	7·0	·2584
·16	·9353	·60	·7885	1·40	·6115	8·0	·2366
·18	·9276	·65	·7745	1·45	·6031	9·0	·2185
·20	·9201	·70	·7609	1·50	·5950	10·0	·2033
·22	·9126	·75	·7478	1·60	·5795	15·0	·1527
·24	·9053	·80	·7351	1·70	·5649	20·0	·1236
·26	·8980	·85	·7228	1·80	·5511	30·0	·0910
·28	·8909	·90	·7110	1·90	·5379	50·0	·0611
·30	·8838	·95	·6995	2·0	·5255	100	·0350

Another formula reasonably accurate, but not involving K is given by: $L = r^2N^2/(9r + 10l)$, where r and l are in inches.

To obtain maximum inductance from a given length of wire wound in the form of a solenoid, the diameter should be $2{\cdot}54 \times$ the length.

Inductance of multilayer coils of rectangular cross-section.

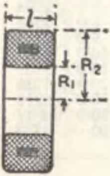
As for solenoids the inductance of a multi-layer coil is dependent upon the proportions of length, radius and depth of winding. A formula which is reasonably accurate for coils where $R/(t + l)$ and l/t are each not greater than 10:1, and which follows:

$$L = \frac{{\cdot}02303N^2(2R + t)\ (1\sqrt{1{\cdot}125t/R})}{(1 + 1{\cdot}15l/R)}$$

microhenry

The greatest inductance is given by a coil of square cross-section (i.e., $t = l$) and of as large a radius as possible.

Toroidal coils of rectangular and circular cross-section



Square section

$$L = {\cdot}0046lN^2\log_{10}(R_2/R_1)\mu$$

Circular section

$$L = {\cdot}0126N^2(R - \sqrt{R^2 - r^2})\mu$$

μ = permeability of the core material

Where the permeability, μ is high and the leakage is low, inductance is given by: $L = {\cdot}0126N^2\mu A/c$, where A is the cross-sectional area and c is the length of the magnetic path.

Capacitance

Definition

The capacitance of a conductor is defined as the quantity of charge required to raise its potential by one unit. If a quantity Q coulombs raises the potential by V volts, the capacitance is given by $C = Q/V$. The constant C is known as the capacitance and its unit is the farad. This unit is a very large one and for most practical purposes the unit used is one-millionth of a farad (a microfarad), or a millionth of a microfarad (a picofarad). A capacitor is a conducting surface whose capacitance has been artificially increased by bringing another conducting surface near to it. The capacitance depends upon the surface area of the conducting plates, the distance between them and the interposing dielectric, or insulating material between parallel plate capacitor. C = ·0885 Ks/d picofarads, where s = surface area (sq. cms.); d = distance between plates and K = dielectric constant (also called specific inductive capacitance, or permittivity). A table of values for K, for some of the commoner insulants, is given below. The power factor (cos ϕ) is also given, but is only approximate.

Electrolytic capacitors are polarised for varying d.c. only or non-polarised (back to back polar units) suitable for a.c. The volume efficiency is high but leakage currents can be ·002 to ·25 m/A per μF.

Capacitors in parallel

The total (effective) capacitance of a number of capacitors in parallel is given by the sum of the individual capacitances.

$$C_{eff} = C_1 + C_2 + C_3 + \ldots\ldots$$

Capacitors in series

$$1/C_{eff} = 1/C_1 + 1C_2 + 1/C_3 + \ldots\ldots$$

Energy of a charged capacitor

The work done in charging a capacitor is given by:

$W = CV^2/2$, in farads, V in volts and W in Joules.

Capacitance

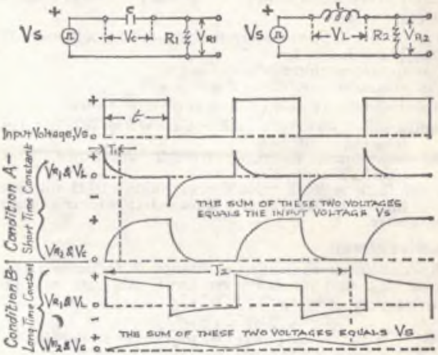
VALUES FOR K AND $\cos\phi$.

Material	K	$\cos\phi$
Air (at N.T.P.)	1·000	—
Ebonite	2·8	·006
Glass, crown	7·0	·007
Glass, flint	6·6	·01
Glass, plate	8·4	·01
Glass, Pyrex	4·9	·004
Gutta percha	4·2	·03
Gypsum	6·3	·002
Hydrogen	0·9998	—
Marble	9·3	·01
Mica	8·0	·00017
Oil, paraffin	2·7	·01
Oil, vaseline	2·0	·02
Paper (dry)	2 to 3	·04
Paraffin wax	2·3	·009
Pitch	1·8	·05
Cellulose acetate	3 to 7	·06
Nylon	3 to 4	·04
Paxolin 'T' Grade	4·9	·030 Av
Paxolin 'V' Grade	4·6	·028 Av
Penhol-formal-dehyde	5 to 20	·03
Polystyrine	2·5 to 3	·0004
Poly-vinyl-chloride	4 to 12	·03
Porcelain	6·5	·006
Quartz	4·5	·00015
Resin	3·3	·003
Rubber, pure	2·2	·05
Vulcanised	3·9	·03
Shellac	6·0	·07
Silica	3·6	·02
Slate	12	·55
Steatite	6·5	·001
Sulphur	3·0	—
Tufnol *(Kite Brand)	5·07	·038
Water (pure)	75	—
Wood, birch	5·2	·065
Wood, oak	3·3	·035
Wood, teak (oiled)	2·7	·015
Wood, whitewood	1·7	·025
Vacuum	·9994	—

*Tested at 1MHz

Time constant

Defined in (L,R) and (C,R) circuits as $T = L/R$ and $T = CR$ seconds respectively. These circuits modify input wave forms, the extent being gauged by comparing the period t of the input wave with the time constant. A short T is $< t/10$, a long $T > 10t$ and the effect on a square wave is shown below.



Case A is a short T, when C charges completely well within t producing V_{R1} or V_L which is substantially $\propto dV_S/dt$ and acts as a differentiator circuit. Case B is a long T, C being unable to charge more than a few per cent of V_S within t, producing V_{R2} or V_C substantially $\propto \int V_S dt$ and acts as an integrator. Note that medium and long T circuits alter datum level with time; this permits a method of sorting different pulse widths, e.g. for $T = 40\mu S$.

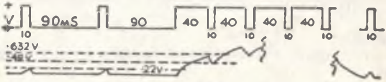


Fig 2

Alternating current

Waveform

Most alternating current or voltage phenomena follow a sinusoidal (sine wave-form) law, which is written as—

$$i = I_{max} \sin \omega t; \; e = E_{max} \sin \omega t.$$

where i and e are the instantaneous values of current or e.m.f. I_{max} and E_{max} are the maximum values of current or e.m.f. ω = angular velocity $= 2\pi f$; and t = time in seconds

Average values

For a sinusoidal waveform: $E_{av} = E_{max} \times 2/\pi = \cdot637 \, E_{max}$; $I_{av} = I_{max} \times 2/\pi = \cdot637 \, I_{max}$

For square waveforms: $E_{ac} = E_{max}$

For triangular waveforms: $E_{av} = E_{max}/2$

Root mean square value (virtual, or effective value)

Sinusoidal waveform: $E = E_{max}/\sqrt{2} = \cdot707 E_{max}$ $I = I_{max}/\sqrt{2} = \cdot707 \, I_{max}$

Sq. waveforms, $E = E_{max}$; triangle waveforms, $E = E_{max}/\sqrt{3}$

Form factor = *RMS* value/average value = 1·111 for sine wave = 1·000 for square wave = 1·1574 for triangular wave.

Series circuit

A series circuit containing resistance R, inductive reactance, X_L, and capacitive reactance, Xc, has a total IMPEDANCE, $Z = \sqrt{R^2 + (X_L - Xc)^2} =$

$$\sqrt{R^2 + (2\pi fL - 1/2fC)^2}.$$

Power factor = true power (I^2R)/apparent power $(I^2Z) = R/Z = \cos\phi$. Phase angle, $\phi = \tan\text{-i}(X\text{u} - Xc)/R$.

Parallel circuit

The simplest method of calculating parallel networks is to turn resistance, reactance and impedance into their reciprocals of conductance $(G = 1/R)$, subsceptance $(b = 1/X)$ and admittance $(Y = 1/Z)$ and add these reciprocal quantities vectorially as for series networks, i.e., $Y = \sqrt{G^2 + b^2}$

Power factor $= G/Y = Z/R$: phase ang. $\phi = \tan^{-1} b/G = \tan^{-1} R/X$

Resonance

In a series circuit, the condition for resonance occurs when $X_L = Xc$.

$$\therefore 2\pi fL = 1/2\pi fC; \text{ and } f = 1/2\pi\sqrt{LC}.$$

In the series circuit, known as the *acceptor circuit*, the impedance, $Z_r = R$ at resonance, and maximum current occurs.

In the parallel circuit, know as the *rejector circuit*, the combination offers infinite impedance if no resistance is present. In practice, resistance must occur and the effective, or Dynamic resistance, of a tuned circuit at resonance is given by: $R_{dyn} = L/CR$, where R_{dyn} is equivalent series resistance.

Q Factor. Practical inductors must have resistance, which prevents the tuned circuit presenting infinite or zero impedance in the case of parallel and series circuits respectively. The ratio of inductive reactance to the resistance of an inductor is a measure of its *goodness* and is known as the Q factor, $= 2\pi f_0 L/R$, where f_0 is the resonant frequency of the circuit in which it is used. It must be remembered that the resistance R is the effective resistance at the frequency involved, and, due to skin effect, may be considerably higher than its d.c. value. $R_{dqn} = QX_0 = Q^2R$.

When the Q of a parallel circuit is below 10, the term resonance is not so easily defined—there is a set of values for L and C that will make the parallel impedance a pure resistance, but with these values the impedance does not have its maximum possible value. Another set of values for L and C will make the parallel impedance a maximum, but this maximum value is not a pure resistance. Either condition could be called *resonance*, so with low Q circuits it is necessary to distinguish between *maximum impedance* and *resistive* parallel resonance.

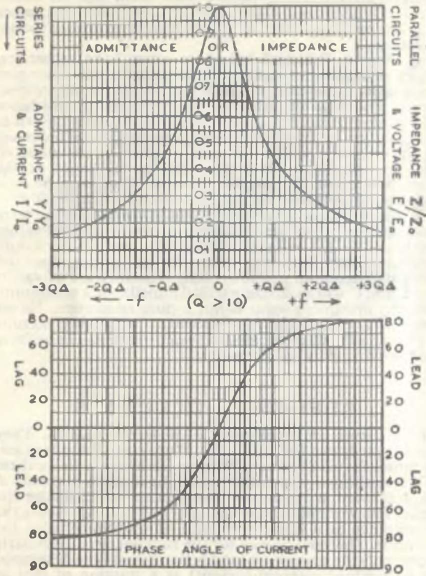
RESONANT CIRCUITS

These curves are applicable to all tuned circuits. They have been drawn for a Q of 30 but are correct to graphical accuracy for all Q's above 10 or so. In order to make them of general application their scales have been *normalised*, i.e. expressed in parameters independent of particular circuits. The vertical axis for the series circuit represents the admittance as a fraction of that at resonance (also the current). The vertical axis for the parallel circuit similarly represents the impedance (and hence the voltage developed by a constant external current) as a fraction of that at resonance.

For a series circuit, $Y_0 = G = 1/R$, or $Z_0 = R$.

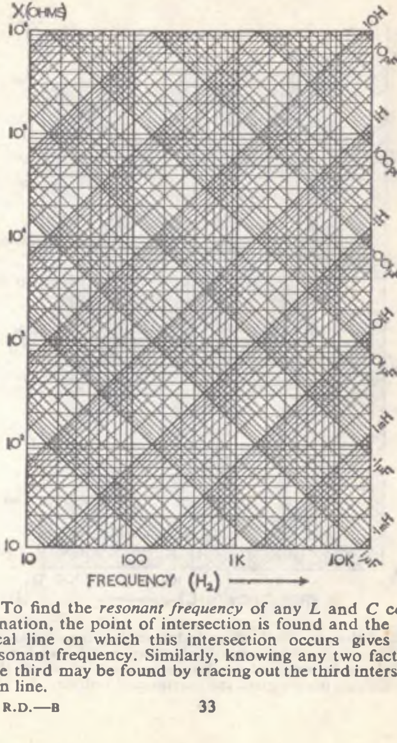
For a parallel circuit, $Z_0 = L/CR$, if the Q is fairly high.

The frequency axis is expressed in terms of the fractional de-tuning Δ ($= \delta f/f_0$) multiplied by Q Hence, increasing the Q for a given fractional de-tuning reduces the value of admittance (or impedance) read off from the curve for that amount of de-tuning corresponding to greater selectivity.



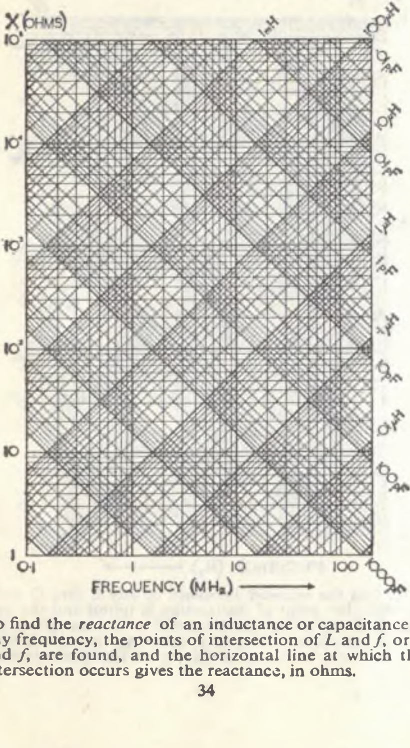
Nomographs for admittance, impedance and phase angle

Reactance and Resonance Chart for Audio Frequencies



To find the *resonant frequency* of any *L* and *C* combination, the point of intersection is found and the vertical line on which this intersection occurs gives the resonant frequency. Similarly, knowing any two factors, the third may be found by tracing out the third intersection line.

Reactance and Resonance Chart for Radio Frequencies



To find the *reactance* of an inductance or capacitance at any frequency, the points of intersection of L and f, or C and f, are found, and the horizontal line at which this intersection occurs gives the reactance, in ohms.

Calculation of Q

Provided that $Q > 10$ and that $\delta f < fo/20$ the impedance presented by a tuned circuit may be expressed

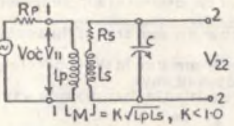
$$|z| = R\sqrt{1+(Q2\ \delta f/fo)^2} \text{ for series}$$

and $|z| = Rdyn/\sqrt{1+(Q2f\ \delta/fo)^2}$ for parallel circuits. Voltage (E) across a tuned circuit δf cycles off tune relative to the voltage at resonance (Eo), is

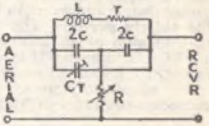
$$E/Eo = 1/\sqrt{1+(Q2\ \delta f/fo)^2}$$

and when $Q = fo/2\ \delta f$, $E = (1/\sqrt{2})\ Eo$. $E = \cdot 707\ Eo$.

R. F. coupling



The tuned circuit together with an R.F. transformer is a common form of coupling. The maximum possible gain at fo is given by $V_{22}/V_{11} = \omega Ls/2\sqrt{RpRs}$ (assuming coefficient of coupling $K = 1{\cdot}0$). If C, Ls and Rp be fixed, then the optimum Lp for this gain is $LpOPT. = Rp/2\pi foQs$. If $Lp\ OPT.$ is employed Qs is reduced and the selectivity degraded. Up to 30MH$_z$ it is usual to put $Lp = 1/36\ Lp\ OPT.$ giving 40% maximum gain but 95% maximum Qs. attenuate an undesired frequency a wave trap is often employed, being a parallel circuit in series with the receiver aerial input or more effective a bridged-T network as below:



Star network ($2C$, $2C$, R) converts to delta equivalent. L now paralleled by capacity $C^1 = C$ in series with negative resistance $-R^1$. At $f = 1/2\pi\sqrt{L.(C+C_T)}$ resistance $-R^1$ depends on R, and cancels out r (self resistance of L) when $R = Rdyn/4$, where $Rdyn = L/Cr$. Procedure is

(a) tune C_T

(b) then adjust R for max., attenuation with receiver tuned to undesired signal.

Sound

Sound is the transfer of energy, in the form of longitudinal waves of compression and rarefaction, from one part of a conducting medium to another. At a temperature of 0°C, a sound wave travels at 1087 feet per second, in air. As the temperature rises, the velocity of the sound increases—at a rate of 2 feet per second per degree centigrade increase in temperature.

The chief properties of sound are:

Loudness. The subjective sensation produced on the human ear.

Pitch. The impression that denotes the comparative frequency of the fundamental (or lowest tone) of a complex sound.

Timbre. The tone quality, as determined by the rate of rise and decay of the sound and the number, frequencies and relative amplitudes of the overtones superimposed on the fundamental frequency.

Intensity. The acoustical power per unit area of the sound conducting medium.

Pressure. The r.m.s. change of pressure in the conducting medium in the path of the sound wave.

Intensity and pressure are related in the following way:

$$\text{intensity, } J = \frac{P^2}{R_r}$$

where P is the pressure and R_r is the radiation resistance.

Human beings vary with age in their range of audibility. The lowest frequency audible as a note is about 16 H_z., but the highest note varies from over 15,000 H_z for very young people and decreasing with age to 8,000 or 10,000 H_z for elderly persons. The ideal upper limit for perfect transmission of all audible sounds would therefore be 15,000 H_z, but 10,000 is considered as perfectly satisfactory for high quality reproduction.

For satisfactory speech transmission (telephone communication) a range of 200 to 2,600 H_z is sufficient.

MUSICAL SCALES

Notes of the gamut	*C*	*D*	*E*	*F*	*G*	*A*	*B*	*C*
Frequency ratios, based on C = 1	1	$\frac{9}{8}$	$\frac{5}{4}$	$\frac{4}{3}$	$\frac{3}{2}$	$\frac{5}{3}$	$\frac{15}{8}$	2

Interval ratios between successive notes	$\frac{9}{8}$	$\frac{10}{9}$	$\frac{16}{15}$	$\frac{9}{8}$	$\frac{10}{9}$	$\frac{9}{8}$	$\frac{16}{15}$

Intervals, $\frac{9}{8}$ and $\frac{10}{9}$ are whole tones; $\frac{16}{15}$ is a semitone.

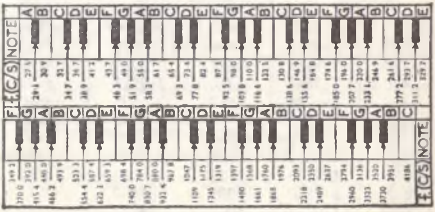
Other intervals, and their frequency ratios, are:

2/1, e.g. from C' to C, is an *octave*
3/2, e.g. from G to C, is a *fifth*
4/3, e.g. from C' to G, is a *fourth*
5/4, e.g. from A to F, is a *third*

The above refers to the true diatonic scale, and enables one to calculate the frequency of any note if one note, say C is fixed. Thus, if middle C is given the (old) physical pitch of 256 cycles/sec. the corresponding A above it would be $256 \times \frac{5}{3} = 426{\cdot}67$ Hz.

It is not possible to tune a keyed instrument (e.g. a piano or organ) to the diatonic scale exactly, so the *equal tempered scale* has been evolved in which every interval has the same frequency ratio. This ratio is: $\sqrt[12]{2} = 1{\cdot}05946$ . . . a complete octave of ratio 2/1 being effected in twelve notes.

Orchestral pitch has now been fixed by international agreement, as: $A' = 440{\cdot}00$ Hz. From this, the piano keyboard frequencies are as shown below.



DECIBEL CALCULATIONS

Power ratios

$N(\text{decibels}) = 10\log_{10}P_2/P_1$. $N(\text{nepers}) = \cdot 5\log_\varepsilon P_2/P_1$

Voltage and current ratios

$N(db) = 20\log_{10}E_2/E_1$. $N(db) = 20\log_{10}I_2/I_1$

$N(\text{nepers}) = \log_\varepsilon E_2/E_1$. $N(\text{nepers}) = \log_\varepsilon I_2/I_1$

db	*Voltage ratios*		*Power ratios*	
	E_2/E_1	E^1/E_2	P_2/P_1	P_1/P_2
0·1	1·0116	·9885	1·0223	·9772
0·2	1·0233	·9772	1·0471	·9550
0·3	1·0315	·9660	1·0715	·9332
0·4	1·0471	·9550	1·0965	·9120
0·5	1·0593	·9441	1·1220	·8912
0·6	1·0715	·9332	1·1482	·8710
0·7	1·0839	·9226	1·1749	·8511
0·8	1·0965	·9120	1·2023	·8318
0·9	1·1092	·9016	1·2303	·8128
1·0	1·1220	·8912	1·2589	·7943
1·2	1·1482	·8710	1·3183	·7586
1·4	1·1749	·8511	1·3804	·7244
1·6	1·2023	·8318	1·4454	·6923
1·8	1·2303	·8128	1·5136	·6608
2·0	1·2589	·7943	1·5849	·6310
2·2	1·2882	·7762	1·6595	·6025
2·4	1·3183	·7586	1·7382	·5754
2·6	1·3490	·7413	1·8198	·5501
2·8	1·3804	·7244	1·9055	·5249
3·0	1·4125	·7079	1·9953	·5012
3·4	1·4789	·6761	2·1884	·4570
3·6	1·5136	·6608	2·2910	·4365
3·8	1·5488	·6457	2·3986	·4168
4·0	1·5849	·6310	2·5119	·3981
4·2	1·6218	·6167	2·6305	·3802
4·4	1·6595	·6025	2·7539	·3631
4·6	1·6982	·5888	2·8837	·3467
4·8	1·7382	·5754	3·0206	·3311
5·0	1·7783	·5623	3·1623	·3162
5·5	1·8836	·5309	3·5480	·2819
6·0	1·9953	·5012	3·9811	·2512

Sound

db	Voltage ratios E_2/E_1	Voltage ratios E_1/E_2	Power ratios P_2/P_1	Power ratios P_1/P_2
6·5	2·1134	·4732	4·4668	·2238
7·0	2·2387	·4467	5·0119	·1995
7·5	2·3712	·4218	5·6234	·1778
8·0	2·5119	·3981	6·3096	·1585
8·5	2·6605	·3757	7·0795	·1412
9·0	2·8184	·3548	7·9433	·1259
9·5	2·9851	·3350	8·9125	·1122
10·0	3·1623	·3162	10·000	·1000
11·0	3·5480	·2819	12·589	·07943
12·0	3·9811	·2512	15·849	·06310
13·0	4·4668	·2238	19·953	·05012
14·0	5·0119	·1995	25·119	·03981
15·0	5·6234	·1778	31·623	·03162
16·0	6·3096	·1585	39·811	·02512
17·0	7·0795	·1412	50·119	·01995
18·0	7·9433	·1259	63·096	·01585
19·0	8·9125	·1122	79·433	·01259
20·0	10·000	·1000	100·00	·01000
22·0	12·589	·07943	158·49	·00631
24·0	15·849	·06310	251·19	·00398
26·0	19·953	·05012	398·11	·00251
28·0	25·119	·03981	630·96	·00158
30·0	31·623	·03162	1000·0	·00100
32·0	39·811	·02512	1584·9	·00063
34·0	50·119	·01995	2511·9	·00039
36·0	63·096	·01585	3981·1	·00025
38·0	79·433	·01259	6309·6	·00016
40·0	100·00	·01000	10,000	·00010
42·0	125·89	·00794	15,849	·000063
44·0	158·49	·00631	25,119	·000039
46·0	199·53	·00501	39,811	·000025
48·0	251·19	·00398	63,096	·000016
50·0	316·23	·00316	10^5	·000010
60·0	1,000·0	·00100	10^6	10^{-6}
70·0	3,162·3	·00032	10^7	10^{-7}
80·0	10^4	·00010	10^8	10^{-8}
90·0	31,623	·00003	10^9	10^{-9}
100·0	10^5	·00001	10^{10}	10^{-10}

The relationship between power and voltage ratios is found since $N = 10^0 \log_1 E^2_2/R_2/E^2_1/R_1$ *db*., giving $N = 20 \text{long}_{10} E_2/E_1 + 10 \log R_1/R_2$ *db*. Usually the important relationship is that of the voltages, power being of less interest in some applications, and the last term above is omitted. It is convenient to work with ratios greater than unity; should $E_2 < E_1$, invert the ratio and precede the logarithm with a negative sign (indicating loss):

$$10 \log_{10} P_2/P_1 = -10 \log_{10} P_1/P_2 \; db$$

In addition to the relative *db* scales above there are two absolute *db* scales. One is the acoustic, zero level being taken at ·0002 dynes/cm^2 (pressure scale) or 10^{-16} watts/cm^2 (intensity scale), giving the *db* level of any pressure Px dynes as, $db = 20 \log_{10} Px/{\cdot}0002$. The other is the absolute electrical (*dbm*.) scale, zero level being $1mW$, any power Wx having a *dbm* level, $dbm = 10 \log_{10} (Wx/10^{-3})$. A standard communication impedance of 600Ω is assigned to this scale giving a corresponding voltage scale, the zero level when $1mW$ is dissipated in 600Ω being ·775 volts. The *db* level of any voltage Vx on this scale is $db = 20 \log_{10} (Vx/{\cdot}775)$. In television an absolute scale with a zero of 1 volt into 75Ω may be quoted.

Relationship between intensity and distance

The energy contained in a single sound wave remains practically constant and, since each wave is distributed over a spherical surface which is enlarging rapidly, the energy passing through each unit area of a wave surface depends upon the distance of that surface from the source of the sound. Because the area of a sphere varies directly as the square of the radius ($a = \pi r^2$), the energy of the wave at a distance of 2 metres from the source, for example, will be distributed over an area four times as great as when the wave was at a distance of 1 metre. The intensity of the sound therefore varies as the inverse square of the distance from the source (when no solid object in the vicinity causes absorption or reflection).

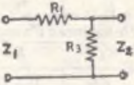
As the intensity of a note decreases, so does the loudness, *but not in step*, The difference in loudness is roughly proportional to the differences in the logarithm, to the base ten, of their relative intensities.

Useful networks

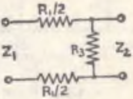
ATTENUATOR DESIGN

In all cases, the voltage ratio, *V*, corresponding to the number of decibels loss, *N db*, must first be found; i.e., $V = \text{antilog}(N/20)$.

Potentiometer, or L type



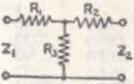
Unbalanced



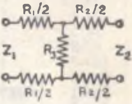
Balanced

$$R_3 = 2VZ_1Z_2/(V^2Z_2 - Z_1)$$
$$R_1 = Z_1(V^2Z_2 + Z_1)/(V^2Z_2 - Z_1) - R_3$$

T and H types



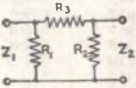
Unbalanced



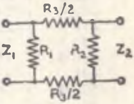
Balanced

$$R_3 = 2VZ_1Z_2/(V^2Z_2 - Z_1)$$
$$R_1 = Z_1(V^2Z_2 + Z_1)/(V^2Z_2 - Z_1) - R_3$$
$$R_2 = Z_2(V^2Z_1 + Z_2)/(V^2Z_1 - Z_2) - R_3$$

π and square (or box) types



Unbalanced

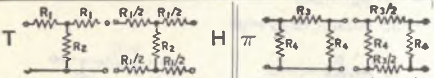


Balanced

$$R_3 = (V^2Z_2 - Z_1)/2V$$
$$R_1 = Z_1(V^2Z_2 - Z_1)/(V^2Z_2 + Z_1 - 2VZ_1)$$
$$R_2 = Z_2(V^2Z_2 - Z_1)/(V^2Z_2 + Z_1 - 2VZ_2)$$

ATTENUATOR TABLES FOR $Z_1 = Z_2 = Z = 600\Omega$

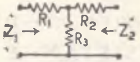
For other values of Z, multiply all resistors by Z/600



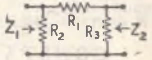
Unbalanced *Balanced* *Unbalanced* *Balanced*

db	R_1	R_2	R_3	R_4
·1	3·46	52·10*k*	6·9	104·2*k*
·2	6·90	26·06*k*	13·8	52·12*k*
·3	10·36	17·38*k*	20·7	34·75*k*
·4	13·82	13·02*k*	27·6	26·06*k*
·5	17·26	10·42*k*	34·5	20·87*k*
·6	20·72	8·68*k*	41·5	17·38*k*
·7	24·17	7·44*k*	48·4	14·90*k*
·8	27·62	6·50*k*	55·3	13·04*k*
·9	31·1	5·78*k*	62·2	11·60*k*
1·0	34·5	5·21*k*	68·6	10·43*k*
2	68·8	2·58*k*	139·4	5232
3	107·7	1·70*k*	212·5	3505
4	135·8	1·25*k*	287·5	2651
5	168·1	987·6	264·5	2141
6	199·3	803·4	447·5	1807
7	229·7	685·2	537·0	1569
8	758·4	567·6	634·2	1393
9	285·8	487·2	738·9	1260
10	310·0	421·6	854·1	1154
11	336·1	367·4	979·8	1071
12	359·1	321·7	1119	1002
13	380·5	282·8	1273	946·1
14	400·4	249·4	1443	899·1
15	418·8	220·4	1632	859·6
16	435·8	195·1	1847	826·0
18	466·5	152·5	2344	772·8
20	491	121·2	2970	733·3
25	536·4	67·7	5318	670
30	563	37·99	9500	639
35	579	21·36	16·86*k*	622
40	588	12·00	30·00*k*	612
45	594	6·74	53·35*k*	607
50	596	3·79	94·87*k*	604

Attenuator Pads to Work Between Unequal Impedances



T *pad (unbalanced)*



π pad (unbalanced)

		T pad			π pad		
Z_1/Z_2	Loss	R_1	R_2	R_3	R_1	R_2	R_3
600/500	5*db*	256	64	902	332	4709	1169
	10	282	170	385	780	1764	1067
	20	502	400	111	2712	761	606
	30	566	466	35	4652	643	530
	40	590	490	11	26382	613	509
600/250	10	394	6	272	550	35080	382
	20	534	178	78	1916	849	281
	30	576	226	25	6120	664	260
	40	592	242	8	19348	619	253
600/100	20	586	52	50	1212	1143	107
	30	592	84	16	3872	708	102
	40	598	96	5	12248	631	101
600/30	20	562	4	27	664	5138	31
	30	586	22	9	2108	835	30
	40	596	28	3	6700	659	30
250/30	20	238	12	18	428	572	32
	30	244	24	6	1368	305	31
	40	248	28	2	4330	265	30
100/30	20	90	18	11	272	153	33
	30	94	24	6	886	113	31
	40	98	30	1	2750	104	30
50/30	10	34	10	27	56	160	44
	20	44	24	8	192	70	36
	30	48	28	3	598	54	31
	40	50	30	0·8	1936	51	30

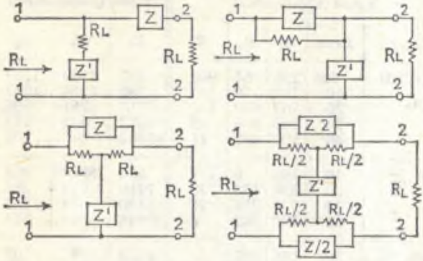
Attenuators between matched impedances, Z, T and H types:

$R_3 = 2VZ/(V^2 - 1)$: $R_1 = R_2 = Z(V - 1)/(V + 1)$

π and square types:

$R_3 = Z(V^2 - 1)/2V$: $R_1 = R_2 = Z(V + 1)/(V - 1)$

CONSTANT RESISTANCE NETWORKS

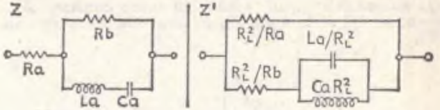


A network designed to alter the frequency response of a system (equaliser network) contains reactors and will in general present a varying input impedance (Z_{11}) with frequency. If, however, a constant resistance structure is employed Z_{11} is constant and equal to the terminating resistor, R_L. Such structures are as shown above.

The necessary condition is that $ZZ^1 = RL^2$; if Z, Z^1 meet this condition they are *inverse* w.r.t. R_L. Inverse relationships are:

Z	R	L	C	SERIES Za Zb
Z'	R^2/R_L	L/R_L^2	CR_L^2	PARALLEL Z'a Zb'

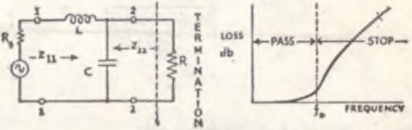
A typical example is shown below:



The loss in all these structures is 20 $\log_{10}$ $(1+R_L/Z^1)$ or 20 $\log_{10}$ $(1+Z/R_L)db$.

FILTERS

The basic 1/2 section low pass filter and loss curve:

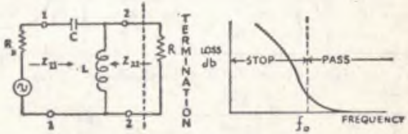


Procedure

1 given termination R, cut-off frequency fo, obtain c from $c=1/2\pi fo\ R$

2 $L=CR^2$. For $f>fo$ loss is approximate $6+20$ log. f/fo db.

The basic 1/2 section high pass filter and loss curve:



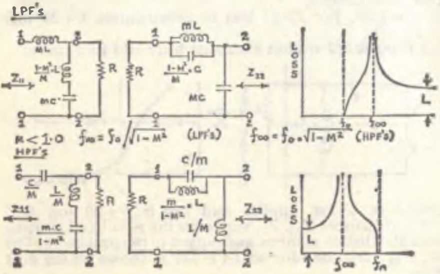
Procedure above applies and loss is $6+20$ log fo/f. The impedances Z_{11}, Z_{22} vary over the pass band causing mismatch losses at input and output in the pass band. The Z_{11}, Z_{22} variations for an LPF are as shown on the next page.

The process of M derivation reduces this variation. This process develops from the basic 1/2 section; series M 1/2 section has Z_{11} of basic, but more constant Z'_{22}; shunt M 1/2 section has Z_{22} of basic but more constant Z'_{11}.

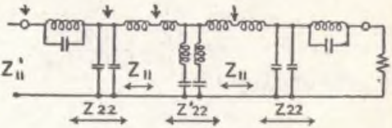
These 1/2 sections and loss curves are:

SERIES DERIVED

SHUNT DERIVED



Filter 1/2 sections may be cascaded to form a composite filter, *if similar impedances always face each other.* The tendency to constant loss L for an M derived 1/2 section can be overcome by cascading with basic 1/2 sections, the M sections going to either end of the composite filter to maintain a better match with source and load. The best M terminating section is shunt derived $M = \cdot 6$. Typical LPF:

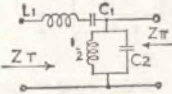


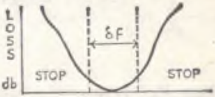
Bandpass filters

An analogy between LPF's and BPF's permits easy design approach viz. (i) design an LPF for desired R of BPF, with an $fo = \delta f$ the desired bandwidth of the BPF; this directly yields L_1 and C_2 values of BPF. (ii) from $fc = 1/2\pi\sqrt{LC}$ find C_1 resonant with L_1 also L_2 anti-resonant with C_2 at centre frequency fc.

Bandstop filters

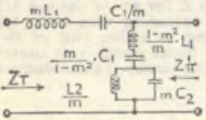
These are analogous with HPF's, stopwidth $\delta = fo$ of HPF; series arm anti-resonates, shunt arm resonant at fc.





$L_1 = R/2\pi\ \delta f$, $C_1 = \delta f/2\pi fc^2 R$, $C_2 = 1/2\pi\ \delta f R$, $L_2 = \delta f R/2\pi fc_1$

The series *M* derived BPF 1/2 section.



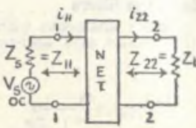
$$f\infty_1.f\infty_2=f_1f_2=fo^2 \qquad P=\sqrt{\frac{f_2}{f_1}-\frac{1-(f\omega_1/f_1)^2}{1-(f\omega_1/f_2)^2}}$$

$$m=(\sqrt{f_2/f_1}+\sqrt{f_1/f_2})/(P+1/P)$$

IMAGE OPERATION OF NETWORKS

If the network input and output impedances (Z_{11}, Z_{22}) match the source and load impedances respectively, the only loss is that due to the network under matched conditions. This requires $Z_{11}=Zs$ when 22 loaded with Z_L, and $Z_{22}=Z_L$ when Zs faces 11. If a network is designed on this basis it is said to be *image operated.*

$Z_{11}=\sqrt{Zoc_1\,Z_{22}}=\sqrt{Zoc_2.Zsc_2}$
Zoc_1— Z into 11 when 22 open
Zsc_1— Z into 11 when 22 shorted
Zoc_2— Z into 22 when 11 open
Zsc_2— Z into 22 when 11 shorted



Now $V_{11}i_{11}/V_{22}i_{22}=e^{2}\,\theta$ where θ is the image transfer constant. $\theta=tanh^{-1}\sqrt{Zsc_1/Zoc_1}$ and may be complex. The real part constitutes attenuation, the imaginery part phase shift. When $Zs=Z_L$, $e\,\theta=V_{11}/V_{22}$ direecly.

Power supplies

POWER TRANSFORMER DESIGN

The design of small power transformers for electronic use is greatly simplified by the fact that efficiency and power factor are secondary considerations. The transformer is therefore initially designed for the temperature rise permitted by conventional insulating materials. Since manufacturers produce scrapless laminations, which are of uniform proportions, the design procedure has become conventional. If the procedure below is followed, only minor adjustments are necessary to bring the design to an optimum:

1. Calculate the total secondary volt-amperes, $(EI)s = E_1I_1 + E_2I_2 \ldots.$ The r.m.s. value of current and voltage must be used. All the voltages will be sinusoidal, but for the rectifier H.T. winding I_{rms}/I_{dc} is: approximately 2 for halfwave capacitor input, 1 for fullwave capacitor input, and precisely 0·707 for fullwave inductor input.

2. For scrapless *EI* or *TU* laminations with a stack equal to, or up to $1\frac{1}{2}$ times, the width of the centre limb, the area of cross-section of the core *A* (sq. cm.) is given by:

$$A = 1{\cdot}16\sqrt{(EI)s} \text{ sq. cm.}$$

This area can be provided, say, by the next narrowest stock centre limb width below $\sqrt{2A/3}$ stacked to slightly more than $\sqrt{3A/2}$. If an unusual amount of insulation is to be provided, or there are four or more secondaries, more window space will be needed, so that the next *widest* centre limb width above $\sqrt{2A/3}$ should be chosen. The window width will then be greater in the same proportion.

3. Calculate the TPV (turns per volt) from the peak flux density to be used. (This is determined by the permissible heating). $\text{TPV} = 1/(\sqrt{2}\pi B_{max}Akf)$ (*mks* units) where *A* is the area of cross section in square metres (=area in sq. cm. $\times 10^{-4}$), B_{max} is the peak

flux density in weber/sq. metre, k is the core stacking factor (which takes account of insulation and air between laminations, and for the usual thickness of 375μ (0·015″) has a value of about 0·92), and f is the frequency. For a reasonable temperature rise B should not exceed about 1·2 Wb/m^2 for Stalloy, and similar silicon steels. (1 Wb/m^2 = 10,000 gauss).

4. Calculate various secondary turns. The normal ratings for copper wire give a regulation of 10% or so, so all secondaries should be given 10% more turns than that given by the TPV figure, i.e. N_s = TPV $\times E_s \times 1{\cdot}10$. The heater windings should be calculated first, since it is desirable to keep heater voltages as close to nominal as possible, and the windings should therefore come to a whole number of turns; rounding off by half a turn or so may increase the voltage by several per cent. If necessary therefore the TPV figure should be altered slightly to make Ns a whole number for L.V. windings. The other secondaries can then be calculated.

5. Calculate primary turns. On the assumption that primary resistance drop is negligible, N_p = TPV $\times E_p$.

6. Find the approximate primary current, assuming the probable values of 90% efficiency and 90% power factor, i.e.

$$I_p = (EI)_s/(0{\cdot}81E_p)$$

7. Knowing I_p and the secondary ratings, determine the various wire gauges on the basis of about 1·5 A/mm^2 (1000 Ain.^{2}) for conservatively rated or large transformers; up to $2A/mm^2$ for smaller transformers. For the smallest (e.g. heater) transformers, which have a better surface area/volume ratio, up to $3A/mm^2$ is permissible.

8. Plan arrangement of windings. If the primary is to be screened then it should be the inner winding, and an *open circuit* single-layer copper-wire or foil sheet wound above it; this should be earthed. The finer-gauge windings should be on the inside since these wind more smoothly, but the outer heavy-gauge windings should then not be allowed to crush these by over-tight winding.

The tables on page 25 give the necessary information for planning windings. Margins of 3-4 *mm*. should be left to avoid spilling turns or creepage between layers: this is sufficient for peak working voltages up to 1 kV., and should be increased roughly in proportion for higher voltages. The enamel insulation of the wire, taking account of the inevitable bruising during winding, should not be called upon to stand working voltages of more than 30 v. or so. Very desirably, each layer should be interleaved with one layer of paper. Test voltages should be a minimum of 1000 v., and double the working voltage above that. Safe test voltage ratings are, in volts peak per layer: Kraft paper 10v./micron (250v./mil); Empire cloth (yellow) 40v./micron (1000v./mi). The total depth or build of the windings, including bobbin, should not exceed 0·9 of the window depth, while the bobbin should be about 1*mm* narrower than the window. Suitable clearances should also be allowed for the centre limb through the bobbin.

9. The regulation can then be roughly calculated. At 80°C. copper has a resistivity 1·25 times that at 20°C. The figures given on page 18 of resistance per metre should be multiplied by 0·0125, by the length of mean turn, and by the number of turns to give the winding resistance. The length of mean turn for any winding is:

$lmt = 2(a+b) - 8r + \pi(2r+d)$ cm. $or = 2(a+b) - 1{\cdot}7r + \pi d$ cm. where a and b are the internal dimensions, r the internal corner radius, and d the depth, of the winding.

Very roughly, $lmt = 2(a+b)$. Copper losses on normal load can then be calculated. Iron losses for the core material are given by the manufacturer in watts/*kg* for various peak flux densities. Adding these gives the total loss W. The temperature rise is then roughly:

$$\Delta T = 30\ W/A\,°C.$$

where A is the total surface area in sq. cm. of the core *considered as a solid block* (i.e., ignoring windows).

The resultant temperature ought not to be above 80°C. If the environmental temperature is already high, the temperature rise must be restricted.

POWER SUPPLY CIRCUITS

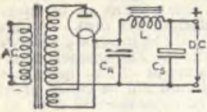


Fig. 1. Half-wave rectifier with smoothing circuit

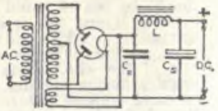


Fig. 2. Full-wave rectifier with smoothing circuit

Half-wave rectifier circuit for ac/dc receiver. (*Fig. 3*). R_2 serves the double purpose of smoothing and preventing heavy surges of current through the valve. It should be between 100 and 500 ohms. For 0·3 amp. heater R_1 is found by subtracting the total voltage of heaters from the supply voltage and dividing by 0·3.

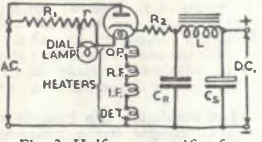


Fig. 3. Half-wave rectifier for a.c., d.c. receiver

Example

Value heaters voltages = 6·3 + 6·3 + 6·3 + 26 + 26 + 4 = 75 volts. Supply voltage = 240, then E_1 = (240 − 75)/·3 = 165/·3 = 550 ohms. *r* should be about 10 ohms for a 4-volt dial lamp.

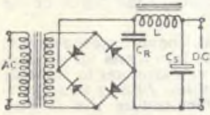


Fig. 4. Full-wave bridge circuit using metal rectifiers

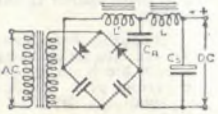


Fig. 5. Voltage doubler circuit using metal rectifiers

In all of the above circuits, the values of *L*, *Cr*, *Cs* depend largely upon the value of H.T. current and the degree of smoothing required. The reservoir capacitor *Cr* is usually 4μF. for currents up to 200 mA. and a voltage of anything up to 400 v. Paper or oil-filled capacitors are preferable to electrolytics in this position. The smoothing capacitor *Cs* is usually either an 8 μF. or 16μF. electrolytic type. The inductor *L* should have a value of not less than 10 Henries at the current it has to carry. A gapped inductor, sometimes called a swinging inductor, is often put in front of the reservoir capacitor, as *L'* in circuit No. 5, to give better regulation (i.e. less variation of output voltage with change of output load).

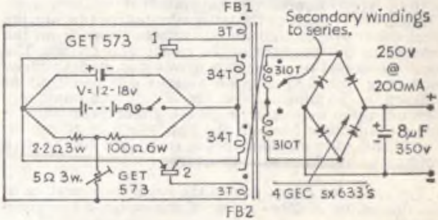


Fig. 6. 50W. invertor by G.E.C. laboratories

When say Tr1 conducts the induced voltage in FB1 causes further heavy conduction, whilst FB2 biases Tr2 into cut-off. The battery voltage is now applied across the half primary winding inductance and the flux increases ($V= -Nd\Phi/dt.10^8$) until $+\Phi s$ saturation when FB1 begins to cut off Tr1, FB2 opening Tr2, which remains open until $-\Phi s$ saturation of core occurs and the cycle recommences. Frequency is $\simeq V.10^8/4N\Phi s$, where N is half primary turns and Φs in gauss. The transistors are subjected to twice battery voltage on the collectors when being switched off. In this circuit mount transformers on 3" × 3" × 16SWG aluminium heat sinks and adjust the 5Ω preset to ensure that $Im < 0{\cdot}8Ipk$. The frequency is about 400H_z.

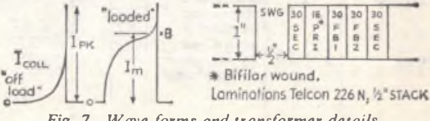


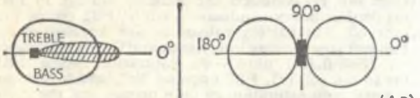
Fig. 7. Wave-forms and transformer details

Audio frequency

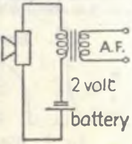
MICROPHONES

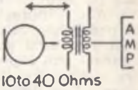
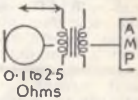
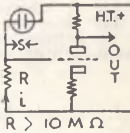
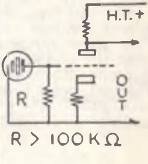
A microphone is a transducer which transforms acoustic pressure fluctuations into electrical output. The pressure wave drives either the front only (pressure operated) or both sides (pressure difference operated) of the diaphragm. The force developed on the latter vibrates the transducing mechanism proper, and output voltage depends on the velocity or displacement of the mechanical system. The force developed on a pressure driven mic. varies widely, increasing with frequency on average due to effects of diffraction. To minimise this, size is kept to a minimum. The net force developed in pressure difference mics. is due to difference between the opposite face pressures. The sound wave at the rear face is delayed by the path difference round the mic. structure.

Typical polar diagrams and the basic facts relating to the operation of most types of microphone are given below:



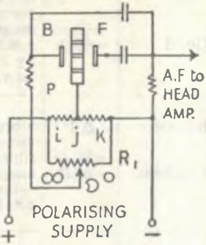
PRESSURE (P) — PRESSURE DIFFERENCE (ΔP)

Type & quality	*Operation & output*	*Transducing system*	*Circuitry*
Carbon Poor	P Approx. 1 v.	Variation of resistance of of carbon granules by by vibration of diaphragm	

Type & quality	*Operation & output*	*Transducing system*	*Circuitry*
Moving coil **Good**	**P** **−70 *to* −40*db***	**vibration of coil in field of permanent magnet—e.m.f. induced by Faraday's Law.**	
Ribbon **Excellent**	**△ P** **−80 *to* −60*db*.**	**vibration of metal-foil ribbon in magnetic field —induced e.m.f.**	
Condenser **Old types, fair.** **new types, excellent**	**P (old)** **△ P(new)** **—30 *to* —50*db* (pre-amp output)**	**diaphragm is one plate of condenser. Variation of Salters charge Hence change in i × R gives output.**	
Crystal **Fair to good**	**P** **1 v. to 0·2 v. high quality lower approx. −50 *db*. −80 *db*.**	**Piezo-electric effect. Plates of quartz or Rochelle salts stressed by diaphragm.**	

Variable polar diagram condensor microphones

These comprise two diaphragms mounted at back (B) and front (F) of a common central electrode plate (P). A number of holes are drilled in this plate, sound waves passing through the holes undergoing a delay. As far as any one diaphragm is concerned it is pressure difference operated since both internal and exposed faces are driven by the sound wave. The delay via the internal hole route is designed equal to the external path difference delay, the resulting polar diagram being a cardioid. If one diaphragm only is polarised a cardioid results; both excited with same polarity and audio outputs added—an omni-directional pattern; both oppositely polarised—a figure of eight pattern. When R_1 slider is at j, B diaphragm not excited giving cardioid; at k, B and F are polarised in same sense, giving omni pattern; at i, B and F oppositely polarised, giving figure of eight.



A.f. hiss noise level

For a hi-fi system the noise should be $>55db$ down on output signal. Hiss is mainly contributed by the thermal noise of the source resistance. $e_n = 1{\cdot}55.10^{-8}\ \sqrt{R\mu V} = -154 + 10\ \log_{10}\ R\ db$. For $R = 300\Omega$ and mean microphone level of $-70db$ initial separation is $58db$ The first stage anode current contributes shot noise hiss, with the addition of partition noise (I_{G2} leaving I_A) in pentodes. These effects assigned a hypothetical equivalent thermal noise resistance; triodes $R = 2{\cdot}5/G_m$, pentodes $R = I_A[(2{\cdot}5/G_m) + (20I_{G2}/G^2{}_m)]/(I_A + I_{G2})$. At a.f., shot noise may practically be neglected.

AF AMPLIFIERS COUPLING AND DECOUPLING

Resistance capacitance coupling

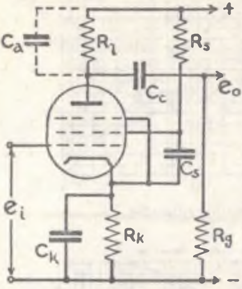


Fig. 1.

Ra = valve ac resistance, $R = R_l\ R_g/(R_l + R_g)$ $r = R\ R_a/(R + R_a)$, $\omega = 2\pi f$, other symbols as diagram.

Stage gain at medium frequencies
$Am = \mu R(R + Rc)$; phase shift $= \pi$.

Gain at low frequencies
This falls (due to C_c), and is: $\dfrac{A_l = \mu R/(R + R_a)}{\sqrt{1 + 1/\ \omega^2 C_c^2 R_g^2}}$.
The attenuation is shown in Fig. 2, in which the frequency scale is in terms of f_1, given by $1/2\pi R_c C_g$, at which $Xc = Rg$ and the response is down $3db$. (It is useful to regard the $3db$ points as the limits of the pass band.) Phase shift is $\theta_l = \tan^{-1} 1/\omega C_c R_g$ (leading) $+ \pi$.

Gain at high frequencies

The gain falls due to shunting of R_l by C_a and is:

$$A_h = \mu R/(R + R_a)\sqrt{1 + \omega^2 C_a^2 r^2}.$$

Fig. 2 also includes this curve, but the *upper* frequency scale must be used. Here the $3db$ point corresponds to $2\pi f_2 C_a = r$. Phase shift is $\theta_h = \tan^{-1} \omega C_a r$ (lagging) $+ \pi$.

Screen decoupling

At low frequencies C_s is not negligible and the gain drops (Fig. 3) according to the same law as for coupling, but the same curve cannot be used as the presence of the screen ac resistance R_{sg} across C_s causes the loss to become constant, below a frequency determined by C_s and R_{sg}, and at a level determined by R_s and R_{sg}. The parameter R_{sg} is rarely available, but is typically $5R_{at}$ where R_{at} is the anode ac resistance as a *triode* at the screen voltage. Making this assumption, Fig. 3 gives curves for various values of R in terms of R_{at}. At frequency f_3, $1/\ \omega C_s$ = parallel resistance of R_s and $5R_{at}$ ($= R_{sg}$).

R.C. Coupling Response Curves

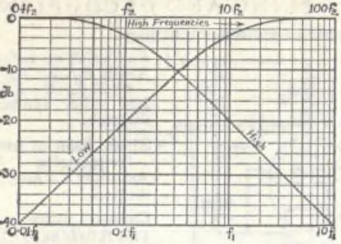


Fig. 2. Coupling response

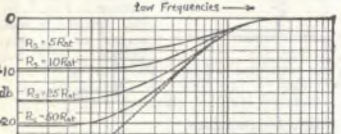


Fig. 3. Loss due to Screen decoupling

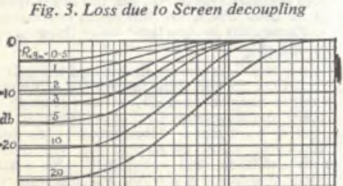


Fig. 4. Loss due to Cathode decoupling

Cathode decoupling

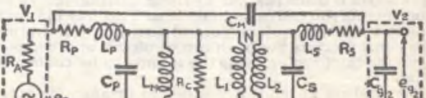
When C_k is not negligibly small, gain is reduced considerably by current feedback, which depends upon the product $R_k g_m$. Fig. 4 gives curves for this loss for various values of $R_k g_m$, in terms of the frequency f at which $1/\omega C_k = R_k$. This curve reveals that the droop begins at frequencies well above this, so that for consistent decoupling the product CR should be at least ten times that for the screen. The loss will be reduced by any falling-off in screen decoupling but this interaction can be ignored since the error will be a pessimistic one and usually not large. The overall response can therefore be obtained by direct addition of the four separate ones.

Transformer coupling

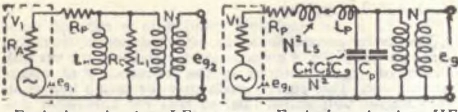
The information given here applies also to input and output transformers. The main feature of the use of a.f. transformers is the production of a band-pass response whose *LF* cut-off is due to the decreasing primary impedance with frequency and *HF* cut-off is due to both leakage inductance and shunt capacitance. Very often the *HF* falling-off is accelerated by a resonant peak which precedes it, due to the leakage inductance resonating with the shunt capacitance.

Equivalent circuit diagram of transformer

The stage gain at any frequency is best studied by reference to an equivalent diagram embodying all the parameters of both the transformer and its accompanying circuits. The equivalent circuit may be broken down further into simpler diagrams which include only those elements which have appreciable effects at the frequencies involved.



Equivalent circuit of transformer at any frequency



Equivalent circuit at LF *Equivalent circuit at HF*

Fig. 5. Equivalent diagrams

μ = amplification fac'r of V_1
e_{g_1} = input voltage of V_1
R_a = ac resistance of V_1
R_p = res. of transf'r pr'y.
R_s = res. of transf'r sec'y.
L_p = pr'y leakage inductance
L_s = sec'y leakage inductance.
C_p = pr'y self capacitance.
C_s = sec'y self capacitance.
C_m = mutual capacitance between P and S.
L_m = magnetising induct.
R_c = iron losses of tr'f'r.
L_1L_2 = ideal transf'r with no losses.
N = turns ratio of transf'r.
C_{g_2} = total input shunt capacitance of V_2
e_{g_2} = voltage applied to V_2

Stage gain at low frequencies

$$A_l = A_m\sqrt{1+(R_c(R_a+R_p)/\ \omega L_m(R_c+R_a+R_p))^2}.$$

Stage gain at high frequencies

$$A_h = A_m\sqrt{(f/f_0Q)^2+(1-f^2/f_0^2)^2}.$$

where f = frequency at which H.F. gain is measured.
f_0 = frequency at which total leakage inductance resonates with shunt capacitance.
$Q = 2\pi f_0(L_p+N^2L_s)/(R_a+R_p+N^2R_s)$.

Gain variation by varying mu

Volume control by varying the amplification of the valves themselves is made possible by using variable-mu valves and varying the grid or cathode bias. This has the great advantage that the volume control carries only dc and not the signal voltages, so that it can be placed at a distance from the amplifier, enabling the volume to be controlled remotely.

Using these valves, it is possible to arrange for automatic control of gain by the signal itself, so that a steady level of volume is maintained, just as avc maintains a steady r f signal.

TONE CONTROL CIRCUITS

The chief types of tone control circuits required in A.F. amplifiers are treble cut or lift, and bass cut or lift. These four types are illustrated by the frequency response curves shown in Fig. 6; all may be obtained by simple circuits of the type given below.

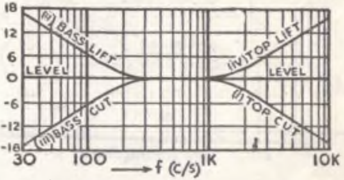
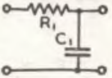
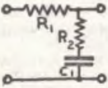


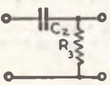
Fig. 6. Frequency response curves



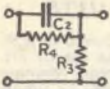
***Treble cut*. If $R_1 = 100\ k\Omega$; $C_1 = \cdot 001\ \mu F$, degree of treble cut is approximately 14 *db* at 10 KHz relative to response at 1 KHz.**



***Bass lift*. If $R_1 = 100\ k\Omega$; $R_2 = 22\ k\Omega$; $C_1 =$ 0·05 μF, degree of bass lift is approximately 15*db* at 50 Hz relative to response at 1 KHz.**



***Bass cut*. If $C_2 = 0{\cdot}001\ \mu F$; $R_3 = 1\ M\Omega$, degree of bass is cut approximately 12 *db* at 50 Hz relative to response at 1 KHz. This circuit has the same form as the R.C. coupling circuit of an amplifier and shows the necessity of making C_3 large if *no* bass cut is required.**



Treble lift. If $C_2 = 0{\cdot}0001\ \mu F$; $R_3 = 100\ k\Omega$ $R_8 = 500 k\Omega$, degree of treble lift is approximately 12 *db* at 10 KHz relative to response at 1 KHz.

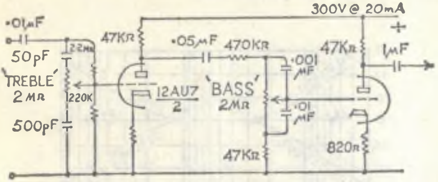


Fig. 7. Tone control circuit

In the above circuit the flat gain is about unity. It should be placed between pre-amp and power output stages. To avoid noise, the input should be $>{\cdot}01\,V$. r.m.s., but $<2{\cdot}0\,V$. r.m.s. to avoid distortion. Maximum boost or cut is ± 16 *db* at 30 Hz and 10 KHz rolling away from 750 Hz cross over.

RESONANT R.C. CIRCUITS

The Twin T amplifier shown below is very useful for A.F. equalisation where the inductors of the equivalent L.C. circuit would be impracticable or undesirable.

If $A =$ valve amplification without feedback, p is the fraction of output voltage fed back (as determined by the tapped anode load), and n is the parameter indicated in the circuit diagram, then:

Resonant frequency $f_0 = 1/2\pi CR$.

Gain at resonance $A_0 = A/(n+1) = A/6$ for $n=5$.

Gain remote from resonance (at frequency limits) may be less than 1.

$A_l = 2A/n(1+pA) + 2 = 2A/(5pA+7)$ for $n=5$.

Total lift of peak relative to frequency limits:

$A_0/A_l = [n(1+pA)+2]/2(n+1) = (5pA+7)/12$ for $n=5$.

1. the valve should be well decoupled to avoid phase shift within the band of interest;
2. the driving impedance (output of previous stage) should not exceed R;
3. pR_l should not exceed $R/5$;
4. C_f should be negligible compared with R.

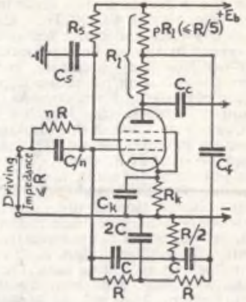


Fig. 1. Twin T amplifier

Design procedure should commence by referring to the curve to be equalised. The frequency up to which it is to be equalised becomes f_o and the amount it is to be lifted relative to distant frequencies becomes A_o/A_l. This gives pA. The actual gain desired from the stage gives A, and thence p. nR is then made as high as the grid circuit permits; this determines the bridge elements R and C, and R_l is calculated as $< Rp/5$. A valve stage is then designed with this value of load and amplification A. Component tolerances in the bridge elements should be $\pm 5\%$, and nR and C/n should be $\pm 10\%$.

As an example, component values are given for equalisation of the typical magnetic recording curve on page 72.

Valve EF86: $E_b = 250$, $I_k = 2{\cdot}0mA$, $R_l = 82k + 15k$ $(pR_l = 15K)$, following grid leak $=$ $330k$, $Rs = 390k$, $C_s = 10\mu F$, $Rk = 1k$, $C_k = 100\mu F$; $f_o = 10{\cdot}6$ KHz, $n = 5$, $nR \simeq 750K$, $c/n = 20pF$ (10%); $R = 150k$, $C = 100pF$ (5%); $C_f \simeq 5\mu F$; $A = 41db$, $A_o = 25db$, $A_l = +7{\cdot}5db$, $A_o/A_l = 17{\cdot}5db$; $3db$ frequencies at 4·6 KHz and 24·4 KHz.

LOUDSPEAKER OPERATION

Matching to the speech coil by a transformer requires a turns ratio $tp/ts = \sqrt{zp/zs}$ where *zp* and *zs* are valve output and speech coil impedances respectively. The conversion efficiency of electrical input to acoustic output varies over the A.F. band. The cone mass and spider compliance resonates at 40-100 Hz. Above the resonance cone velocity (*Uc*) and hence acoustic power output (*Wa*) is governed by the cone mass (inertia). Thus *Uc* is choked off as frequency increases, but $Wa = Uc^2Rr$, *Rr* the acoustic radiation resistance depends on shape and dimension of radiator. For a cone *Rr* increases with frequency and off-sets the choking, but between 300-600 Hz *Rr* ceases to rise. Beyond this choking would set in, but corrugations in cone become effective, one or more of which radiate over H.F. range. This variable H.F. behaviour is overcome by using a small separate L.S. or tweeter feeding both loudspeakers via cross-over networks diverting treble to tweeter only. To maintain bass, a baffle is necessary otherwise back radiation will cancel frontal output. The worst baffle circular, the best rectangular 3:2 ratio. By cabinet design, back radiation can reinforce frontal output extending bass. The cabinet walls must be lined with H.F. absorbent quilting and, apart from a vent at bottom, front should be sealed. The enclosed volume should resonate below speaker resonance (at least 12 cu. ft.)

STEREOPHONY

A listener decides the position of a sound source partly by the difference in loudness produced at the two ears, and partly by the difference in time at which the sound arrives at each ear, plus certain apparent differences in sound quality and reverberation. The various systems of stereo microphone arrangements make use of either time or intensity difference or both. The loudspeaker arrangement shown in Fig. 1 is satisfactory for most systems; note that each ear in fact hears sounds from both L.S., though at different times and intensities. Most systems are not suitable for headphone listening.

Fig. 2 shows the intensity difference system using coincident microphones (within a few inches at most so that no phase differences occur) having a figure 8 characteristic *i.e.* ribbon or condenser types. A sound source at A is on the dead axis of the R.H. mic. and the live axis of the L.H. mic., and so appears to come from the L.H. loudspeaker. A source at C produces equal outputs on L. and R. channels and so appears to come from midway between the speakers (which must be connected in phase), and so on for any point between A and B.

Fig. 1

Fig. 2

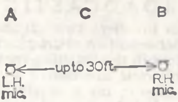


Fig. 3

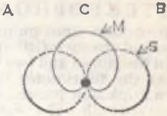


Fig. 4

Fig. 3 shows one of the layouts using spaced mics., usually cardioid or omni-directional, giving mostly time difference information. With this system the sound image tends to be concentrated near the L.S., giving the effect of a hole in the middle, which is often partly filled up by placing a third mic. in the centre, fed equally into both channels.

Most systems of stereo transmission (*see below*) require the signals (left + right) (M signal) and (left − right) (S signal), which are obtained by using a sum and difference network. See Fig. 5. Alternatively the M and S signals may be obtained directly from the mics. using an arrangement like Fig. 4.

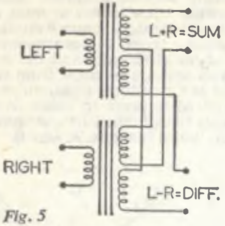


Fig. 5

Decreasing the S signal relative to the M signal decreases the apparent width of the stereo image at the loudspeakers, while increasing the S signal increases the width up to a certain limit where out-of-phase effects arise. The width can also be reduced simply by cross-mixing the L. and R. signals with a variable attenuator connected between the two channels; 8-12 db gives about half width and zero attenuation gives a central (monophonic) image; unwanted cross-talk has the same effect. The image can be offset in either direction by a balance control which raises the gain of one channel relative to the other.

STEREOPHONIC BROADCASTING

Broadcast stereophony provides in effect two separate sound channels (left and right) between two microphones in the studio and two loudspeakers in the listening room. At the transmitter the two channels are coded by a multiplexing process for transmission on a single vhf wavelength. The signal is decoded by a special vhf receiver in order to extract the two signals required by the loudspeakers.

The BBC, in common with other European broadcasting authorities, uses the pilot-tone (Zenith-G.E.) system of stereophonic broadcasting. The system may be envisaged as a process of alternately switching the left and right channels to a single vhf transmission. A similar switching process at the receiver performs the reverse operation and extracts the left and right signals for feeding to the respective loudspeakers. The switches at the transmitter and in the receiver operate at the rate of 38,000 times per second and are kept in synchronism by the pilot-tone transmission. The system is fully compatible in that it allows ordinary vhf receivers to obtain satisfactory monophonic reception of stereophonic broadcasts. Moreover, it does not require separate transmitter wavelengths for the two channels.

To hear these programmes stereophonically special receivers are required. Ideally, the two loudspeakers should be spaced 6 to 12 feet apart and the listener should sit at a point which is equidistant from them and preferably not closer to them than their distance apart.

The stereophonic programmes on the vhf Third Network are broadcast in the Music Programme and in the Third Programme at times marked by a special sign Ⓢ in the appropriate regional edition of the *Radio Times*, which also gives the correct vhf frequencies. The emphasis is mainly on music, because this type of programme derives the greatest benefit from stereophonic reproduction and also provides the necessary degree of compatibility for monophonic reception.

Initially, the stereophonic transmissions are available to listeners within the service area of the Third Network vhf transmitters in South-East England at Wrotham and Swingate (Dover) and in the Midlands from the Sutton Coldfield vhf transmitter, and in the North of England from the transmitter at Holm Moss.

REPRODUCTION FROM GRAMOPHONE RECORDS

78 rpm, or coarse-grove discs have approximately 90-120 grooves per inch; long-playing or fine-groove discs (33⅓ or 45 rpm) have 200-300 grooves per inch. Fine-groove discs frequently use variable groove pitch, the spacing being increased to allow high modulation on loud passages, and decreased on quiet passages giving wide dynamic range with a long playing time.

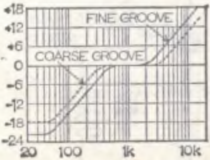


Fig. 1

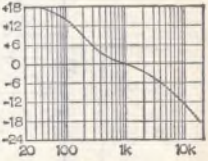


Fig. 2

If all frequencies were recorded with a constant rms velocity, the bass would have excessive amplitude requiring wide groove spacing, while the treble would have very small amplitude and so a poor signal-noise ratio. Frequencies below about 250 H_z are attenuated at approximately 6 db per octave and, except on earlier 78 rpm discs, the high frequencies are pre-emphasised. Fig. 1 shows the BSI 1928: 1955 specification for recording equalisation; early 78 rpm discs had approximately the same L.F. roll-off but no top lift.

The ideal reproducing equaliser would produce the inverse of this curve, assuming a perfect velocity-sensitive pick-up. Most record companies use recording characteristics which differ from the B.S.I. curve, no single reproducing equaliser being correct for all discs, but the R.I.A.A. standard reproducing characteristic shown in Fig. 2 gives results within a few db for most discs. Fig 3 shows a circuit for a velocity sensitive pick-up (moving coil, moving iron or ribbon). Crystal pick-ups produce voltage α recorded amplitude, and can be used unequalised for many purposes but for high quality they can be shunted to produce an approximate velocity characteristic and equalised as shown, or used with a specially designed equaliser.

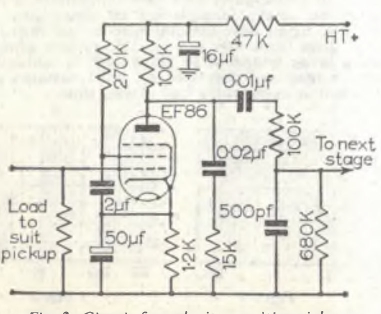


Fig. 3. Circuit for velocity sensitive pick-up

Crystal pick-ups give an output of about 0·1-2·5 volts; magnetic types only 10-200 mV. Magnetic pick-ups, their associated wiring and transformers are liable to hum induction; care is necessary with the low output to avoid trouble from turntable motor and mains wiring. The standard 78 r.p.m. reproducing stylus has tip radius 0·0025 in. and for long-playing discs a radius of 0·001 in. For stereo a radius of 0·0005 in. is recommended, though 0·007 in. is often used. The stylus pressure on the disc should be less than 10 grams for long-playing discs and preferably less than 5 grams for stereo.

The groove on a monophonic disc is of constant depth but has lateral variations which cause the needle to vibrate horizontally. In an early system, the variations were *hill-and-dale* and produced vertical vibrations. In stereophonic recording, a right-angled cutter is driven by two drive rods (Fig. 4) extended from separate moving coil units, in such a way that two distinct hill-and-dale recordings are made on the 90° displaced opposite faces of the v-shaped groove.

The stereo information can then be recovered by scanning the groove with a single needle which drives two separate voltage generators in accordance with the information recorded on the two groove walls.

Fig. 4. How a stereo disc is cut

Fig. 5. Crystal stereo pick-up

A typical stereo pick-up consists of two crystal bimorphs, clamped at their rear ends and embedded at their front ends in a diamond-shaped sheet of plastic (the resolver) as shown in Fig. 5. The top of the resolver is held in a rubber clamp and the needle tip is fixed to the end of a rod which fits into a notch at the lower corner of the resolver.

If only the right-hand groove wall is hill-and-dale modulated, the arm of the resolver BC is vibrated longitudinally and turns arm AB about A, stressing the left-hand (R) bimorph, whilst arm CD pivots about D and produces relatively little stress in the right-hand (L) crystal.

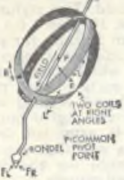


Fig. 6. Moving coil pick-up

If, on the other hand, only the L.H. groove wall is modulated, it is the R.H. crystal (L) which is most stressed. A moving-coil resolver, which does not make use of pivoted levers is shown in Fig. 6. Two coils fixed at right angles can turn about their common centre, and are situated in a magnetic field parallel to their common diameter. L modulation alone turns the coils as shown; coil (r) remains in its own plane and so produces no output and coil (1) cuts lines of force and so has a signal induced.

MAGNETIC RECORDING

The storing of information in the form of a magnetic pattern on a wire or tape dates back to 1900, when Poulsen exhibited the first magnetic recorder at the Paris Exhibition of that year. Nowadays the medium is generally a ¼″ wide plastic or paper tape coated with a 0·5 mil layer of very finely divided ferric oxide ($Fe_2O_3\gamma$).

Wiping process

Any previous recording is erased by passing the tape over the twenty mil long gap in a ring type electromagnet, energised by high frequency alternating current. The magnetic flux, due to the current, is concentrated in the tape coating over the gap. Because the density of the flux in the tape is not uniform, but varies from zero value on the left-hand side of the gap to maximum in the middle and then back to zero at the right-hand side, the segments of coating passing over the gap are subjected to alternating magnetising forces which gradually increase in strength to saturation value and then diminish to zero. Thus any previously recorded programme is erased and the tape coating leaving the gap is in a completely demagnetized condition, ready for re-recording.

Recording process

The recording head is a similar ring-type electromagnet, though with a much narrower gap. It is energised by the amplified output of the microphone, to which a supersonic bias (obtained from the oscillator, which also provides the erase current) is added. The function of the bias, which is of critical amplitude, is to inhibit the distortion of low audio frequencies, which would otherwise arise because of the s-shape of the initial magnetisation curve of ferric oxide. Another important function of the bias is to increase the output level, thus ensuring a good programme-to-noise ratio.

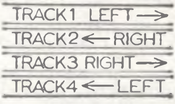


Fig. **1**

Domestic recorders have recording heads of half or quarter tape width and therefore record two or four tracks. On stereo, four-track, recorders the disposition of the tracks is as shown in Fig. 1.

Reproducing process

Reproduction is effected by passing the tape (after rewinding) over the same ring-type electromagnet as used for recording. On most professional machines there is a separate reproducing head; with such an arrangement the recording can be monitored as it is being made. A change-over switch enables the operator to listen, at will, to the incoming sounds or their reproduction from the tape. During reproduction (other than whilst recording as described above) the erase head is not energised. As the recorded tape moves over the gap in the replay head the magnetic pattern on the tape causes changing lines of magnetic force to thread the core of the replay head and this produces at the terminals of the coil surrounding the core, the electrical replica of the recorded signal.

After suitable equalisation and amplification the signal is fed to the loudspeaker and becomes audible. Over the range 100-3,000 H_z the output voltage is proportional to the rate of change of flux in the core and therefore, if unequalised, rises 6dB per octave as shown in the graph.

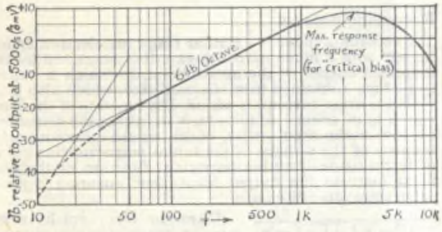


Fig. 2. Magnetic tape recording: unequalized response for typical good commercial recorder (tape speed 19cm/sec)

Below 100 H_z the fall off with decrease in frequency is at a sharper rate, whilst at frequencies above 3 KH_z the output begins to fall off, until it reaches a minimum at what is called *the extinction frequency*. This is the frequency at which the recorded wavelength is equal to the effective gap length in the replay head.

Equalisation

With suitable equalisation the output signal can be made linear up to a frequency which is 2/3 of the extinction frequency.

It is normal for the replay pre-amplifier to include bass-lift equalisation to the C.C.I.R. characteristic for the tape speeds required (response equivalent to a series R-C circuit of time-constant 100 μ sec for 7·5″/sec and 50 μ sec for 15″/sec). This is conveniently obtained by means of a simple anode-grid feedback circuit of R and C in series, e.g., IM and 100 p.f. for 7·5″/sec. Additional very low frequency losses occur when the wavelength on the tape is comparable with the length of pole-piece in contact with the tape and high frequency losses occur when the wavelength is comparable with the reproducing gap width,

$$\text{loss} = 20 \log_{10}[(\sin(\pi\delta/\lambda)/(\pi\delta/\lambda)]\text{db}$$

where δ = the effective gap length in the replay head.

The recording equalisers and bias are adjusted to produce a recording having a C.C.I.R. characteristic with a given head and type of tape. Nearly all losses occurring during recording are at high frequencies: partial demagnetisation due to high bias level; self-demagnetisation of the tape as each half-cycle is short relative to its thickness (both these effects are less with high coercivity tapes) and losses in the record head itself. The top-lift necessary to correct for these losses is obtained from a tuned circuit either L.C. or twin-T negative feedback in the recording amplifier. The recording and reproducing amplifiers are frequently the same amplifier with only the equalisers switched for the different purposes.

A mis-alignment angle $\propto$ between record and replay head gaps with a track of width W gives an HF loss of: $20 \log_{10}[\sin(\pi W \tan\propto/\lambda)/(\pi W \tan\propto/\lambda)]$db

Standard speeds

For the standard speeds (in inches per second), the frequency responses obtained with a 0·0003″ replay head gap are as follows: 30″, originally used for broadcasting, now almost obsolete—15-20 KH_z; 15″, broadcasting and professional recording—15 KH_z; $7\frac{1}{2}$″, high quality domestic music—10 KH_z; $3\frac{3}{4}$″, normal domestic use—6 KH_z (but much greater frequency response is often obtained by reducing bias level and tolerating a greater level of distortion). Lower speeds than these, $1\frac{7}{8}$″ and $\frac{15}{16}$″ sec, are in use for dictating machines.

Electro-magnetic waves

Comprising electric (E) and magnetic (H) fields crossed 90° and mutually perpendicular to line of travel. In free space $E/H = Zo = \sqrt{\mu o \varepsilon o} = 377$ ohms/M^2 and velocity $C = 1/\sqrt{\mu o \varepsilon o} = 299{,}793$ km./sec. $\simeq 3.10^8$ M/sec. where permeability $\mu o = 1{\cdot}257\ 10^{-6}$; henry/$M$, and permittivity $\varepsilon o = 8{\cdot}85.10^{-1}$ farad/M. Wavelength in space is $\lambda =$ (3.10^8/frequency) metres/sec. Wave speed 5·376 μS per mile, 3·335 μS per Km.

Propagation

The wave may proceed to the receiver along one or more of three possible routes:

1. Direct (line of sight)
2. Reflected (bounce off conducting object)
3. Surface wave (groundwave).

The ionosphere (ionised gas layers above earth) and earth surface constitute reflecting surfaces. Earth surface presents a parallel resistive/capacative impedance to a wave and attenuates surface waves as frequency increases. Above 2 H_z attenuation is severe. Skywaves are reflected from earth with a reflection coefficient r=(reflected/ incident) wave amplitude.

Ionosphere

There are four reflecting layers: D, 50—90 Km; E, 100 Km; F_1 200 Km; F_2, 250—350 Km up by day. By night D and E partially de-ionise, leaving a combined F_1+F_2 layer at 250 Km. These layers have increasing free electron density maxima (N), from D to F_2. The electron density determines the critical frequency, f_c, the highest frequency reflected from the particular layer at vertical incidence. $f_c = 9\sqrt{N}$

FREQUENCY BANDS

Allocation to communications services

In order to transmit intelligence, whether in the form of a simple code (morse), telephonic communication, or video (television) signals, a radio frequency has to be modulated, and the process of modulation causes the rf signal to occupy a channel of so many Hertz bandwidth. The width of this channel depends upon the nature of the intelligence to be transmitted, being fairly low for telegraphic messages in morse (below 1,000 Hz wide), about 10,000 Hz for am sound broadcasting, and over 2·5 megacycles/sec. wide for the video signals used in television. If interference between two separate channels is to be avoided, it is obvious that some form of international agreement must be made on the allocation of frequency bands to each service. Such an agreement has been made on a world basis for general allocation of radio communication services. The world is divided into three main areas, viz., American, European, Australian, for this purpose, the allocations for each service being fairly similar at the hf end of the spectrum, due to the ability of such waves to travel long distances and thus cause interference at the other side of the world.

SHORT-WAVE BROADCASTING

Broadcasting on short waves is confined to a number of relatively narrow wavebands. Although quite a number of high-quality (10 KHz) channels can be accommodated in each band, the number of s.w. stations is rapidly increasing and satisfactory reception is often limited to the advertised service area for the particular transmission. The following wavebands have been allocated to international broadcasting (excluding tropical bands):

3·95– 4·00 MHz	75 metre band
5·95– 6·20 MHz	49 metre band
7·10– 7·30 MHz	41 metre band
9·50– 9·775 MHz	31 metre band
11·80–11·975 MHz	25 metre band
15·10–15·45 MHz	19 metre band
17·70–17·90 MHz	16 metre band
21·45–21·75 MHz	13 metre band
25·60–26·10 MHz	11 metre band

The use of particular channels for long-distance transmission is determined largely by ionospheric conditions. Schedules of transmission times and frequencies are worked out with reference to the ionospheric conditions at particular times of the day and year. Most countries now co-operate with one another to minimise interference, but many transmissions are still deliberately jammed.

From the U.K. a regular world service on s.w. is maintained by transmitters working on the following frequencies:

Call sign	*Freq. MHz*	*Wave-length*
75 metre band		
MCM	3952·5	75·90
GRC	3975	75·47
49 metre band		
MCP	5·975	50·21
MCQ	5·99	50·08
GRB	6·01	49·92
GSY	6·04	49·67
GSA	6·05	49·59
GSX	6·06	49·50
GRR	6·07	49·42
GWM	6·09	49·26
GSL	6·11	49·10
GWA	6·125	48·98
GRW	6·15	48·78
GWK	6·165	48·66
GSZ	6·17	48·62
GRO	6·18	48·54
GRN	6·195	48·43
41 metre band		
MCS	7·11	42·19
GRM	7·12	42·13
GRS	7·135	42·05
GRT	7·15	41·96
GRK	7·185	41·75
GWZ	7·20	41·67
GWL	7·21	41·61
GSW	7·23	41·49
GWI	7·25	41·38
GSU	7·26	41·32
GWN	7·28	41·21
GRJ	7·325	40·96
31 metre band		
GRI	9·41	31·88
GSB	9·51	31·55
GWJ	9·525	31·50
GWB	9·55	31·41
GWX	9·57	31·35
GSC	9·58	31·32
GRY	9·60	31·25
GWO	9·625	31·17
GVZ	9·64	31·12
GWP	9·66	31·06
GWT	9·675	31·01
GRX	9·69	30·96
GWY	9·70	30·93
GWF	9·735	30·82
—	9·75	30·77
MCR	9·760	30·74
MCN	9·770	30·71
GRH	9·825	30·53
GRU	9·915	30·26
25 metre band		
GRG	11·68	25·68
GVV	11·73	25·58
GSD	11·75	25·53
GVU	11·77	25·49
—	11·78	25·47
GWV	11·79	25·45
GWH	11·80	25·42
GSN	11·82	25·38
GWQ	11·84	25·34
GSE	11·86	25·30
GRE	11·88	25·25
GWW	11·89	25·23
MCO	11·91	25·19
GVX	11·93	15·15
MCQ	11·945	25·12
GVY	11·955	25·09
MCT	11·96	25·08
GRV	12·04	24·92
GRF	12·095	24·80
19 metre band		
GWC	15·07	19·91
GWG	15·11	19·85
GSF	15·14	19·82
GSO	15·18	19·76
GWU	15·21	19·72
GWD	15·23	19·70
GSI	15·26	19·66
GWR	15·30	19·61
GSP	15·31	19·60
—	15·36	19·53
—	15·375	19·51
—	15·40	19·48
—	15·42	19·45
GWE	15·435	19·44
GRD	15·447	19·42

Call Sign	*Freq. MH*	*Wave-length*	*Call sign*	*Freq. MHz*	*Wave-length*
16 *metre band*			13 *metre band*		
GVP	17·70	16·95	GSH	21·47	13·97
GRA	17·715	16·93	GSJ	21·53	13·93
GVQ	17·73	16·92	GST	21·55	13·92
GRQ	17·74	16·91	GVT	21·63	13·87
GSG	17·79	16·86	GRZ	21·64	13·86
GSV	17·81	16·84	GVR	21·675	13·84
GRP	17·87	16·79	GVS	21·71	13·82
GVO	17·89	16·77	11 *metre band*		
—	18·08	16·53	GSQ	25·75	11·65
			GSK	26·08	11·50

Optimum wavebands for short-wave reception

Owing to world-time-differences and the reliance of s.w. broadcasting upon ionospheric conditions, s.w. stations in general do not attempt to maintain a continuous service. Instead, they run to a schedule in which wavelengths and directional aerials are changed from hour to hour and season to season. To assist s.w. listeners, *Wireless World* publishes each month tables showing muf's and luf's over different paths. Propagation forecasts, showing wavebands for optimum reception at given periods of the day, also appear in the monthly R S G B publication *Radio Communications* (see page 78).

EXAMPLE TABLE

(*Figures relate to wavebands in metres*)

Season	*U.K. time*	*Europe*	*South Africa*	*North America*	*South America*	*Australia*	*South Asia*	*East Asia*
Spring	Morn.	75	41	31	31	31	19	25
and	Aft.	41	16	16	16	19	25	31
Autumn	Eve.	M/W	31	31	31	31	41	49
	Morn.	49	31	25	31	25	16	19
Summer	Aft.	31	16	19	16	25	19	25
	Eve.	41	25	25	25	31	25	31
	Morn.	75	41	49	49	41	25	31
Winter	Aft.	41	16	19	19	25	31	41
	Eve.	M/W	31	31	31	31	41	49

The SINPO code

This is an international code for reporting details of reception of broadcasting stations and it takes the form of a five-figure code following the word SINPO. The code derives its name from the initial letters of the five qualities observed and each observation is reported in terms of one out of a range of five different degrees.

S = carrier strength	Excellent	5
	Good	4
	Fair	3
	Poor	2
	Barely audible	1
I = Interference	Nil	5
	Slight	4
	Moderate	3
	Severe	2
	Extreme	1
N = noise (QRN)	Nil (–40*db.*)	5
	Slight (–30*db.*)	4
	Moderate (–20*db.*)	3
	Severe (–10*db.*)	2
	Extreme (0*db.*)	1
P = propagation disturbance	Nil	5
	Slight	4
	Moderate	3
	Severe	2
	Extreme	1
O = overall readability	Excellent	5
	Good	4
	Fair	3
	Poor	2
	Unusable	1

EXAMPLE:

A good signal with slight het. but rapid fading would be: SINPO 44333.

Radio Society of Great Britain

The R S G B, a national society for radio amateurs in the United Kingdom, issues a monthly journal *Radio Communication*, free to members, of information on amateur radio activities and technical developments. Subscription: Corporate Members £2 10s. 0d., Associates £1 5s. 0d. Address: 35 Doughty Street, London, W.C.1. Tel: 01-837 8688.

Some international amateur prefixes

Prefix	*Country*
AC4	Tibet
AP	Pakistan
BV	Taiwan
BY	China
CE	Chile
CM, CO	Cuba
CN2,8-9	Morocco
CP	Bolivia
CR4	Cape Verde Is.
CT1	Portugal
CT2	Azores
CT3	Madeira
CX	Uruguay
DJ, DK, DL DM	Germany
DU	Philippine Is.
EA	Spain
EA6	Balearic Is.
EA ϕ	Spanish Guinea
EI	Eire
EL	Liberia
EP	Iran
9F3	Ethiopia
F	France
FC	Corsica
FG7	Guadeloupe
FH8	Comoro Is.
FL8	French Somaliland
FM7	Martinique
FO8	Clipperton Is. and French Oceania
FP8	St. Pierre and Miquelon
FR7	Reunion Is.
FU8	New Hebrides
FY7	French Guiana
G	England
GC	Channel Islands
GD	Isle of Man
GI	N. Ireland
GM	Scotland
GW	Wales
HA	Hungary
HB	Switzerland
HC	Ecuador
HE	Liechtenstein
HH	Haiti
HI	Dominican Republic
HK	Colombia
HM	Korea
HP	Panama
HR	Honduras
HS	Thailand
HV	Vatican City
HZ	Saudi Arabia
I1, IT1	Italy
IS1	Sardinia
JA, KA	Japan
JT	Mongolia
JY	Jordan
K	U.S.A.
KA-KJ	U.S. Stations in Pacific
KL7	Alaska
KN	U.S.A. (Novices licence)
KR-KX	U.S. Pacific Possessions
KZ5	Panama Canal
LA, LB	Norway
LU	Argentina
LX	Luxembourg

Prefix	*Country*
LZ	Bulgaria
M1,9A1	San Marino
MP	British stations in Persian Gulf
OA	Peru
OD5	Lebanon
OE	Austria
OH	Finland
OK	Czechoslovakia
ON4, ON5	Belgium
OX, KG1, XP	Greenland
OY	Faroe Islands
OZ	Denmark
PA, PI	Netherlands
PJ	Dutch West Indies
PX	Andorra
PY	Brazil
PZ1	Netherlands Guiana
SM, SL	Sweden
SP	Poland
ST2	Sudan
SU	Egypt
SV	Greece
TA	Turkey
TF	Iceland
TG	Guatemala
TI	Costa Rica
TN	Congo Republic
UA1-6	European Russia
UA9	Asiatic Russia
UY5,	Ukraine
UC-UO	Russian Republics
UP2	Lithuania
UQ2	Latvia

Prefix	*Country*
UR2	Estonia
VE, VO	Canada, Labrador and Newfoundland
VK	Australia and Dependencies, e.g. VK9–Papua
VP1-7	W. Indies
VP8	Falkland Islands and dependencies
VP9	Bermuda
VQ1	Zanzibar
VQ2	Zambia
VR1-6	British Poss'ns in Pacific
VS6	Hong Kong
VS9	Aden and Maldive Is.
VU	India
W/K	U.S.A.
XE, XF	Mexico
XW8	Laos
XZ	Burma
YA	Afghanistan
YI	Iraq
YK	Syria
YN	Nicaragua
YO	Rumania
YS	Salvador
YU	Yugoslavia
YV	Venezuela
ZA	Albania
ZB1	Malta
ZB2	Gibraltar
ZD3	The Gambia
ZE	Rhodesia
ZK1	Cook Islands and Manihiki Is.
ZK2	Niue

Prefix	*Country*	*Prefix*	*Country*
ZL	New Zealand	5U7	Niger Republic
ZP	Paraguay	5W,	Samoa
ZS1-6	South Africa	7X	Algeria
3A	Monaco	8F	Indonesia
3V8	Tunisia	9G1	Ghana
3W8	S. Vietnam	9K2	Kuwait
4S7	Ceylon	9K3	Neutral Zone
4W1	Yemen	9L1	Sierra Leone
4X4	Israel	9M2	W. Malaysia
5A	Libya	9M6/8	E. Malaysia
5B4	Cyprus	9N1	Nepal
5H3	Tanzania	9Q5	Congo
5N2	Nigeria	9V1	Singapore

The number following the letter K or W (prefix for U.S.A.) or VE (Canada) indicates the State:

K/W1	Connecticut, Maine Mass., New Hampshire, Vermont
W2	New Jersey, New York
W3	Delaware D.C., Maryland, Penn.
W4	Alabama, Florida Georgia, Kentucky, Carolina, Tennessee, Virginia
W5	Arkansas, Louisiana, New Mexico, Oklahoma, Texas
W6	California
W7	Arizona, Idaho, Nevada, Oregon, Utah Washington, Wyoming
W8	Michigan, Ohio, West Virginia
W9	Illinois, Indiana, Wisconsin
W	Colorado, Iowa, Kansas, Minnesota, Missouri, Nebraska, N.
VE1	New Brunswick, Nova Scotia, Prince Edward Islands, Newfoundland
VE2	Quebec, Labrador
VE3	Ontario
VE4	Manitoba
VE5	Saskatchewan
VE6	Alberta
VE7	British Columbia
VE8	Yukon
VE	Radio Clubs (any district)

Transmission

CARRIER WAVE PRODUCTION

Various methods may be used to produce the rf carrier essential for the propagation of em waves, such as rotary alternators, spark generators, tuning-fork drives, and valve oscillators with lc or crystal control of the frequency. Only the two latter find favour in modern transmitters, where precise control of frequency is essential.

All oscillators consist fundamentally of four elements:

1. A frequency-determining network (either electrical or electro-mechanical, eg, crystal, tuned circuit).

2. An amplifying stage to make good the inevitable losses occurring in the fd network.

3. A system of positive feedback from the output of the amplifier to the fd network.

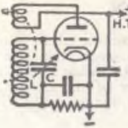
4. A limiting device to prevent the oscillations building-up to a dangerous value. Often this is automatic in the sense that grid current is used to bias back the valve.

Circuits shown on the next page embody inductance, L, and capacitance, C, for frequency-determining elements. Unless otherwise stated, frequency of oscillation is $f = 1/2\pi\sqrt{LC}$. Feedback is obtained in circuits (i) and (iii) by mutual electro-magnetic induction between grid and anode units; in circuit (ii) by grid-to-anode capacitance of valve; in circuits (iv) and (vi) by currents flowing in common grid and anode circuits.

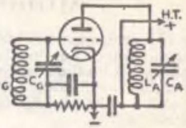
In circuit (v) extremely good frequency stability is obtained as $C_1 \doteq \cdot 003\mu F$ ($C_2 \doteq \cdot 006\mu F$) swamps valve variations—used below 30 MHz; frequency is given by

$$f = f_0\sqrt{1 + \frac{C}{C_1} + \frac{C}{C_2}};\ C \doteq 300pF \text{ max for } f < 7\text{MHz}.$$

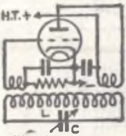
Capacitances (unmarked) other than those in the tuned circuit are for decoupling or self bias.



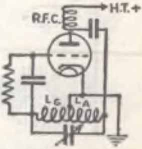
(i) *Tuned grid oscillator.*



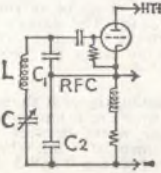
(ii) *Tuned grid/tuned anode oscillator.*
$f = 1/2\pi\sqrt{L_G C_G}$



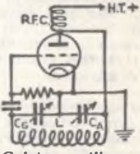
(iii) *Meissner oscillator.*



(iv) *Hartley oscillator.*
$f = 1/2\pi\sqrt{C(L_A + L_G + 2M)}$



(v) *Clapp oscillator.*

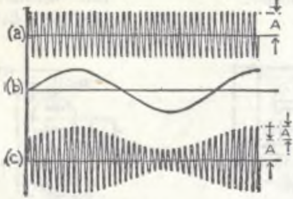


(vi) *Colpitts oscillator.*
$f = \sqrt{C_A + C_G}/2\pi\sqrt{LC_A C_G}$

Basic oscillator circuits

MODULATION

The rf carrier wave radiated from a transmitter does not in itself convey any intelligence to the receivers. The af variations from the microphone must be made to control either the amplitude or the frequency of the carrier, a process known as amplitude, or frequency-modulation respectively.

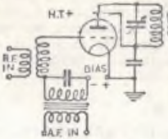


Amplitude modulation

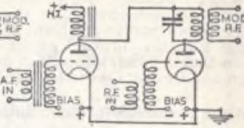
An unmodulated carrier of amplitude A is shown at (*a*); beneath it (*b*) is the af wave with which it is to be modulated. The final wave (*c*) shows the effect of varying the amplitude of (*a*) at the af rate of (*b*). The waves are not drawn to scale, since, even with the longest rf wave (150 Hz), this would represent many cycles for every cycle of af signal.

The amplitude Am of the modulating signal is shown as m times the carrier amplitude, where m is known as the *modulation factor*. The carrier varies between a maximum of $A\ (1+m)$ and a minimum of $A(1-m)$. When $m=1$, the carrier amplitude swings between 0 and 2 A, a condition known as 100% modulation.

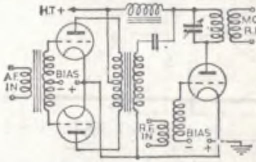
Modulation may either be performed at an early stage in the transmitter (low-power modulation) or the modulating stage may be the final stage (high-power modulation).



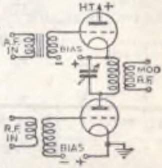
Grid modulation



Anode, or Heising modulation



Push-pull Heising modulation



Series modulation

In these example circuits, triodes (rarely now used) are shown for simplicity. For the same reason, neutralisation circuits are omitted.

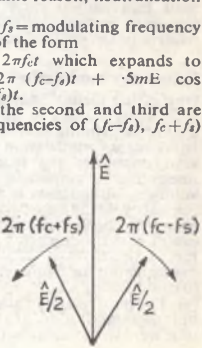
If f_c = carrier frequency and f_s = modulating frequency then the transmitter output is of the form

$E_0 = E(1 + m \sin 2\pi f_s t)$. Sin $2\pi f_c t$ which expands to

$$E_0 = E \sin 2\pi f_c t + \cdot 5mE \cos 2\pi (f_c - f_s)t + \cdot 5mE \cos 2\pi (f_c + f_s)t.$$

The first term is the carrier, the second and third are lower and upper sideband frequencies of $(f_c - f_s)$, $f_c + f_s$) Hertz respectively. The bandwidth requirement is thus ± the highest f_s anticipated centred about f_c.

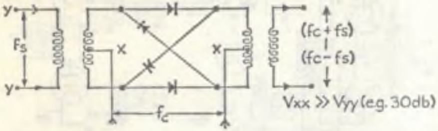
The vector shows the situation for 100% mod., ie, m = 1·0 when the sideband vectors swing the amplitude between zero and twice *B*. The carrier does not itself bear information—providing the receiving system is a given means to artificially recreate an accurate carrier (for demodu-



lation purposes) only the two sidebands need be transmitted or alternatively only the single sideband. As the high power carrier is absent in the transmitter output stage a higher sideband power is possible as also is a reduction in bandwidth with consequent improvement in signal/noise over the carrier plus double sideband system.

Balanced modulator/demodulator

A simple low level device which when fed with frequencies f_s, f_c, produces only (f_s+f_c), (f_s-f_c), ie, no carrier component. If filters are subsequently employed, one of the two sidebands may also be eliminated.

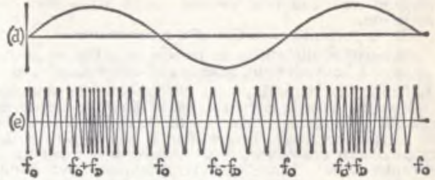


Demodulation may be performed in a similar circuit if reconstituted identical carrier is fed to the points xx. If a frequency ($\triangle f$) or phase ($\triangle\phi$) error exists between mod and demod carriers the demodulated signal is frequency or phase shifted by the same error respectively, relative to the original modulator input signal.

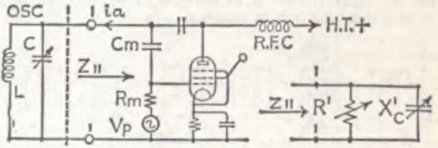
Frequency modulation

One of the main disadvantage of amplitude modulation is its susceptibility to interference, because the receiver cannot discriminate between noise and signal voltages. In frequency modulation, the amplitude of the carrier is kept constant; the receiver can therefore be made insensitive to amplitude variations and hence to noise voltages. Modulation is achieved by making the carrier frequency deviate above and below its normal frequency at a rate corresponding to the af signal impressed, the extent of the swing being proportional to the intensity of the af signal. It is difficult to represent this diagrammatically but the figure on the next page shows an af signal (d) modulating a carrier, f_0, which then assumes the shape (e). The particular deviation frequency, f_D, corresponds to only one amplitude of af and would be greater if the amplitude were to increase. In practice, the maximum deviation

necessary (corresponding to 100% modulation in the am system) is ± 75 KHz, so that it is necessary to use the very high frequency band (above 30 KHz).



Reactance modulator (f.m.)



$$X'c = \frac{Rm^2 + X^2cm}{gmXcm} \qquad R' = \frac{Rm^2 + X^2cm}{gmRm}$$

The voltage across the lc circuit (V_{11}) of the master oscillator is potentially divided, the voltage across *Rm* (*Vg*) being almost 90° phase advanced on V_{11}. This voltage generates a rf anode current $ia = gm.Vg$ which flows back to the tuned circuit ≃ 90° phase leading V_{11} hence Z_{11} predominantly capacitive reactive; magnitude of synthesized capacity depends on *Gm* which is controlled by programme bias voltage $V\rho$. For small *fd* oscillator frequency $\propto V\rho$, A pure tone modulated fm voltage is $e = E \sin(2\pi fot + (fd/fm) \sin 2\pi fmt.)$ This when expanded shows a carrier component and an infinite set of sidebands $\pm fm, \pm 2fm, \ldots \pm fm$.

Reception

The wave form of current induced in the aerial is a replica of the modulated wave transmitted. The function of the receiver is to amplify the aerial voltage and then perform the converse process of demodulation or detection.

The important properties of a receiver are:

1. Selectivity, the ability to receive only the frequency band of interest whilst rejecting all other signals and
2. Sensitivity, the ability to amplify weak aerial voltages.

A simple receiver is the *straight* in which r f amplifiers precede the detector stage and the lc circuits are tuned to carrier frequency. This system is satisfactory if the range of carrier frequencies to be covered is small but if a range of say, 1,500-500 KH_z (medium wave band) is required sensitivity and selectivity vary widely. It is usual to overcome this by converting r f to a fixed *intermediate* frequency (if); as the majority part of receiver gain is effected in the if amplifier the fixed selectivity/sensitivity of the if determines a constant performance for the receiver over a wide tuning range. This super heterodyne system is schematically:

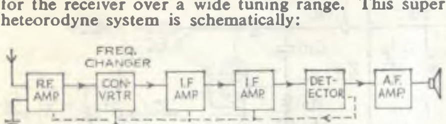


Fig. 1. Superheteorodyne system

Typical ifs are 465 KH_z for am up to hf, 10·7 MH_z for am or fm on vhf and 45, or 60 MH_z for TV or radar receivers. R f amplication is still necessary (i) to reject the *image* rf, 2 × if above the desired rf as otherwise the image would convert into the if (ii) to boost the rf signal amplitude prior to the convertor since the latter introduces a highnoise level. In these circumstances the noise in the rf stage itself sets the overall noise level of the receiver.

A circuit suitable for detection is that of a diode valve rectifier. Fig. 2 shows a typical diode detector circuit, *LC* being the last of the rf tuned circuits. The diode output is developed across the load resistor R_1, but at this stage it has a strong r f component. It is therefore followed

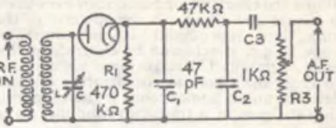


Fig. 2. Diode detector circuit

by a smoothing circuit, consisting of $C_1R_2C_2$, but there is still the dc component, which may be removed by the blocking capacitor C_3. R_3 is the grid leak of the a f amplifier which follows.

Where a sensitive detector is required, a triode or pentode is employed, the grid acting as a diode anode, rectified r f appearing on the grid is amplified as the desired a f signal, filtering is effected in the anode. This cumulative or leaky grid detector is illustrated in Fig. 3. An ordinary a f amplifier biased practically to cut-off point will act as a rectifier and so detect. It is, in fact, an anode-bend detector and gives good amplification but poor quality. If the bias is obtained by making the cathode resistor a high value, and the latter also acts as the load from which the output is taken, the quality is greatly improved but the gain is reduced to less than unity. Its input impedance is very high, which gives it its name, and its performance is comparable with that of the diode detector. One of the main advantages is that it presents very little damping to the r f circuit from which it is fed and so gives better selectivity than the diode circuit. Fig. 4 gives the fundamental circuit.

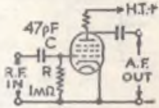


Fig. 3. Leaky grid detector

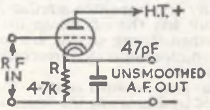


Fig. 4. Infinite impedance detector

Automatic gain control

Most receivers use some form of automatic gain control to minimise the effects of ionospheric or man-made fading, and to ensure that all signals, irrespective of their input amplitude, are reproduced at substantially-constant amplitude. A.g.c. is achieved by controlling the gain of pre-detector, usually i.f., stages by a voltage (or current) derived from the signal at the detector output or at a post-detector point. Some means is therefore required of adjusting the gain of a transistor amplifier by a control voltage.

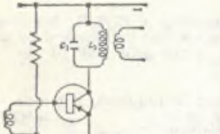


Fig. 5. A circuit illustrating reverse control a g c (after Amos)

The gain of a transistor falls at low collector currents and at low collector-emitter voltages and there are two corresponding ways of achieving gain control. The first is by applying the control voltage to the base as a reverse bias: this is known as *reverse control* and an example, using a pnp transistor and positive-going control bias, is given in Fig. 5. An npn transistor would, of course, require a negative-going bias for reverse control. For both types of transistor the control bias increases when a strong signal is received thus biasing back the transistor and reducing the gain. An unfortunate feature of this type of control is that the signal-handling capacity of a transistor is reduced by reverse bias but the circuit has the advantage that collector current is reduced when strong signals are received: this is important in battery-operated receivers where current economy is desirable.

In the second method, known as *forward control*, the gain of the transistor is reduced by increasing the forward bias, thus increasing collector current. An essential feature of the circuit, illustrated in Fig. 6, is the decoupling circuit R_2C_2: as the collector current increases, the voltage drop across R_2 is increased thus reducing the collector-emitter voltage and forcing the operating point to move into the knee of the I_c- V_c characteristics where the char-

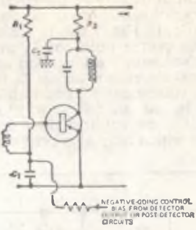


Fig. 6. A circuit illustrating forward control a g c (after Amos)

acteristics are more crowded and the g_m therefore lower. For this type of control a pnp transistor requires a negative-going bias and an npn a positive-going bias. Forward control has the advantage of increasing the signal-handling capacity of the transistor when this is needed for large-amplitude signals. Not all transistors are suitable for forward control and it is important to select a type which has been specifically designed for use in this type of circuit. By increasing the collector current of a suitable transistor from 4 to 13 mA, it is possible to reduce the gain by more than 40 dB.

Both methods of a.g.c. are in common use, sometimes in the same receiver. Both forms of control reduce the power output of the controlled stage. This is of little significance in early i f stages but it is not usual to apply a.g.c. to the final i f stage because this is required to supply appreciable power to the detector.

F.m. receivers centred on 10.7 MHz

Superhets are employed with 200 kHz wide i f band local oscillator on high side of v h f signal frequency; the f m signal contains spurious a m which must be removed in order to give freedom from interference. The majority of receivers use ratio detectors which have up to 30 dB of a m rejection but to supplement this it is common practice to make the final i f stage into an amplitude limiter stage.

F.m. detectors

The Foster-Seeley (phase) discriminator relies on the fact that the signal at the secondary of a double tuned i f transformer is 90° phase advanced on that at the primary at centre frequency (*fc*).

The Foster-Seeley discriminator contains two diode detectors so arranged that their outputs are connected in series opposition. The diodes are fed from a double-tuned transformer, the primary and secondary windings of which are resonant at the centre frequency of the pass-band to be covered. An essential feature of the circuit is that a fraction of the primary voltage is fed to the centre point of the secondary winding. In Fig. 7 this is achieved by a connection between the centre point of L_2 and a tapping point on L_1 but the secondary connection could be to the junction of two equal capacitors across L_2 (they could together constitute the tuning capacitance) and the primary connection could be to an inductor closely coupled to L_1 or to a capacitive potential divider across L_1 (formed by two capacitors which may also provide the tuning capacitance).

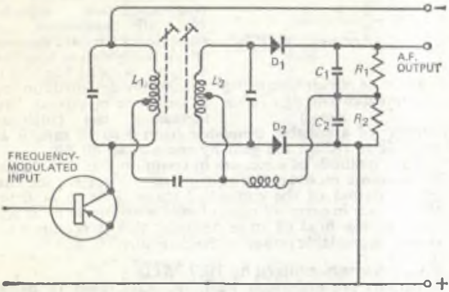


Fig. 7. One form of Foster-Seeley discriminator

For signals at the centre frequency, diodes D1 and D2 receive equal inputs and the voltages generated across R_1 and R_2 are equal, giving zero resultant voltage across $(R_1 + R_2)$. The effect of the interconnection between primary and secondary windings is that for signals displaced from the centre value one diode receives a bigger input than the other, and there is a net output across $(R_1 + R_2)$, the polarity depending on the direction of the frequency displacement and the magnitude depending on the extent of the displacement. If, therefore, a frequency-modulated signal is applied to the discriminator, a copy of the modulation waveform is generated across $(R_1 + R_2)$.

The Foster-Seeley discriminator gives zero output at the centre frequency. At other frequencies the output of the discriminator is proportional both to frequency displacement and to signal input. The Foster-Seeley discriminator has poor ability to reject a.m. signals and is normally used with a separate limiter stage.

Ratio detector

The ratio detector has much better a.m. rejection than a Foster-Seeley circuit and nearly all commercial receivers employ a ratio detector.

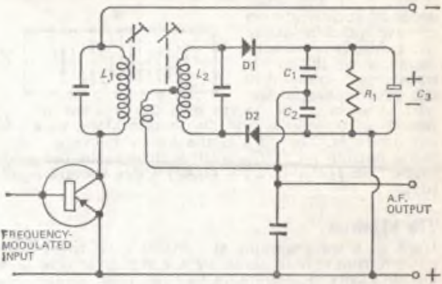


Fig. 8. One form of ratio detector (after Amos)

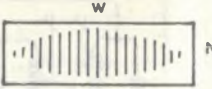
The circuit diagram of one form of ratio detector is given in Fig. 8. It has two diodes and a double-tuned transformer with a primary-secondary connection similar to that employed in the Foster-Seeley circuit but the diodes are connected in a series-aiding arrangement and supply a common load resistor. This resistor has a low value to give the heavy damping of the secondary tuned circuit on which the limiting properties of the detector depend. The diodes conduct continuously when a signal is applied to the detector and give a voltage across the load circuit proportional to the carrier input: The maximum value of this voltage gives an indication of the correct tuning point. The inputs to the two diodes vary with frequency displacement and the voltages generated across the reservoir capacitors C_1 and C_2 vary also although the total voltage across ($C_1 + C_2$) is independent of input frequency, being stabilised at a value proportional to carrier input amplitude by the long time constant R_1C_3

Pages 90-93 abridged by permission from Principles of Transistor Circuits *(4th Edition) by S. W. Amos, Iliffe*

Microwaves

Waveguides

As opposed to the electro-magnetic field in space (see p. 74) when both E and H fields are transverse to direction of propagation, either E or H has a component in the direction of the guide depending on orientation of wave launching device wrt guide. It is usual to designate by the purely transverse component viz. TE (transverse electric).



The subscript figures give the number of half periods of transverse field distribution along wide and narrow faces. The TE_{10} is the lowest frequency wave mode passing all $f > f_0 = c/2W$ with a λ_g in the guide found from $1/\lambda^2_g = (1/\lambda^2) + (1/\lambda_0^2)$ where λ_0 corresponds to f_0.

The klystron

Used as a cw generator at $f > 1000$ MHz to produce powers from mW to about 30 Kw at present time. The double cavity type employs two resonant cavities, the

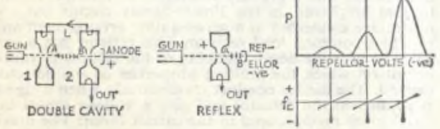


Fig.1 . Double cavity klystron

electron beam traversing lips of cavity 1, shock excites the latter, which in turn velocity modulates the beam electrons periodically, resulting in a bunching of the electron density in space AB. When this rf modulated beam current passes the lips of cavity 2, strong oscillation results, a portion of this being fed back via the loop L to sustain the original shock oscillation and hence bunching action by the first cavity. The reflex type employs only one cavity; a repellor electrode, negative with respect to the cathode, causes electrons emanating from the lips to decelerate and reverse back through the cavity. The bunching space is about twice AB, the one cavity acting as its own buncher and catcher. The repellor potential controls both phase and amplitude of return bunch, the former providing means of frequency modulating a klystron ($\simeq$ ·7MHz/volt), the latter controls output power P as shown.

Microwave Reception

Conventional rf amplification has no advantage above 600 MHz. Silicon diode mixer stage fed from aerial has lower noise, eg 10*db* at 3000 MHz. A typical mixer assembly is show in Fig 2.

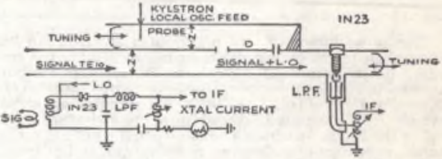


Fig. 2. Typical mixer assembly

The necessity to improve on F=10db has lead to development of several microwave rf amplifiers. One such device is thc travelling wave tube shown in Fig 3.

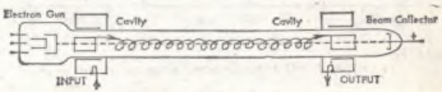


Fig. 3. Travelling wave tube

The input cavity induces the signal on to the helix. The helix field velocity modulates the electron beam. This modulation propagates at slightly less than the speed of the helix wave, interacting with it to produce amplification, the output being coupled via a terminating cavity. The performance (1969) of certain other devices is tabulated below:

System	*Frequency (KMHz/) limits*	*F (db)*	*Gain*	*Band width %*
Maser	·3—optical	$\simeq 0$	5-20	$\simeq 1$
Parametric				
(i) varactor diode	·3—10	3	10-20	10
(ii) electron beam	·3— 8	4	10-20	10
Travelling wave tube	1—30	4-25	10-30	10-20
tunnel diode	·1— 3	5	15	10

The noise factor (F, see p. 92) of a receiver can be quoted in terms of a noise temperature Te (°K). Te = rcvr, noise output/KBxrcvr. gain (when aerial input terminated with a resistor equalling aerial resistance). Now $F = 1 + (Te/To)$, $To = 290°K$ the standard temp. at which F is measured. Aerial noise temperature (Ta) °K is the temp. at which a resistor equal to aerial resistance produces the observed noise power at the rcvr., input. Whereas F as the noise factor assumes $Ta = 290°K$, at any other aerial temperature $F^1 = (F-1) + (Ta/To) = (Te + Ta)/To$.

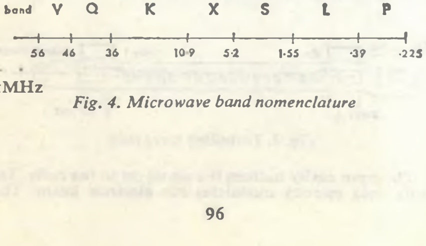


Fig. 4. Microwave band nomenclature

Television

625-LINE MONOCHROME

Nominal specification of transmitted signal

Channel width		8 MHz
Spacing between unmodulated sound and vision carriers		6 MHz
Vision modulation (am negative)		
upper sideband		5·5 MHz
lower sideband		1·25 MHz
synchronising level	As percentage of maximum vision carrier amplitude	100%
blanking level		77%
white level		20%
Sound modulation (fm)		
peak deviation		50 KHz
pre-emphasis		50 μs
Ratio of vision power during synch pulses to sound power		5 : 1
Lines per picture		625
Interlace		2 : 1
Field frequency		50 Hz*
Line frequency		15,625 Hz*
Approximate gamma of picture signal		0·5
Aspect ratio		4 : 3

* The transmissions are asynchronous; ie, the synchronising signals are derived from a stable oscillator and are not locked to the mains.

The idealised vision carrier amplitude as a function of time is shown in Figs. 1 and 2.

The vision carrier-amplitude waveform indicated in Figs. 1 and 2 represents the amplitude of a double-sideband am signal from which the transmitted vestigial-sideband signal is derived. Sideband frequencies more more than 1·25 MHz below the vision carrier are attenuated.

Test-line signal

A test-line signal is transmitted on lines 16 and 329. (When equipment modifications are completed, these will be changed to 18 and 331, to accord with international practice on test signals.) It consists of a 10-μs white bar containing an inverted sine-squared pulse (half-amplitude duration, 0·2 μs) and followed by an erect sine-squared pulse (half-amplitude duration, 0·2 μs), a chrominance pulse (half-amplitude duration, 1 μs) and a five-step staircase. The duration of each of the first four steps is 4 μs and that of the last step is approximately 3·5 μs. The steps are of nominally equal height

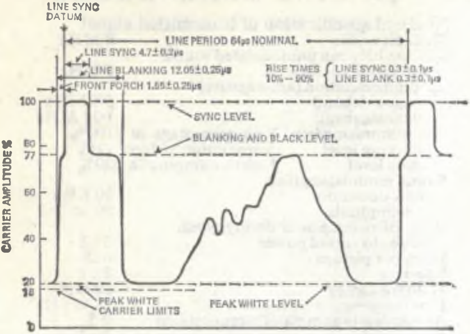


Fig. 1. Vision waveform showing line synchronising signals

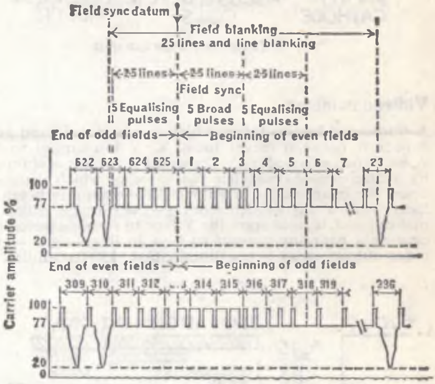
and the top step is at peak white. A colour sub-carrier signal, having a peak-to-peak value equal to the step height, is superimposed on the whole staircase.

At times when the 625-line network is carrying colour tests, those transmitters not radiating colour will transmit the test-line signal in a slightly modified form.

CAMERAS

Image orthicon camera

The image is focused on to the photo-cathode (see fig. 3). This emits electrons from A proportional to image brightness, the resulting electron density image is accelerated through a mesh B striking the target C. The target emits secondary electrons which are collected by the mesh (at small + ve voltage) leaving a + ve image pattern on the target face. The target is made of lateral conducting glass and the + ve image leaks to the rear face where the electron deficiency is made good by the low velocity scanning beam. The deficiency (picture) modulated beam returns towards the gun entering an electron multiplier, the magnified beam current being passed through a resistor developing a + ve video output. Scene illumination: 75 ft. candles.



Equalising pulses 2·4 μs
Interval between broad pulses 4·7 μs

Rise times 10%–90%: Field blanking 0·6 μs; Field syncs 0·3 μs; Equalising pulses 0·3 μs

Fig. 2. Vision waveform showing field synchronising signals

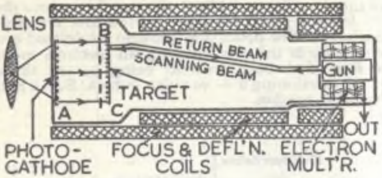


Fig. 3. Image orthicon camera

Vidicon cameras

A photo conductive target which can be considered as a capacitor between target faces X, Y is charged to 20 volts. When exposed each elemental capacitor is shunted by a discharging resistance the value of which depends upon the image point brightness. The scanning beam is faced with a charge deficiency (+ ve) image which it makes good, ie, recharges the Y face to cathode potential causing a charging current to flow in R giving a — ve video output signal. Scene illumination: 130 ft. candles.

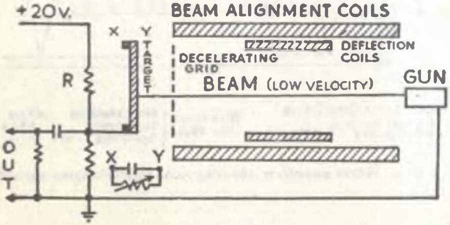
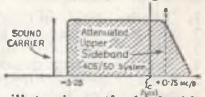


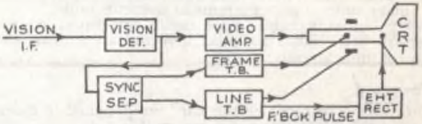
Fig. 4. Vidicon camera

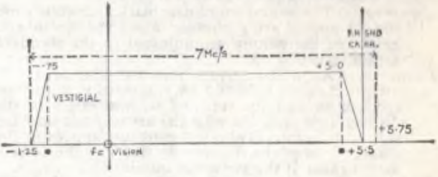
RECEIVERS

Superhet's with if about 11 MHz but higher, eg, 38 MHz in modern sets. Tuning accomplished by adjusting local oscillator frequency for loudest sound. The standard 405 line UK transmitted spectrum is vestigial side band to conserve channel space. The oscillator converts to the 2 ifs



separated by 3·5 MHz, if the oscillator is on the high side of carrier frequency Snd if = vis if + 3·5Hz. The CCIR European standard channel is 7 Hz wide as shown. The circuit schematic after the vision if is:





Besides the adoption of negative modulation for vision, the sound channel is fm, with maximum deviation of $\Delta f = \pm$ 50 KHz and $50 \mu S$ pre-emphasis. Sound to vision carrier spacing is 5·5 MHz, giving just over 5 MHz of vision equivalent to resolving 500 lines across the picture.

The CRT averages a 1,500 hour life, limitation due to residual gas being ionised by the beam. This results in ion bombardment of cathode destroying emissive surface and burn up of screen due to those ions which strike the latter. This minimised by (i) protective aluminised backing to screen (also improves brilliance), (ii) bending gun and employing ion trap magnet to deflect electrons but not ions on to correct line of approach to screen.

Receiver adjustments

1. Tuning—controls local oscillator frequency.
2. Contrast—alters rf/if gain.
3. Brilliance—controls grid bias of CRT.
4. Focus—controls size of spot on screen.
5. Line hold—alters frequency and/or degree of locking by line sync. pulse.
6. Frame hold—as above but applies to vertical scan.
7. Line/frame linearity—ensures spot scans raster with uniform velocity.
8. Picture height/width (line/frame amplitude)—controls magnitude of scan currents in deflector coils.

Note: Lethal voltages exist inside televisors; *do not* remove back unless the risks involved are known and clearly understood.

Test card

The various patterns on this card are designed to assess certain characteristics of the system thus:

Aspect ratio: The central concentric black and white rings should appear truly circular when the width and height of the picture are adjusted to the standard aspect ratio of 4 : 3.

Picture size: As most receivers have a display area with an aspect ratio of about 5 : 4, it is usual to adjust the receiver so that the top and bottom edges of the display area coincide with the arrowheads and the side castellations of the test card just appear in the display area of the receiver. In this way the correct aspect ratio of the picture is obtained.

Contrast: At the centre of the test card is a column of five squares with a contrast range of about 30 to 1 between the top and bottom squares. The difference in brightness between adjacent squares should be constant on a correctly adjusted receiver. Within the top and bottom squares are small lighter spots; white or black crushing is shown by the merging of the top or bottom spot into its surrounding area. The areas of the test card which are at peak white include the spot in the top square of the contrast pattern and the white background with the exception of the white vertical line in the black surround and the white surround of the black vertical line.

Resolution and bandwidth: At the sides of the contrast pattern are six gratings consisting of vertical stripes designed to produce, after gamma correction, signals of approximately sine-wave form corresponding to the following frequencies in MHz.

1·0, 1·5, 2·0, 2·5, 2·75 and 3·0

The range of brightness in the gratings is the same as that from the first (top) square to the fourth square of the contrast pattern; the brightest parts of the stripes have the same brightness as that of the area surrounding them.

Scanning linearity: The background of white lines should be reproduced in all parts as enclosing equal squares and the central black and white rings should appear truly circular.

Line Synchronisation: The border of the test card is a pattern of alternate black and white rectangles. The right side of this border serves as a test signal to check the line synchronisation of receivers. Faulty line synchronisation shows as horizontal displacement of those parts of the picture on the same level as the white rectangles in this side; it will also give the central rings the appearance of cog-wheels.

Low-frequency response: This can be checked by means of the black rectangle within the white rectangle at the top centre of the test card. Poor low-frequency response shows as streaking at the right-hand edges of the black and white areas and also of the border castellations.

Reflections: The white vertical line with the black surround and the black vertical line with the white surround should appear free from displaced images (ghosts). If there are reflections of the television signal, from hills or large buildings, these may result in displaced ghost images of any significant feature of the picture. This effect will be most readily seen as displaced images of the white and black vertical lines. The lines represent pulses having a duration of 0·3 microseconds.

Uniformity of focus: In each corner of the test card there is a diagonally-disposed area of black and white stripes; the focus of these areas and of the central area of the test card should be uniform. The stripes correspond to a fundamental frequency of about 1 MHz.

Colour television

The information in the following pages is based, with permission, partly upon information obtained from the BBC, partly upon Investigation Report No. L.113 by I. McWhirter of Thorn-AEI Radio Valves & Tubes Ltd., and partly upon Gouriet's monograph, An Introduction to Colour Television, published for the Royal Television Society by Norman Price (Publishers) Ltd.

The ability to transmit colour television signals in a three-colour system depends upon the fact, discovered by James Clerk Maxwell, that practically any spectral colour can be matched by mixing together (in suitable proportions) rays of three *primary* colours, red, blue and green. In a television camera the multi-coloured rays from a colour scene are filtered by dichroic mirrors—one of which will pass all rays but blue ones, which it reflects, another which will pass all but red rays, which it reflects. After filtering, the three resultant rays, respectively red, green and blue, are passed to the photo-cathodes of three television tubes built into one camera, which produce three output voltages proportional to the strength of the impinging rays. A typical arrangement of the dichroic mirrors is shown in Fig. 1, though relay lenses shown are usually only necessary in 4-tube colour cameras (p. 112).

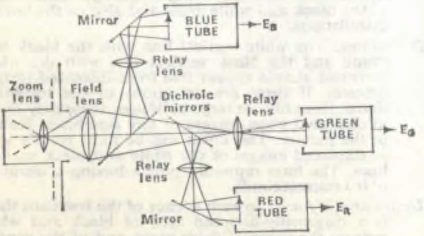


Fig. 1. Optical system of three colour camera with relay lens arrangement.

Because the human eye responds to different colour hues in a non-linear way—it is most sensitive to energy in the middle of the visible spectrum and responds only feebly to violet and deep red (*see Fig.* 2.)—the relative brightness content of the constituents of the mixture of red, (R), blue, (B), and green, (G), that makes white light is found to be approximately R 30%, G 60% and B 10%. In other words the luminance (brightness), Y, of white = 0·3R + 0·6G + 0·1B. These proportions hold good for *standard white* which is very similar to the average colour of the north sky and is called illuminant C'.

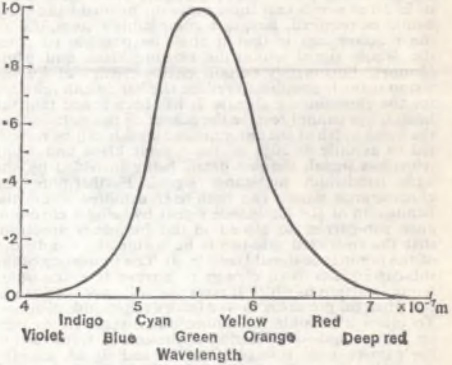


Fig. 2

Using another white as the standard would alter the relative proportions of R, B and G. For example, if the white chosen were similar to the colour of an illuminated slide projector bulb, the equation would be nearer to

Y (white) = 0·44R + 0·53G + 0·03B

The three properties of light which have to be transmitted are luminance, (brightness), hue (the colour wavelength) and saturation (strength). Luminance information is sent out on the main carrier and hue and saturation on a sub-carrier which is phase-modulated with the hue information and amplitude modulated with the saturation information.

Now the luminance signal alone—which corresponds to the normal black and white signal of monochrome transmissions, requires a bandwidth depending upon the number of lines per picture, the number of pictures per second and the fineness of the detail which is to be reproduced along each line. In the British 625/50 system the required video bandwidth is 5·5 MHz. It might have been expected that the hue and saturation (chrominance) signals would each occupy similarly wide bandwidths or in other words that three times the normal bandwidth would be required. But, in a compatible system, one of the requirements is that it shall be possible to place the whole signal within the existing black and white channel. Fortunately certain characteristics of human vision make it possible to reduce the bandwidth required for the chrominance signals. It has been found that the human eye cannot resolve the colour of fine details, thus the band width of the chrominance signals can be restricted to as little as 20% of the regular black and white television signal, the fine detail being provided by the wide bandwidth luminance signal. Furthermore the chrominance signals can both be transmitted within the bandwidth of the luminance signal by using a chrominance sub-carrier so placed in the frequency spectrum that the restricted side-bands lie within the bandwidth of the luminance signal (*see Fig. 3*). The frequency of the sub-carrier has been chosen to ensure that the interference pattern to which it gives rise will appear to cancel out when the picture is viewed from a reasonable distance. To make it possible for monochrome receivers to pick up a black and white version of the colour transmission the camera tube voltages (E_R, the red signal, E_G, the green signal and E_B, the blue signal), before being used for transmitter modulation, are encoded, that is converted into three combination signals:

(*a*) luminance, $Y = (0{\cdot}3E_R + 0{\cdot}59E_G + 0{\cdot}11E_B)$

(*b*) a red colour-difference signal $(E_R - E_Y)$ and

(*c*) a blue colour-difference signal $(E_B - E_Y)$.

The way in which the sub-carrier can be doubly modulated so that the two chrominance signals, $(E_R - E_Y)$ and $(E_B - E_Y)$, can be recovered in the colour receiver, without mutual interference, can best be explained by assuming the presence of two-sub-carriers of the same frequency but separated in time by 90°. In the phase alternation line (PAL) system the sub-carrier carrying $(E_B - E_Y)$ is regarded as the reference and that carrying $(E_R - E_Y)$ is therefore phase advanced upon the reference signal by a quarter of a wavelength (90°).

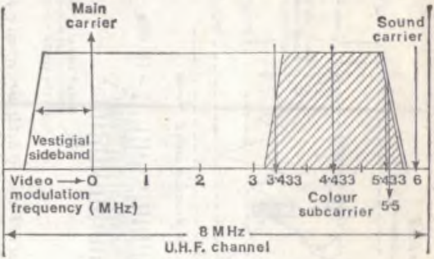


Fig. 3

At the receiver, the three voltages representing the primary colour signals, E_R, E_G and E_B, are recovered by addition, the luminance signal E_Y being applied to the cathodes of the three-gun CRT and the difference signals to the respective control grids, for example:

$$E_R = E_Y + (E_R - E_Y)$$
$$E_B = E_Y + (E_B - E_Y)$$
$$\text{and } E_G = E_Y + (E_G - E_Y).$$

Note: Although the $(E_G - E_Y)$ signal is not transmitted as such, it can be abstracted in the receiver, because $(E_G - E_Y) = -\frac{1}{2}(E_R - E_Y) - 1/5\,(E_B - E_Y)$.

Bearing in mind that the colour difference signals are severely bandwidth restricted compared with the luminance signal, a more accurate equation is

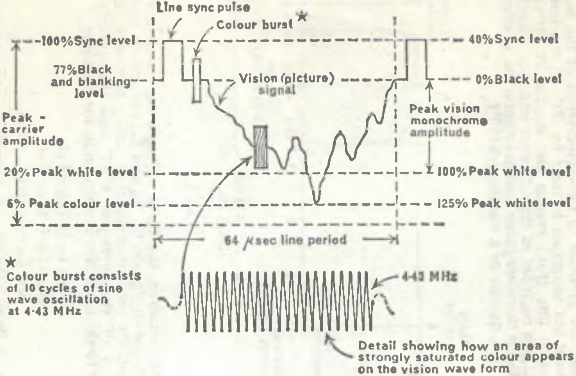


Fig. 4. Waveform of single line of 625-line, 50 fields per second colour transmission

$$E_R \text{ (recovered)} = [E_Y]_0^{F_{max}} + [E_R - E_Y]_0^{\frac{F_{max}}{5}}$$

$$= [E_R]_0^{\frac{F_{max}}{5}} + [E_Y]_{\frac{F_{max}}{5}}^{F_{max}}$$

Similar equations can be derived from the green and blue primary signals and they show that the reproduced image is in full colour up to detail components corresponding to $\frac{F_{max}}{5}$ and the fine detail is described entirely by the luminance signal.

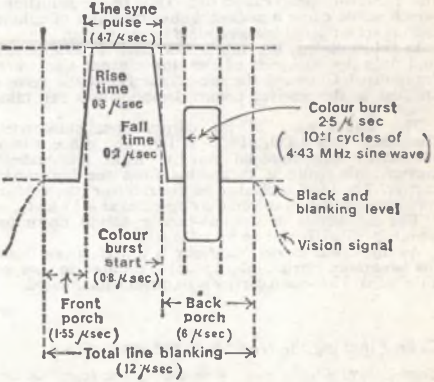


Fig. 5. Line sync and colour burst pulses in 625-line negative picture modulation colour transmission.

A danger of the double modulation system is that differential phase distortion in either the transmitter or the receiver may alter the apparent phase of the sub-carrier and this would result in a change in the hue of of the reproduced picture. This danger was overcome in the early (American) colour phase alternation system by changing the phase of the I signal by 180° every other line. This had the effect of causing the developed error signals to be of opposite sign, ie, if line one became too blue, the next line became too red. In this way the eye integrated the successive errors and appeared to see the true colour. The West German redevelopment of this idea is to cancel out the phase error electrically. This requires that in the receiver an electrical delay line, of time length of one scanning line, be used to provide the signal from the preceding line of the field. This is subtracted from an undelayed chrominance signal so that the error signals cancel out. Thus phase distortions which would cause a noticeable deterioration of colour quality are rendered imperceptible to the viewer.

In transmission, the sub-carrier itself is suppressed and only the sidebands of the chrominance signal are transmitted. Of course the sub-carrier has to be reconstituted in the receiver before demodulation can take place.

The actual value of the chrominance sub-carrier frequency, f_c is 4,433, 618·75 ± 1 Hz and, since it is a sub-carrier superimposed on the main (luminance) carrier, this figure is its spacing from the luminance carrier. The local oscillator in the receiver (the colour reference oscillator) accordingly operates at 4·43 MHz.

The side-bands of the sub-carrier extend down to about 3·433 MHz and up to 5·433 MHz.

As the sound carrier frequency is 6 MHz away from the luminance carrier, interference is reduced to acceptable limits. The sound carrier is frequency modulated.

Colour locking the receiver to the transmitter

Correct sampling is made possible by the transmission of a colour burst (consisting of 10 cycles of 4·43 MHz sine wave) on the back porch of the television signal which immediately follows the line sync pulse. In the receiver this burst is extracted from the signal and is used:

(*a*) to control the accurate timing of a sub-carrier regenerating oscillator operating at 4·43 MHz

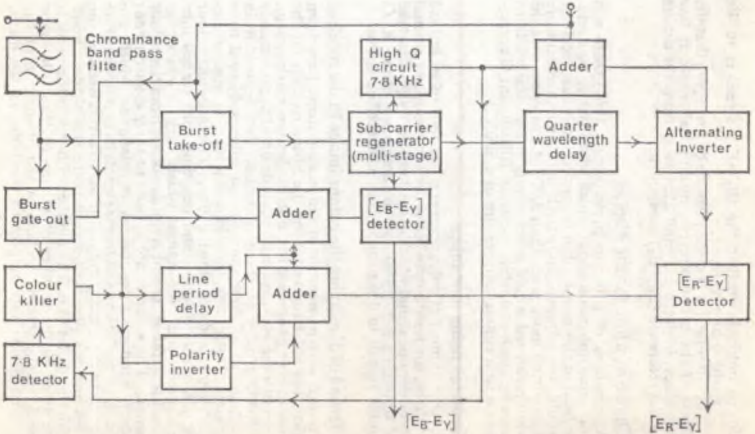


Fig. 6. Block diagram of PAL decoder. By courtesy of Thorn-AEI.

(*b*) to indicate whether the $(E_R - E_Y)$ signal is to be positive or negative.

The latter facility is made possible by arranging that the phase of the colour burst is made to alternate at line frequency, between 135° and 225° away from the timing of the reference $(E_B - E_Y)$ component.

Choice of white for the display

At the time of going to press (spring 1969) there was still some indecision about which white was to be standardised in this country but it seemed likely that receivers would be adjusted to match a colour temperature fractionally more green than 9,027° C, in order that the white achieved would more nearly match the white of existing black and white receivers. (Illuminant C corresponds to a colour temperature of 6,500°C.)

Cameras using four tubes

A large number of colour cameras use four tubes instead of three, the fourth tube being used to produce a normal monochrome signal which is used as the Y signal instead of obtaining a Y signal by the summation of the R, G and B signals in due proportions, as described above.

Method of displaying the colour image in the receiver

Though a large number of different systems have been tried, the most successful for the home receiver (at the time of going to press) is the RCA shadow mask system.

In this the viewing screen is made up of approximately 400,000 groups of three phosphor dots—one dot in each group of three glowing red when subjected to a stream of electrons, one glowing blue in similar circumstances and one glowing green. The incoming colour signals, after reception and decoding, are turned into red, green and blue signals which vary as exactly as possible as the R, G and B signal output from the colour camera. These signals are applied to three electron guns which produce three beams of electrons which are deflected by the line and frame signals in exactly the same way as in the normal monochrome receiver. Between the guns and the phosphor screen there is a shadow mask with tiny holes in it. The number of holes in the mask is the same as the number of groups of three phosphor spots on the screen. Because of the orientation of the focused beams with respect to the holes in the mask,

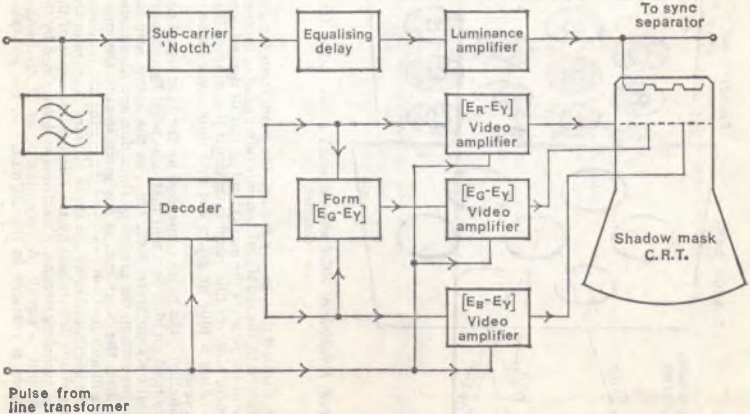


Fig. 7. Block diagram of PAL decoder and video stages. By courtesy of Thorn-AEI.

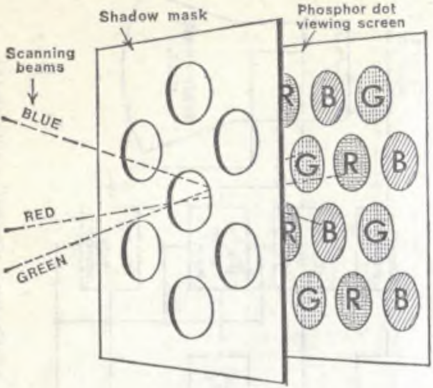


Fig. 8. Principle of RCA shadow mask colour tube. After Gouriet.

it is ensured that, at any instant, the beam from the gun operated by the red signal can only (after passing through one of the holes in the mask) land on one of the red dots. Hence, however this beam is deflected it can never land on either a blue or green dot and thus cause these to fluoresce.

Similarly, the beam controlled by the blue signal can only land on blue dots and that controlled by the green signal only on green dots. Thus, as the beams scan the viewing screen, the groups of three dots are illuminated one after the other and, at any given instant, the group being illuminated will convey to the eye the hue produced by the combination of the three primary colours in the proportions determined by the relative intensities of the three impinging streams of electrons as determined by the R, B and G signals. A very high order of mechanical accuracy is required in the alignment of the phosphor screen and difficulties had to be overcome in deflecting the electron beams accurately whilst still maintaining the convergence angles necessary for colour selection.

Compatability

The PAL colour television system is fully compatible. In addition to ensuring the reproduction of colour images at least on a par with accepted colour photographic systems, it has the following advantages:

1. it makes it possible to place the whole of the signal within existing black and white channels and therefore involves no change in established international planning agreements for channel positions, spacings and transmitter power and sitings.
2. as far as is practicable, it is possible to transmit the signal through existing television transmission equipment.
3. existing unmodified 625/50 black and white receivers can reproduce the colour transmission as an acceptable black and white picture.
4. colour receivers are able to reproduce an acceptable black and white image from an incoming black and white signal.

Gamma correction

For a pleasing reproduction of a scene, it is desirable that there should be a linear relation between light input to the camera and luminance of the display; ie, the gamma of the overall television chain should be 1·0. This can be achieved if the output signals from the camera, E_R, E_G and E_B undergo an inverse non-linearality to that of the display. In practice it has been found desirable not to compensate completely for the display and a commonly used inverse gamma index is 1/2·5.

Feeders and aerials

FEEDERS

The conductors of a feeder cable possess distributed self-resistance ($R, \Omega/M$), self-inductance (L, H/M), shunt capacitance (C, F/M) and shunt conductance (G, mho/M). The equivalent circuit is:

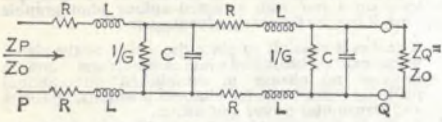


Fig. 1

The feeder exhibits a characteristic impedance (Z_0) ohms at P when termination at Q is also Z_0 ohms, ie, when matched $Z_0 = \sqrt{Z_{poc} . Z_{psc}}$ (Z_{poc} is Z_p with $Q_o/_c$, Z_{psc} is Z_p with $Q_s/_c$) giving $Z_o = \sqrt{(R+i\omega L)/(G+i\omega c)} \simeq \sqrt{L/C}$ closely at rf. Physical dimensions governs L and C hence Z_0:

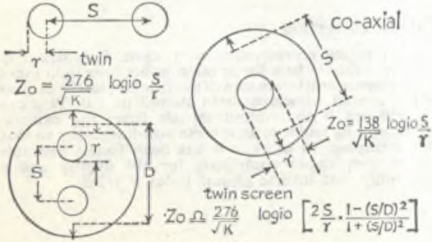


Fig. 2

Dielectric constant, $K = 1{\cdot}0$ for air, $2{\cdot}3$ for polyethylene. Propagation constant, $\gamma = \sqrt{\gamma (R+i\ \omega L)\ (G+i\ \omega C)} =$

$\alpha + i\beta$, α the attenuation constant in db/M, β the phase constant in radians of phase delay per metre. When $Z_Q \neq Z_0$, ie is mismatched the travelling wave from P is partially reflected back from Q; σ = reflected volts (V_R)/ travelling volts ((V_T) = $(Z_Q - Z_0)/(Z_Q + Z_0)$. V_R adds with V_T to give a standing wave distribution of rms rf voltage (V_S) with peaks (antinodes) and minima (nodes), the nearest antinode to Q is at $X = (2\pi - \theta)\ \lambda/4\pi$ where θ = angle of reflection coefficient σ, λ is wavelength on line. The standing wave ratio (swr) = $V_S$max/ $V_S$min. = $(1 + |\sigma|)\ (1 - |\sigma|)$ indicates degree of mismatch, is 1·0 when $\sigma = 0$ ie $Z_Q = Z_0$. Complete feeder equations are:

$$V(x) = V_Q \cosh \gamma x + I_Q Z_0 \sinh \gamma x.$$

$$I(x) = (V_Q/Z_0) \sinh \gamma x + I_Q \cosh \gamma x.$$

giving impedance at distance x from Q:

$$Z(x) = Zo.(Zq + iZo \tanh \gamma x)/(Zo + iZq \tanh \gamma x) \text{ ohms.}$$

When low loss cable is employed, hyperbolic functions can be replaced by corresponding trigonometric function. When Q s/c,

$$Zx = Zo \tanh \gamma x, \quad \delta = -1{\cdot}0$$

when Qo/c,

$$Zx = Zo \coth \gamma x, \quad \delta = 1{\cdot}0.$$

Variation of $Zp\ sc$ with x is shown for short at $x = O = Q$. For $x = \lambda/4$ $Zpsc$ (very high) $= Zo \coth \alpha x$ resistive $\simeq$ $Zo.8fL/R$ ohms. f is frequency. L/R is inductance/resistance ratio of feeder. At $\lambda/2$ $Zpsc$ = zero ohms again, but is reactive as shown between $\lambda/4$ points.

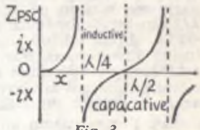


Fig. 3

Note that λ on line is approximately $1/\sqrt{K}$ of free space λ, (for polyethylene 68% of free space λ). Coaxial covers range 50 – 250 Ω, twin 200 – 500 ohms. Devices for matching Z_O to Z_Q:

(*a*) $\lambda/4$ transformer; insert a $\lambda/4$ section of cable of impedance $Zo_1 = \sqrt{Zq\, Zo}$ between feeder and Z_Q.

(*b*) stub match; when $x = (\lambda/2\pi) \tan^{-1} \sqrt{Z_Q/Z_0}$ metres, or feet distant from Q, Zx comprises $R = Zo$ paralleled by a reactance $\pm iX$.

An s/c or o/c stub is tee'd in at this distance, stub length and hence reactance adjusted to cancel the $\pm iX$ on the line at x.

AERIALS

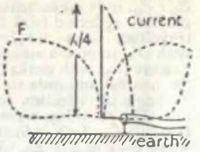


Fig. 4

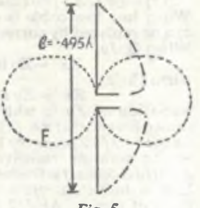


Fig. 5

The direction of electric flux lines defines the polarisation of received or transmitted wave, eg, horizontal or vertical. The simplest aerial is the vertical unipole. $Z_Q = 40\Omega$ azimuth polar diag. omni-directional, elevation pattern as for F. used at mf, lf on broadcast working—vertical polarisation. The $\lambda/2$ dipole widely employed from hf to shf shown for vertical polarisation—$Z_Q = 75\Omega$ resistive ($Z_Q = 73 + j42{\cdot}5$ ohms when $l = \lambda/2$). Azimuth polar diag. omni-directional, elevation—as for F. Z_Q varies with height for horizontal $\lambda/2$ dipoles as shown converging to 73 ohms in free space. When $l = {\cdot}925\lambda$, pattern is narrower version of F for $\lambda/2$ dipole; Z_Q for horizontal $\cdot925\lambda$ dipole = 3,200 ohms (using 12swg wire). For $l = 1{\cdot}425\lambda$, $Z_Q =$ 100 ohms and elevation pattern as above—azimuth pattern omni-directional. For television and vhf reception a $\lambda/2$ dipole connected via 80Ω feeder is satisfactory within 15 miles of the transmitter. Beyond this distance, increased

Fig. 6

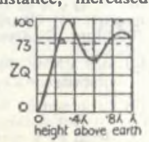


Fig. 7

sensitivity can be obtained by use of a parastic reflector behind the dipole. The reflector is 5% longer than the dipole. At 50 miles additional gain obtained by adding a

director in front of dipole, 5% shorter in length. Besides increasing the sensitivity, the polar diagram narrows down, attenuating unwanted ghost signals and interference.

Horizontally polarised TV and vhf transmissions require horizontal dipoles, etc.

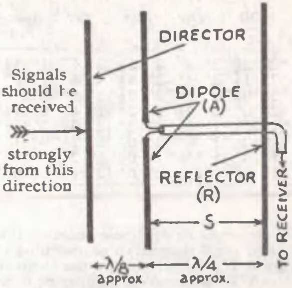


Fig. 8 shows a simple type of 405 line television aerial. Though a folded dipole would offer slight advantages, these would most likely be offset by poor impedance matching between the aerial, down lead and set input.

Fig. 8. 405 line television aerial details

Service	*Frequency MH₂*		*Aerial dimensions*		*Separation S*
	From	*To*	*A*	*R*	
CH1	45·0	41·5	10′ 9½″	11′ 1½″	5′ 3″
CH2	51·75	48·25	9′ 4½″	9′ 8½″	4′ 11″
CH3	56·75	53·25	8′ 6½″	8′ 9½″	4′ 6″
CH4	61·75	58·25	7′ 10″	8′ 1″	4′ 1″
CH5	66·75	63·25	7′ 2½″	7′ 5″	3′ 9½″
VHF/FM	87·5	100	5′ 0″	5′ 3″	2′ 8″
Band III	179·75	211·25	2′ 3½″	2′ 5″	1′ 2″
Band IV	471·25	581·25	1′ ½″	1′ 2″	0′ 6¼″
Band V	615·25	853·25	0′ 9½″	0′ 10½″	0′ 4¾″

Ghosting on band III is much worse than on band I hence a narrow polar diagram is essential. This can be obtained by adding further directors spaced ·2λ apart and each 5% shorter than the previous director element as in table below:

Ch.	*Dir* 4	*Dir* 3	*Dir* 2	*Dir* 1	*Dip*	*Ref*	*Spacing*
6	2′ 0¼″	2′ 1⅛″	2′ 2¾″	2′ 4″	2′ 5½″	2′ 7″	1′ 1⅛″
7	1′ 11¼″	2′ 0½″	2′ 1¾″	2′ 3″	2′ 4½″	2′ 6″	1′ 0″
8	1′ 10¼″	1′ 11½″	2′ 0½″	2′ 1¾″	2′ 3½″	2′ 5″	0′ 11¾″
9	1′ 9¾″	1′ 11″	2′ 0″	2′ 1¼″	2′ 2¾″	2′ 4½″	0′ 11½″
10	1′ 9¼″	1′ 10½″	1′ 11½″	2′ 0¾″	2′ 2″	2′ 3½″	0′ 11¼″
11	1′ 8¾″	1′ 9¾″	1′ 10¾″	2′ 0″	2′ 0″	2′ 2½″	0′ 11″
12	1′ 8¼″	1′ 9″	1′ 10″	1′ 11½″	2′ 0¾″	2′ 2″	0′ 10¾″
13	1′ 7½″	1′ 8″	1′ 9½″	1′ 10¾″	2′ 0″	2′ 1¼″	0′ 10½″

Due to adding more elements *Zq* of dipole becomes low (20Ω for 3 element, 15Ω for 4 element) and matching to 80Ω co-axial is not good. If a folded dipole is used instead of normal type, $Zqf = N^2 Zq$ where *N* is number of folds and match is improved as $Zqf = 4Zq$ (2 fold) $Zqf = 9Zq$ (3-fold).

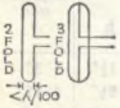


Fig. 9 Folded dipoles

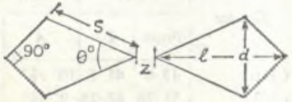


Fig. 10 Cone aerial

The cone aerial (shown for horizontal polarization) of wide band width $\pm ·15\lambda$, $l = 1·025\lambda$, $d = ·75\lambda$, $S = ·75\lambda$, $Z = 160\ \Omega$, $\theta = 60°$. Cones can be made of 12 to 16 wires soldered at base ring and apex; 1/2 cone aerial operates if one cone perpendicular to λ·4 radius sheet of narrow wire mesh, when $Z = 80\Omega$. The slot aerial comprising a

rectangular slot, w (between $\lambda/12$ and $\lambda/10$), $d=3w$ cut in wire mesh.

$Zqs=(377)^2/Zq$ of corresponding length dipole;
for $l=\lambda/2$
$Zqs=365-j200$ ohms;
for $l=\cdot475\ \lambda$ and $w=\cdot01\lambda$
$Zqs=530$ ohms resistive;
for $l=\cdot925\lambda$, $w=\cdot067\lambda$
$Zqs=50$ ohms resistive.

Except in the last case, match to co-ax requires method shown or co-ax connected across slot at $x)\lambda/20$.
Note: Slot shown is for vertically polarised wave operation.

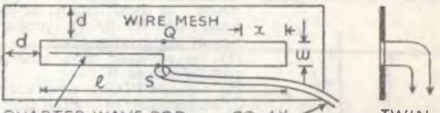


Fig. 11 Slot aerial

For omni-directional vhf communication the $\lambda/4$ ground plane ($Z=40\Omega$) is often employed. When at some height this aerial may be dc earthed for lightning protection by means of a $\lambda/4$ stub.

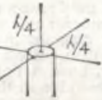


Fig. 12 Co-axial

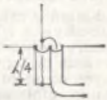


Fig. 13 λ/4 stub

Fig. 14 Discone

The discone is a wide band $Z=50\Omega$ aerial having omni-directional horizontal coverage and with 20% centre frequency band width.

Reflector type aerials

The reflected radiation from a shaped conducting surface can produce narrow beams whilst retaining the simplest forms of primary radiator, eg the dipole.

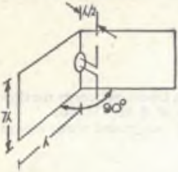


Fig. 15 Corner reflector

Corner reflectors (Fig. 15) may be made up of spaced conductors if spacing $< \cdot 1\lambda$. Shown for vertical polarization, beam width 50° in azimuth, with a gain of 10*db* at $\theta = 0°$ and Z is 70Ω, not a wide band structure.

The parabola (Fig. 16)—geometrical definition $y = x^2/4f$, where f is focal length; at the focal plane $W = 4f$. The properties of importance are:

(a) all reflected rays emerge parallel across W.

(b) distance PQR is constant $= 2f$ independent of $\angle PQR$ hence all waves emerge from W in phase. Assuming a source at A causes uniform illumination across W, the polar diagram for a circular dish is a pencil beam. The first zeros occur at $\pm\theta° = 70 \sin 0{\cdot}61\lambda/W$, half power points $\pm\ \theta° = 31/\lambda W$, side lobes 13% ($-17db$) of the main beam intensity. The gain is a maximum for uniform illumination of W, $G_0 = 4\pi A/\lambda^2$ where A is mouth area in units consistent with λ.

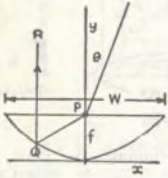


Fig. 16 Parobolic aerial

As the primary radiation is expanding until reflection occurs, the electric intensity is $\propto 1/PQ$. Hence for uniform illumination the primary radiator polar diagram must increase intensity towards dish rim and then cut off at rim to avoid spill over waste of power. Such a polar diagram is virtually impossible and undesirable in that unwanted side lobe level is high. With the small primary radiating systems employed, illumination falls off slowly towards the rim, giving average results $G_e = 60\% G_0$ and $\pm\ \theta° = 36{\cdot}3\lambda/W$ half power bearings with first lobes of 6% (—25*db*). The primary feed may be a dipole providing both arms carry equal rf current; if not the main beam will squint off axis. A good feed is a flared wave guide radiator which gives fairly uniform illumination over W.

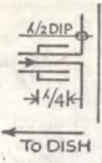


Fig. 17

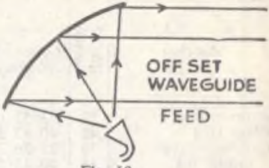


Fig. 18

Aerial gain (G)

Any radiation attenuates in space according to the inverse square law; the hypothetical isotropic aerial radiates an expanding spherical wave; for P watts to aerial power density $S_I = P/4\pi d^2$ watts/m^2 at d metres. A directional aerial would radiate $S = PGF\ (\theta\phi)/4\pi d^2$ watts/m^2 where F is the directivity pattern (having a maximum value of unity at the bearing ($\theta\phi$) of maximum transmission). The gain in any direction ($\theta\phi$) is then $(S/S_I) = (GF\ \theta\phi)$, but the definition of gain relates only to the direction of maximum transmission. *Gain is the factor by which the received power density from a directional aerial exceeds that given by an isotropic aerial.*

Capture area (A)

The power available in the correctly matched terminating resistance (=radiation resistance) of a receiving aerial is $P = AS$ watts, where S in W/m^2, and A is in m^2. The relationship between A and G is $G = 4\pi\ A/\lambda^2$, this formula is useful in estimating space communication loss.

Aerial		*G*	*A*
Isotropic		1	$\lambda^2/4\pi$
Half wave dipole		1·64	$1{\cdot}64\ \lambda^2/4\pi$
Horn	*Physical* area A_P	$5{\cdot}5\ A_P/\lambda^2$	$\cdot 45 A_P$
Parabola or lens		$6{\cdot}3$ to $7{\cdot}5\ A_P/\lambda^2$	$\cdot 5$ to $\cdot 6 A_P$
Broad side array		$4\pi A_P/\lambda^2$	A_P

Ultra high frequencies

SOME OF THE FIRST UHF 625-LINE TV STATION
Relay stations indented under main station of group

Station	*BBC 2 channels*	*Other uhf channels*	*Horizontal /vertical polarization*	*Max vision erp*
Belmont	28	53 57 60	H	500 kw
Black Hill	46	40 43 50	H	500 kw
Crystal Palace	33	23 26 39	H	500 kw
Guildford	46	40 43 50	V	2·5 kw*
Hertford	64	54 58 61	V	500 w*
Reigate	63	53 57 60	V	2·5 kw*
Turnbridge Wells	44	41 47 51	V	10 kw*
Divis	27	21 24 31	H	500 kw
Dover	56	50 53 66	H	100 kw*
Durris	28	22 25 32	H	900 kw
Elmley Moor	51	41 44 47	H	100 kw*
Llanddona	63	53 57 60	H	100 kw*
Oxford	63	53 57 60	H	500 kw
Pontop Pike	64	54 58 61	H	500 kw
Rowridge	24	21 27 31	H	500 kw*
Sudbury	44	41 47 51	H	250 kw
Sutton Coldfield	40	43 46 50	H	1000 kw
Brierley Hill	63	53 57 60	V	10 kw*
Bromsgrove	27	21 24 31	V	4 kw*
Kidderminster	64	54 58 61	V	2 kw*
Lark Stoke	26	23 29 33	V	10 kw*
Talcolneston	55	59 62 65	H	250 kw
Wenvoe	51	41 44 47	H	500 kw
Aberdare	27	21 24 31	V	125 w*
Kilvey Hill	26	23 29 33	V	2·5 w*
Pontypridd	28	22 25 32	V	500 w*
Winter Hill	62	55 59 65	H	500 kw

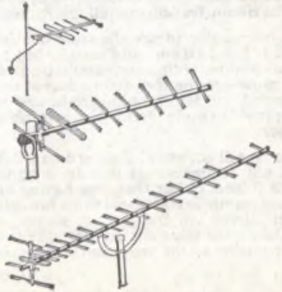
** Directional aerial*

U.H.F. receiving aerials†

The ultra high frequency range for television extends from 470 to 854 MHz. It is divided into two bands, band IV (470 to 582 MHz) and band V (614 to 854 MHz).

Band IV contains 14 channels (21-34) and band V 30 channels (39-68). BBC2 from Crystal Palace is on Channel 33 and from Sutton Coldfield on Channel 40. Typical uhf receiving aerials are shown in Figure 19. They consist of multi-elements, based on a half-wave dipole. Quite near to a transmitter, a simple five-element receiving aerial is usually sufficient. Viewers in reasonably favourable locations within 20 or 30 miles of a high power station will generally find a nine-element aerial satisfactory. In fringe areas an aerial of up to 20 elements may be needed and this type may also be necessary in difficult locations closer to the transmitter. The greater the number of elements, the greater the pick-up power and directivity. This enables reflections to be suppressed to a great extent, but much more care in placing a uhf aerial is required. Hills and tall buildings cast rather sharp radio shadows and may reflect the waves. These effects, which cause appreciable variations in signal strength are much more noticeable than they are in bands I and III.

Indoor aerials and set top aerials are likely to give poor and variable results and are not in general suitable for uhf reception, though in very favourable locations, near to a transmitting station, they may prove sufficient. Even in such locations, the quality of the picture received would probably be improved by the use of a relatively simple outside aerial. In locations directly in the shadow



Simple five element aerial for use near to a transmitter

Nine element aerial, range twenty-thirty miles

Multi-element aerial suitable for fringe areas

Fig. 19 Typical u h f aerials

of a large steel-framed building or immediately screened from the transmitter by a steep hill, really good reception of the uhf transmission may be impossible.

Because of the directional property of uhf aerials (the degree being dependent upon the number of elements in the aerial) the receiving aerial should usually be directed accurately at the transmitter, care being taken to mount it in a position in which there is an unobstructed line of sight to the transmitting aerial. If this is not practicable, the aerial should be placed in a high, open position with as few obstructions in the direction of the transmitter as possible. Obstructions close to the receiving point should be avoided.

It may however be possible to receive a satisfactory picture in a shadow area by pointing the aerial at a large building, outside the shadow area. The building may act as a reflecting surface, but pictures received in this way tend to be less sharp than those received directly and to vary according to weather conditions.

When a new aerial system is being installed which includes aerials for bands I and III as well as for uhf it is usually best to put the uhf aerial at the top of the mounting and to space it at least two feet and preferably four feet from the other aerials, gutters and roof surfaces.

The aerials should be connected to the receiver by low loss coaxial cable. Flat ribbon feeder is unsuitable for uhf.

It will sometimes be possible to use the same coaxial down lead for band I, band III and uhf aerials, but in this case a special junction unit (diplexer) should be used at each end of the cable—at the top end to accept the feeders from the bands I and III and uhf aerials, and at the bottom end to provide two leads for the two input sockets on the receiver.

It is important to install an aerial designed to work satisfactorily on all uhf channels that may be used to serve the area in the future, rather than one having an optimum performance on the first channel to be brought into use. In general all the uhf transmitters serving a particular area will be on the same site so that the same aerial, if suitable, can receive all the programmes radiated on uhf.

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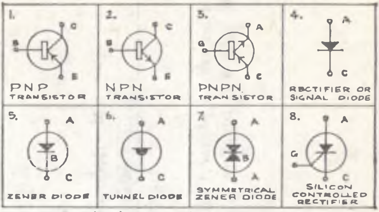
†Abridged with permission from a BBC publication *How to Receive BBC2 and Colour*.

Semi-conductor devices



A-ANODE; B(1&2)-BASE; B(5&7)-BREAKDOWN DEVICE; C (1&2)-COLLECTOR; C(4, 5, 6, & 8)-CATHODE; E-EMITTER; G-GATE;

Fig. 1. Standard symbols

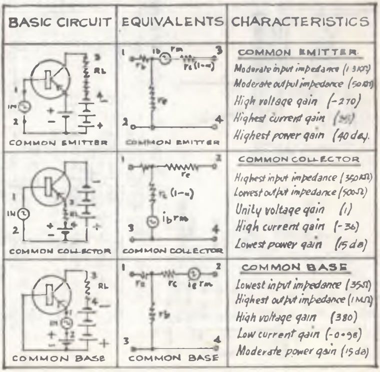


Fig. 2. Equivalent circuits

Transistor data

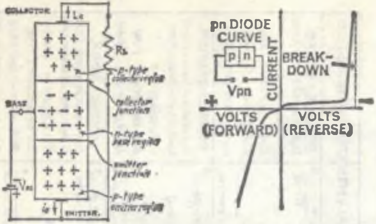


Fig. 3. Conduction in a PNP junction transistor (common emitter connections)

FORMULAE FOR TRANSISTOR CIRCUITS

Exact formulae	*Approximate formulae*
$Z_{11} = re + rb.\dfrac{rc(1-\alpha) + rL}{rc + rL}$	$Z_{11} = re + rb(1-\alpha)$
$Z_{22} = \dfrac{rc.\,re + rb(1-\alpha) + rs}{re + rb + rs}$	$Z_{22} = rc$
$\dfrac{V_{22}}{V_{11}} = \dfrac{\alpha\, rc\, rL}{rc\,re + rb(1-\alpha) + rL(re+rb)}$	$\dfrac{V_{22}}{V_{11}} = \alpha\, rL/Z_{11}$
$\dfrac{i_{22}}{i_{11}} = \dfrac{\alpha}{1 + rL/rC}$	$\dfrac{i_{22}}{i_{11}} = \alpha$

Exact formulae	*Approximate formulae*
$Z_{11} = rb + re.\dfrac{rc + r_L}{rc(1-\propto) + r_L}$	$Z_{11} = rb + \beta re$
$Z_{22} = rc(1-\propto) + re.\dfrac{\propto rc + rs}{re + rb + rs}$	$Z_{22} = re/\beta$
$\dfrac{V_{22}}{V_{11}} = \dfrac{\propto rc r_L}{rc[re + rb(1-\propto)] + r_L(re + rb)}$	$\dfrac{V_{22}}{V_{11}} = \dfrac{\beta rc}{Z_{11}}$
$\dfrac{i_{11}}{i_{22}} = \dfrac{\propto}{1-\propto + r_L/rc}$	$\dfrac{i_{22}}{i_{11}} = \beta = \dfrac{1-\propto}{\propto}$
$Z_{11} = rb + rc.\dfrac{re(1-\propto) + r_L}{re + r_L}$	$Z_{11} = \dfrac{r_L}{1-\propto}$
$Z_{22} = re + rc(1-\propto).\dfrac{rs + rb}{rs + rc}$	$Z_{22} = re + (rb + rs)(1-\propto)$
$\dfrac{i_{22}}{i_{11}} = 1/[(1-\propto) + r_L/rc]$	$\dfrac{i_{22}}{i_{11}} = 1/(1-\propto)$
$\dfrac{V_{22}}{V_{11}} = \dfrac{r_L}{re + rb(1-\propto) + r_L}$	$\dfrac{V_{22}}{V_{11}} = 1$

In these formulae, *rs*, *r*L, *re*, *rb*, *rc*, *rm* are source, load, emitter, base, collector and transfer dynamic resistance parameters resp. $\propto = ic/ie$, $\beta = ic/ib = \propto/1(1-\propto)$, $\propto < 1\cdot0 \simeq \cdot98$ in) junction types, *re* inversely proportional (i) to *Ie* (dc emitter current) and (ii) to collector voltage *Cc* up to about —3*V*, independent thereafter; *rb* comprises two parts (i) intrinsic (interface) (*rb*) inversely proportional to *Ie*, but proportional to *Vc*; (ii) extrinsic *rb′* independent of biases; *rc* inversely proportional to *Ie*. Collector/base capacitance (*Cc*) 5-20*pF* important at R.F.;   varies with *Ie* as shown. Noise increases with *Ie* and is bass heavy increasing below $\simeq$ 100Hz; noise independent of *Vc* when $Vc < -5v$.

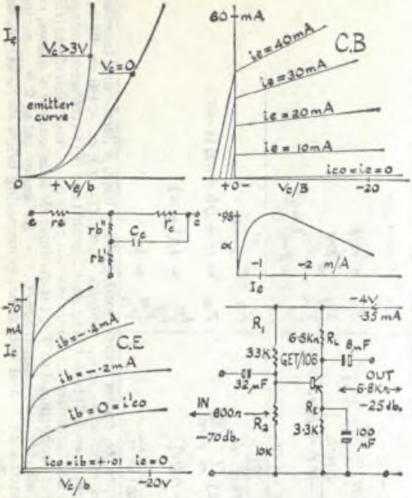


Fig. 4. Pre-amplifier by courtesy of G.E.C.

CB resembles grounded grid and *CC* the cathode follower. Most circuits *CE*. For pre-amps non-linearity due to *re* is small and *ic* (*ie*) reduced to improve noise; compromise between $\propto$ and noise, eg, $\beta = 25/45$ $ic \simeq \cdot 3\ mA$ with $Vc < -5$ say $-2v$. Choose r_L by battery voltage and reduced β to $\beta e = \beta/(1 + r_L/rc(1-\propto))$ eg, $r_L \simeq 5K\Omega$. $\beta e \simeq 85\%\ \beta$. Medium gain stages compromise output versus dissipation. *Ico* ($10\mu A$) the reverse collector leakage current gives in *CB* $Ic = (\propto Ie + Ico)$ important but in CE, $Ic = (\beta Ib + I'co)$, $I'co = (1+\beta)\ Ico$; since β large and *Ico* doubles every 9C° bias point is not stable.

Stability factor $S = dIc/dIco$ (1·5 pre-amps 7 medium gain) $S = 1 + Re/Rb/ (1 - \propto + Re/Rb) \simeq (1 + Rb/Re)$ where Rb is R_1 parallel R_2. Coupling—when RC coupling condenser $> 10\,\mu F$(low Z_{11}, Z_{22}). Volt drop in dc winding resistance of transformers must be small.

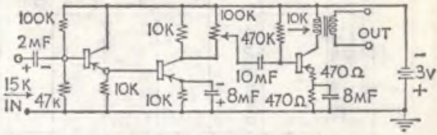
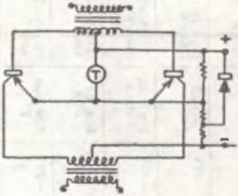
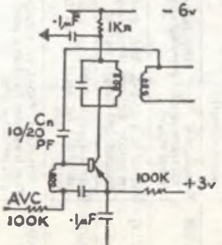


Fig. 5. Typical amplifier chain





Power stages mostly class B push-pull; knowing Po (output power), V_B (battery voltage), r_L to each transistor, $r_L = (V_B - 1)^2/2Po$ ohms.

To avoid threshold distortion forward bias bases by $-0{\cdot}15V$, this should decrease by 2·5 $mV/C°$ to compensate for emitter/base temp. Variation. (Thermistor on transistor (T).) Stabilise against V_B by diode ($SX641$ forward $SX56$ Zener) RF application—$\propto$ falls with frequ., $\propto(f) = \propto/(1 + jf/f\propto o)$ $f\propto o$ is alpha cut off. β cut off is $f\beta o = f\propto o\,(1 - \propto)$. Neutralisation of $Cc\ rb'$ necessary.

Type	*Description*
AA119	**Germanium point—contact diode**
AC107	**Germanium P-N P alloy junction transistor for use in low noise applications**
AC126	**Germanium P-N-P alloy junction transistor for pre-amp and driver stages**
AC127	**Germanium N-P-N high gain transistor for complimentary symmetrical class B output**
AC128	**Germanium P-N-P high gain transistor for class A and B output stages**
AD149	**Germanium P-N-P alloy junction transistor for class B push-pull output stages**

SEMI-CONDUCTOR DATA

by courtesy of Mullard Ltd.

VCB *max*	Ic(AV) *max*	*Ptot max*	Current amplification factor (*common emitter*)	
			Small signal	*Large signal*
max reverse V Peak 45 Av 30 (V)	Max Frd I Peak 100 Av 35 (mA)			
—15V	5 mA	(Tamb = 25°C) 80mW	(Ic = 300μA) 60	
—32V	100mA	(Tj = 75°C) 500mW		140
+ 32V	ICM max 500mA	(Tamb = 25°C) 340mW		(Ic = 500mA) 50
(IE = 0) —32V	ICM max 1·0A	(Tamb = 45°C) 155mW		(IE = 300mA VCB = 0) 60 to 175
—50V	ICM max 3·5A	Tamb = 50°C) 22·5W		(Ic = 1·0A) 30 to 100

AD161	Germanium N-P-N alloy junction transistor with AD162 form complementary pair	($I_E=0$) +32V
AD162	Germanium P-N-P alloy junction transistor with AD161 form complementary pair	($I_E=0$) —32V
AF114	Germanium P-N-P alloy-diffused transistor for rf amp. in am and fm receivers	($I_E=0$) —20V
AF115	Germanium P-N-P alloy-diffused transistor for mixer/oscillator for am/fm receivers	—20V
AF116	Germanium P-N-P alloy-diffused junction transistor for if amplifiers in fm receivers	—20V
AF117	Germanium P-N-P alloy-diffused junction transistor for mixer/oscillator and if amplifier in am receivers	—20V
AF118	Germanium P-N-P alloy-diffused transistor for video amplifier in television receivers	($I_E=0$) —70V
AF124	Germanium P-N-P alloy-diffused junction transistor for rf amplifier in am and fm receivers	($I_E=0$) —20V

I_{CM} max 3A	($T_{amb} \leq 72°C$) 4W		($V_{CE} = +1V$ $I_C = 500mA$) 80 to 320
I_{CM} max 3A	($T_{amb} \leq 63°C$) 6W		($V_{CE} = 1V$ $I_C = 500mA$) 80 to 320
I_{CM} max 10mA	($T_{amb} \leq 45°C$) 5mW		
I_{CM} max 10 mA	($T_{amb} = 45°C$) 50mW	150	
I_{CM} max 10 mA	($T_{amb} = 45°C$) 50mW	150	
I_{CM} max 10 mA	($T_{amb} = 45°C$) 50mW	150	
I_{CM} max 30 mA	($T_{amb} = 45°C$) 250mW		180
I_{CM} max 10 mA	($T_{amb} = 30°C$) 60mW		

Type	*Description*	V_{CB} *max*	I_C (AV) *max*
AF126	**Germanium P-N-P alloy-diffused transistor for if amplifier in fm receivers**	**($I_E = 0$) —20V**	**I_{CM} max 10 mA**
AF127	**Germanium P-N-P alloy-diffused transistor for mixer/oscillator and if amplifier in mw and lw receivers**	**($I_E = 0$) —20V**	**I_{CM} max 10 mA**
AF178	**Germanium P-N-P alloy-diffused transistor for mixer/oscillator at frequencies up to 260 MHz**	**($I_E = 0$) —25V**	**I_{CM} max 10 mA**
AF179	**Germanium P-N-P alloy-diffused transistor for large signal if amplifier in television receivers**	**—25V**	**I_{CM} max 15 mA**
AF180	**Germanium P-N-P alloy-diffused transistor for R.F. amplifier in tel. tuners. f up to 220 MHz**	**($I_E = 0$) —25V**	**I_{CM} max 25 mA**
AF181	**Germanium P-N-P alloy-diffused transistor for television video if amp. with forward A.G.C.**	**($I_E = 0$) —30V**	**I_{CM} max 20 mA**

Ptot max	*Current amplification factor (common emitter)*	
	Small signal	*Large signal*
(Tamb = 30°C) 60mW		
(Tamb = 30°C) 60mW		
(Tamb ≤ 45°C) 75mW	> 20	
(Tamb = 25°C) 140mW		
(Tamb = 25°C) 156mW		Max. unilateralised gain. Typ. 25dB
(Tamb = 25°C) 156mW		Max. unilateralised gain. Typ. 35dB

BA115	Gold bonded silicon diode for use as a television video noise limiter	Max.rev. volts 150V	Max. forward current 50 mA
BA144	Gold bonded silicon diode intended for use in television flywheel synchronising circuits	V_{RM} max 50V	I_{FM} max 50 mA IF (AV) max 2 mA
BA148	A fast general purpose diode	V_{RRM} max350v V_{RWM} max300v V_F max at I_F of 2A=1·5 V	I_F (AV) max averaged over 20 ms period 0·3A I_{FRM} 2A I_R max 200μA
BC107	Silicon N-P-N epitaxial planar transistor for audio driver stages and television signal processing circuits	(I_B=0) +50V	I_{CM} max 100 mA
BC108	Silicon N-P-N epitaxial planar transistor, for af pre-amp. and driver stages in amps.	(I_B=0) +30V	I_{CM} max 100 mA
BC109	Silicon N-P-N epitaxial planar transistor for low noise input stages	(I_B=0) +30V	I_{CM} max 100 mA

Average forward current 2mA	Tamb max 70°C	
50mW		
Tj max 125°C		
(Tamb ≦°25°C) 300mW	(V_{CE} = +5V I_C = 2 mA) 125 to 500	
(Tamb ≦ 25°C) 300mW	(V_{CB} = +5V I_C = 10 mA) 125 to 300	
(Tamb ≦ 25°C) 300mW	(V_{CE} = +5V I_C = 2 mA) 240 to 900	

Type	*Description*	V_{CB} *max*	I_C (AV) *max*
BC187	P-N-P silicon planar epitaxial transistor for use as sync separators and in a.g.c. and oscillator circuits for line and field deflection; also in driver stages of audio amplifiers	V_{CBO} max 30V	I_{CM} max 200 mA
BD121	Silicon n-p-n planar expitaxial power transistor intended for general audio applications	V_{CBO} max 60V	I_{CM}max 5A
BD123	Silicon n-p-n planar epitaxial power transformer intended for general audio applications	V_{CBO} max 90V	I_{CM} max 5A
BD124	Silicon n-p-n planar epitaxial power transistor intended for television field time-base output stages and general purpose medium power applications	V_{CBO} max (I_CF1 mA 70V	I_{CM} max 4A
BF115	Silicon n-p-n planar epitaxial transistor for am and fm applications	V_{CBO} max 50V	I_{CM} max 30 mA
BF167	Silicon N-P-N planar transistor for control stage of video if amplifiers	+40V V_{CE} max +30V	25 mA
BF173	Silicon N-P-N planar epitaxial transistor for output stages of television video if amplifiers	+40V V_{CE} max +25V	25 mA

Ptot max	Current amplification factor (common emitter)	
	Small signal	Large signal
Tamb 25°C 300mW	I_C = 2 mA 100 to 500 I_C = 50 mA 65 to 325	f_T. typ. at I_C = 50 mA 191 MHz
Tamb = 25°C 45W		I_C = 1·0A 55
Tamb = 25°C 45W		I_C = 1·0A 55
Tamb ≤ 60°C 15W		min(I_C = 0·5A) 35
Tamb ≤ 45°C 145mW		
(Tamb ≤ 45°C) 130mW		Max. unilateralised gain. Typ. 42dB
(Tamb ≤ 45°C) 200mW		Max. Unilateralised gain. Typ. 42dB

BF178	N-P-N Silicon planar transistor primarily intended for use in video output stages of monchrome television receivers	V_{CBO} max 145V	50 mA
BF184	N-P-N silicon planar epitaxial transistor recommended for use in the if amplifier of car radios and am/fm receivers; also for use in sound if stages of television receivers	V_{CBO} max 30V	30 mA
BF185	N-P-N silicon planar epitaxial low-noise transistor for use in the input stage of car radios and input and mixer/oscillator stages of am/fm receivers	V_{CBO} max 30V	30 mA
BF195	N-P-N low noise transistor in epoxy resin encapsulation with 3 rigid self-locking strips for insertion into printed circuit boards using standard grids. For use in input stages of am/fm receivers. Also mixer and i.f. stages of a.m. battery operated receivers	V_{CBO} max 30V	30 mA
BF200	V.H.F. Silicon planar N-P-N transistor with forward gain control characteristics intended for use in the r.f. amplifier stage of television v.h.f. tuners TO72 construction with the shield connected to envelope	V_{CBO} max 30V	20 mA

Tmb ≤ 65°C in free air 0·6W Tamb ≤ 100°C 1·7W		I_C = 30 mA V_{CE} = 20V = 20
Tamb ≤ 45°C 145 mW	I_C = 1mA V_{CE} = 10V 75 to 750	
Tamb ≤ 45°C 145 mW		I_C = 1 mA V_{CE} = 10V 34 to 140
Tamb ≤ 45°C 220mW	I_C = 1 mA V_{CE} = 10V 67	
Tamb 25°C 150mW	Stage power gain typical: (f = 200 MHz) 13dB	

Type	*Description*	VCB *max*	Ic (AV) *max*
BY126	**Silicon rectifier diode double-diffused junction rectifier with repetitive peak increase voltage of 650V. For use as a mains rectifier in television receivers. Plastic encapsulation**	**VRRM 650V Crest working reverse voltage 450V**	**IF (AV) 1A**
BY127	**Silicon rectifier diode. Double-diffused junction rectifier diode with repetitive peak reverse voltage of 1250V. For use as a mains rectifier in television receivers. Plastic encapsul-ation**	**VRRM 1250V Crest working reverse voltage 800V**	**IF (AV) 1A**
BY164	**Silicon bridge rectifier consisting of four double diffused junction diodes**	**Max. ac input V = 42V r.m.s.**	**VRRM = 120V**
BYX10	**Silicon rectifier diode. Double-diffused junction diode for low current rectifier applications Plastic encapsulation**	**VRWM max 800V VRRM max 1·6KV**	**IF (AV) max 200 mA**
OA91	**Germanium point-contact diode for detector in am receivers and general purposes**	**Max. rev. V 115 av 90V**	**Max. forward I peak 150 mA**

Ptot max	Current amplification factor (common emitter)	
	Small signal	Large signal
	Tj max 150°C	
	Tj max 150°C	
IFRM = 5A	Output voltage (capacitative load = 60V	Output voltage resistive/inductive load = 38V

OC26	Germanium P-N-P alloy-junction transistor for output stages car radio receivers	—32V	Icм max 3·5A
OC45	Germanium P-N-P alloy-junction transistor for rf and if amplifier stages	—15V	Icм max 10 mA
OC71	Germanium P-N-P alloy-junction transistor for low consumption audio amplifier	—30V	10 mA
OC75	Germanium P-N-P alloy-junction transistor for use in high gain amplifiers	—30V	10 mA

(Tamb = 75°C) 12·5W		20 to 60
(Tamb = 45°C) 43 mW	(V_{CE} = —6V I_E = 1mA) Typ. 50	
(Tamb = 45°C) 75 mW	(V_{CE} = —2V I_C = 1 mA) 41	
(Tamb = 45°C) 75 mW	90	

Semi-conductor replacement hints

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The following points are intended as a guide to some of the problems which may be encountered in radio and audio equipment.

1. *Polarity* P-N-P transistors are more common than n-p-n but it is essential that the correct polarity transistor is used. The collector terminal of p-n-p transistors will be negative with respect to the emitter, and the collector terminal of n-p-n transistors will be positive with respect to the emitter.

2. *Lead Lengths* The leads of all replacement components should be the same length as those of the original devices. If there is a screen lead on the Mullard replacement it should be connected to chassis if possible.

3. *Audio-frequency stages in portables* Arrangements with either output and driver transformers, or a driver transformer only, normally use p-n-p transistors, but if one n-p-n is present every transistor in the arrangement is probably n-p-n. Complementary push-pull arrangements (recognised by the absence of any transformers) usually have at least one n-p-n transistor and frequently more. These can be difficult to service, and it is usually necessary to trace out the circuit if no diagram is available.

4. *A.F. driver transistor* The replacement should be selected with care in circuits where the battery voltage is greater than 12V. The collector voltage rating should be twice the battery voltage, when a driver transformer is used.

5. *A.F. output transistors* If an output transistor has failed, and the cause appears to be over-heating, the Mullard replacement may also be in danger of failing. If there is room, cooling clips should be fitted to the output transistors, or the area of the heat-sink should be enlarged if one already exists. Otherwise the value of the emitter resistor can be increased, or thermistors can be fitted across the base bias resistors.

6. *Car radio output stages* Arrangements with no driver transformer may use a number of circuit configurations, and the pre-amplifier and driver transistors can be p-n-p or n-p-n. A Mullard AD149 should be used as a p-n-p output transistor replacement in all car radio circuits.

7. *A.M. I.F. stages* When transistors in i.f. stages are replaced, a type should be chosen which has a similar value of feedback capacitance. Unfortunately these figures for other manufacturers' types have not always been available. In general, an OC45 is a suitable p-n-p type when neutralising components are used, and an AF117 (also p-n-p) should be used when there is no neutralisation. If there is instability after the replacement has been fitted satisfactory operation may be obtained by making some circuit modifications. For example, if there are neutralising components the value of the neutralising capacitor should be altered. If there is no neutralisation, and if the transformer is single-tuned and of the correct phasing, instability may be removed by inserting a neutralising capacitor (value 1 to 10pF). Another method of making the stage stable is to insert a damping resistor across the primary of the i.f. transformer in the collector circuit.

8. *A.M. oscillator and mixer stages* An AF117 is a suitable p-n-p replacement. If the circuit does not oscillate after the replacement has been fitted, the emitter current should be increased (but not over 3mA). If there is squegging the value of the emitter decoupling capacitor should be reduced, and if this is unsuccessful a damping resistor should be connected across the oscillator tuned circuit.

9. *F.M. I.F. stages* A Mullard AF116 (p-n-p) should be used. If instability occurs the value of the neutralising capacitor should be altered if one is present. Otherwise the emitter current should be reduced (but not to less than half its value) by increasing the value of the upper base bias resistor. A damping resistor connected across the i.f. coil in the collector circuit may cure instability if other methods have failed.

10. *F.M. oscillators and mixers* **Mullard AF114, AF178 (both p-n-p) should be used. It is important to ensure that the lead lengths of the replacements are the same as those of the original devices. Instability can sometimes be cured by adjusting the value of the emitter current (by altering the value of the upper base bias resistor). It may be necessary to alter the value of the emitter feedback capacitor in oscillators.**

11. *F.M. R.F. amplifiers* **A Mullard AF114 or AF178 (both p-n-p) should be used as a replacement. If there is instability the emitter current should be reduced by increasing the value of the upper base bias resistor across the coil in the collector tuned circuit.**

Receiving valves

TYPE NOMENCLATURE SYSTEM

All new Mullard valves are registered with Pro-Electron and have type numbers according to the following code, based on the Pro-Electron type nomenclature system for receiving and amplifying valves.

The type number consists of two or more letters followed by a group of three figures (two figures in earlier types.)

The first letter indicates the heater or filament voltage or current:

D	0·5 to 1·5V filament
E	6·3V heater
G	5·0V heater
P	300mA heater
U	100mA heater

Letters A (4·0V), C (200mA) and K (2·0V) have also been used.

The second and subsequent letters indicate the general class of valve:

A	single diode
B	double diode
C	triode
D	power output triode
E	tetrode
F	pentode
L	power output tetrode or pentode
H	hexode or heptode (hexode type)
K	octode or heptode (octode type)
M	tuning indicator
Y	half-wave rectifier
Z	full-wave rectifier

Two or three of these letters may be combined together, e.g. BC—double-diode triode.

The first figure of the serial number indicates the type of base:

2	B10B (10-pin) base (previously used for B8G base)
3	Octal base
4	B8A base
5	B9D (magnoval) base (previously used for miscellaneous bases)
8	B9A (noval) base
9	B7G base

The remaining figure or figures make up the serial number indicating a particular design or development.

Examples

PCF806	Triode pentode with B9A base for use in 300mA series heater chain
EC90	Triode with B7G base and 6·3V heater.

List of earlier types and types not in common use

AZ31	EBC41	EF55	EZ35
DAF91	EC90	EF92	EZ40
DAF96	EC91	EF95	EZ41
DCC90	ECC32	EL33	GZ30
DF91	ECC33	EL36	GZ32
DF96	ECC35	EL37	GZ33
DK91	ECC84	EL38	GZ37
DK92	ECC91	EL41	PL820
DK96	ECF80	EL42	TY86F
DL92	ECH35	EL85	UAF42
DL94	ECH42	EL91	UBC41
DL96	EF37A	EL95	UCH42
DM70	EF39	EL820	UF41
DM71	EF40	EL821	UY41
EAF42	EF41	EM34	20P4/CL30
EBC33	EF50	EM81	

MULLARD VALVE TYPES

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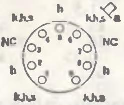
DY86/87 E.H.T. Half-Wave Rectifier



B9A

Vh	1·4	V
Ih	550	mA
Pulsed input		
P.I.V. max.	22	kV
ia(pk) max.	40	mA
Iout max.	500	μA
C max.	2000	pF

Pins 3 and 7 may only be connected to points in the heater circuit and must not be earthed.

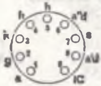
EB91 Double Diode (seperate cathodes)

B7G

Vh	6·3	A
Ih	300	mA
*P.I.V. max.	420	V
*Ia max.	9·0	mV
*ia(pk) max.	54	mA
*vh—k(pk) max.	330	V

*Each section

EBC81 Double Diode Triode



B9A

Vh	6·3	V
Ih	230	mA
Va	250	V
Vg	—3·0	V
Ia	1·0	mA
gm	1·2	mA/V
μ	70	

EBF89 Double Diode Variable-MU R.F. Pentode

B9A

Vh	6·3		V
Ih	300		mA
Va	250	250	V
Vg3	0	0	V
Vg2	80	100	V
Vg1	—1·0	—2·0	V
Ia	9·0	9·0	mA
Ig2	2·7	2·7	mA
gm	4·5	3·8	mA/V
ra	0·9	1·0	MΩ
μg1—g2	20	20	

EC86 U.H.F. Frame-Grid Mixer/Oscillator Triode

B9A

Vh	6·3	V
Ih	200	mA
Va	175	V
Vg	—1·5	V
Ia	12	mA
gm	14	mA/V
ra	4·85	kΩ
μ	68	

ECC81 R.F. Double Triode (separate cathodes)

B9A

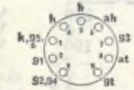
	Series	Parallel	
Vh	12·6	6·3	V
Ih	150	300	mA
Characteristics (each section)			
Va	200	250	V
Vg	—1·0	—2·0	V
Ia	11·5	10	mA
gm	6·7	5·5	mA/V
μ	70	60	

ECC85 R.F. Double Triode (separate cathodes)

B9A

Vh	6·3	V
Ih	435	mA
Characteristics (each section)		
Va	250	V
Vg	—2·7	V
Ia	10	mA
gm	6·1	mA/V
μ	55	

ECH81 Triode Heptode Frequency Changer



B9A

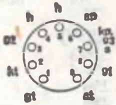
Vh	6·3	V
Ih	300	mA
Vah = Vb	250	V
Rg2 + g4	22	kΩ
Rg3 + gt	47	kΩ
Rk	140	Ω
Iah	3·25	mA
Ig2 + g4	6·7	mA
Ig3 + gt	200	μA
gc	775	μA/V
Vat	100	V
Iat	4·5	mA

ECL82 Triode Output Pentode (pa max. =5·4W)

B9A

	Triode	Pentode	
Vh		6·3	V
Ih		780	mA
Va	100	250	V
Vg2	—	250	V
Ia	3·5	28	mA
Ig2	—	5·7	mA
Vg1	0	−22·5	V
gm	2·5	5·0	mA/V
Ra	—	9·0	kΩ
Pout	—	3·4	W

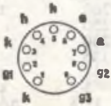
ECL86 Triode Output Pentode (pa max. =9W)



B9A

	Triode	Pentode	
Vh		6·3	V
Ih		660	mA
Va	250	250	V
Vg2	—	250	V
Ia	1·2	36	mA
Ig2	—	6·0	mA
Vg1	−1·9	−7·0	V
gm	1·6	10	mA/V
ra	62	48	kΩ
Ra	—	7·0	kΩ
Pout	—	4·0	W

EF80 High Slope R.F. Pentode



B9A

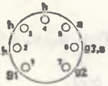
Vh	6·3	V
Ih	300	mA
Va	170	V
Vg2	170	V
Vg3	0	V
Rk	160	Ω
Ia	10	mA
Ig2	2·5	mA
gm	7·4	mA/V
μg1–g2	50	

EF86 Low Noise A.F. Voltage Amplifying Pentode


B9A

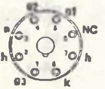
Vh	6·3	V
Ih	200	mA
Va	250	V
Vg3	0	V
Vg2	140	V
Vg1	−2·2	V
Ia	3·0	mA
Ig2	600	μA
gm	2·2	mA/V
μg1–g2	38	

EF91 High Slope R.F. Pentode


B7G

Vh	6·3	V
Ih	300	mA
Va	250	V
Vg2	250	V
Vg3	0	V
Rk	160	Ω
Ia	10	mA
Ig2	2·6	mA
gm	7·6	mA/V
μg1–g2	70	

EL34 Output Pentode (pa max. = 25W)


Octal

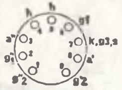
Vh	6·3	V
Ih	1·5	A
Va	250	V
Vg2	250	V
Vg3	0	V
Rk	106	Ω
Ia	100	mA
Ig2	15	mA
gm	11	mA/V
Ra	2·0	kΩ
Pout	11	W

EL84 Output Pentode (pa max. = 12W)


B9A

Vh	6·3	V
Ih	760	mA
Va	250	V
Vg2	250	V
Rk	135	Ω
Ia	48	mA
Ig2	5·5	mA
gm	11·3	mA/V
Ra	4·5	kΩ
Pout	5·7	W

ELL 80 Double Output Pentode (pa max. = 2 × 6W)

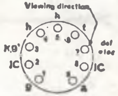


B9A

Vh	6·3	V
Ih	550	mA
Characteristics (each section)		
Va	250	V
Vg2	250	V
*Rk	160	Ω
Ia	24	mA
Ig2	4·5	mA
gm	6·5	mA/V
Ra	10	kΩ
Pout	3·0	W

*Common to both sections

EM84 Voltage Indicator



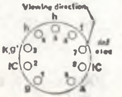
B9A

Vh	6·3		V
Ih	210		mA
Vb	250		V
Vt	250		V
Ra	470		kΩ
Rg–k	3		MΩ
Vg	0	–22	V
Ia	450	60	μA
It	1·0	1·8	mA
*L	21	0	mm

Deflection electrode connected to anode.

*Length of column

EM87 Voltage Indicator



B9A

Vh	6·3			V
Ih	300			mA
Vb	250			V
Vt	250			V
Ra	100			kΩ
Rg–k	3·0			MΩ
Vg	0	–10	–15	V
Ia	2·0	0·5	0·2	mA
t	1·0	1·8	2·0	mA
*L	21	0	–1·5	mm

Deflection electrode connected to anode.

*Length of column. A negative value of L indicates overlapping

EY86/87 High Voltage Half-Wave Rectifier

B9A

Vh	6·3	V
Ih	90	mA
Pulsed input		
P.I.V. max.	22	kV
Iout	800	μA
ia(pk) max.	40	mA
C max.	2000	pF

†Pins 1, 4, 6 and 9 may be used for fitting an anti-corona shield
*Pins 3 and 7 may only be connected to points in the heater circuit and must not be earthed

EZ81 Full-Wave Rectifier

B9A

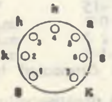
Vh	6·3	V
Ih	1·0	A
Vin (r.m.s.)	2x350	V
Iout max.	160	mA
C max.	50	μF
Rlim min. (per_anode)	230	Ω

PC88 U.H.F. Frame-Grid Grounded Grid Amplifier Triode

B9A

Ih	300	mA
Vh	3·8	V
Va	160	V
Vg1	−1·25	V
Ia	12·5	mA
gm	13·5	mA/V
ra	4·8	kΩ
μ	65	

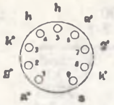
PC900 R. F. Triode



B7G

Ih	300	mA
Vh	4·0	V
Va	135	V
Vg	−1·0	V
Ia	11·5	mA
gm	14·5	mA/V
μ	72	
ra	5·0	kΩ

PCC88 Frame-Grid Double Triode



B9A

Ih	300	mA
Vh	7·0	V
Characteristics (each section)		
Va	90	V
Vg	−1·3	V
Ia	15	mA
gm	12·5	mA/V
μ	33	

PCC89 Variable-MU Frame-Grid Double Triode

B9A

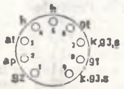
Ih	300	mA
Vh	7·5	V
Characteristcs (each section)		
Va	90	V
Ia	15	mA
Vg	−1·3	V
gm	12·5	mA/V
μ	33	

PCF80 Triode Pentode (separate cathodes)

B9A

	Triode	Pentode	
Ih	300		mA
Vh	9·0		V
Va	100	170	V
Vg2	—	170	V
Vg1	−2·0	−2·0	V
Ia	14	10	mA
Ig2	—	2·8	mA
gm	5·0	6·2	mA/V
μ	20		

PCF84 Triode Pentode



B9A

	Triode	Pentode	
Ih	300		mA
Vh	9·0		V
Va	100	170	V
Vg2	—	170	V
Vg1	−2·0	−2·0	V
Ia	14	12	mA
Ig2	—	3·0	mA
gm	5·0	7·5	mA/V
ra	4·0	400	kΩ

PCF801 Triode Frame-Grid Variable-Mu Pentode

B9A

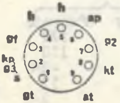
	Triode	Pentode	
Ih	300		mA
Vh	8·5		V
Va	100	170	V
Vg2	—	120	V
Vg1	−3·0	−1·4	V
Ia	15	10	mA
Ig2	—	3·0	mA
gm	9·0	11	mA/V
μ	20	—	
ra	2·2	>350	kΩ

PCF802 Triode Pentode

B9A

	Triode	Pentode	
Ih	300		mA
Vh	9·0		V
Va	200	100	V
Vg2	—	100	V
Vg1	−2·0	−1·0	V
Ia	3·5	6·0	mA
Ig2	—	1·7	mA
gm	3·5	5·5	mA/V
μ	70	—	
ra	20	400	kΩ

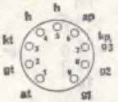
PCL82 Triode Output Pentode (pa max. = 7W)



B9A

	Triode	Pentode	
Ih	300		mA
Vh	16		V
Va	100	170	V
Vg2	—	170	V
Vg1	0	−11·5	V
Ia	3·5	41	mA
Ig2	—	9·0	mA
gm	2·2	7·5	mA/V
μ	70	—	
Ra	—	3·9	kΩ
Pout	—	3·3	W

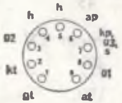
PCL83 Triode Output Pentode (pa max. = 5·4W)



B9A

	Triode	Pentode	
Ih	300		mA
Vh	12·6		V
Va	250	170	V
Vg2	—	170	V
Vg1	−8·5	−9·5	V
Ia	10·5	30	mA
Ig2	—	5·0	mA
gm	2·2	5·5	mA/V
μ	17	—	
Ra	—	5·5	kΩ
Pout	—	2·2	W

PCL86 Triode Output Pentode (pa max. (pentode) = 9W)



B9A

	Triode	Pentode	
Ih	300		mA
Vh	13·3		V
Va	230	230	V
Vg2	—	230	V
Vg1	−1·7	−5·7	V
Ia	1·2	39	mA
Ig2	—	6·5	mA
gm	1·6	10·5	mA/V
ra	—	45	kΩ
μg1–g2	—	21	

PL82 Output Pentode (pa max. = 9W)

B9A

Ih	300		mA
Vh	16·5		V
Va	170	200	V
Vg2	170	200	V
Rk	165	270	Ω
Ia	53	45	mA
Ig2	10	8·5	mA
gm	9·0	7·6	mA/V
Ra	3·0	4·0	kΩ
Pout	4·0	4·2	W

PL508 Field Output Pentode for Colour TV

B9D

Ih	300	mA
Vh	17	V
Va	190	V
Vg2	190	V
Ia	60	mA
Ig2	4·5	mA
Vg1	−17	V
gm	9·0	mA/V
μg1–g2	7·0	
ra	10	kΩ

PL509 Line Output Pentode for Colour TV (pa max. = 30W)

B9D

Ih	300	mA
Vh	40	V
Va	160	V
Vg3	0	V
Vg2	160	V
Vg1	0	V
Ia	1·4	A
Ig2	45	mA

PL802 Video Output Pentode for Colour TV

B9A

Ih	300	mA
Vh	16	V
Va	170	V
Vg3	0	V
Vg2	170	V
Vg1	−0·9	V
Ia	30	mA
Ig2	6·5	mA
gm	40	mA/V
ra	45	kΩ
μg1–g2	70	

PY82 Half-Wave Rectifier

B9A

Ih	300	mA
Vh	19	V
P.I.V. max.	700	V
Vin(r.m.s.) max.	250	V
Iout max.	180	mA
C max.	60	μF
Rlim min.	45	Ω

PY88 Booster Diode

B9A

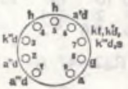
Ih	300	mA
Vh	30	V
P.I.V. max.	6·6	kV
Ia(av) max.	220	mA
vh–k (pk) max. (cathode positive)	6·6	kV

U301/CY30 Efficiency Diode

Octal

Ih	200	mA
Vh	28	V
P.I.V. max.	4·5	kV
Ia max.	150	mA
V(h–k) max.	900	V

UABC80 Triple Diode Triode (one diode having a separate cathode)



B9A

Ih	100		mA
Vh	28		V
Va	170	200	V
Vg	−1·8	−2·3	V
Ia	1·0	1·0	mA
gm	1·45	1·4	mA/V
μ	70	70	

UBC81 Double Diode Triode

B9A

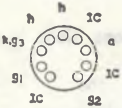
Ih	100		mA
Vh	14		V
Va	100	170	V
Vg	−1·0	−1·6	V
Ia	0·8	1·5	mA
gm	1·4	1·65	mA/V
μ	70	70	
ra	50	42	kΩ

UF89 Variable MU R.F. Pentode

B9A

Ih	100		mA
Vh	12·6		V
Va	170	200	V
Vg3	0	0	V
Rg2	15	24	kΩ
Rk	130	130	Ω
Ia	11	11·1	mA
Ig2	3·9	3·8	mA
gm	3·8	3·85	mA/V

UL84 Output Pentode (pa max. = 12W)

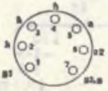


B9A

Ih	100			mA
Vh	45			V
Va	100	170	200	V
Vg2	100	170	*	V
Rk	150	170	270	Ω
Ia	43	70	60	mA
Ig2	3·0	5·0	4·1	mA
gm	9·0	10	8·8	mA/V
Ra	2·4	2·4	2·4	kΩ
Pout	1·9	5·6	5·2	W

*Vg2(b) = 200V, Rg2 = 470 Ω

6CB6 R.F. Pentode



B7G

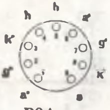
Vh	6·3	V
Ih	300	mA
Va	200	V
Vg3	0	V
Vg2	150	V
Vg1	−2·2	V
Ia	9·5	mA
Ig2	2·8	mA
ra	600	kΩ
gm	8·0	mA/V

6F23/EF812 High Slope R.F. Pentode

B9A

Vh	6·3	V
Ih	300	mA
Va	170	V
Vg2	170	V
Rk	150	Ω
Ia	10	mA
Ig2	2·6	mA
gm	9·2	mA/V
μg1–g2	60	

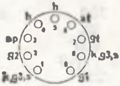
6/30L2/ECC804 Double Triode (separate cathodes)



B9A

Vh	6·3	V
Ih	300	mA
Characteristics (each section)		
Va	200	V
Vg	−7·7	V
Ia	10	mA
gm	3·4	mA/V
μ	18	

30C15/PCF800 V.H.F. Triode Pentode



B9A

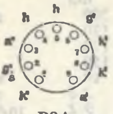
Ih		300	mA
Vh		9·0	V
	Triode	Pentode	
Va	100	170	V
Vg2	—	170	V
Ia	15	10	mA
gm	6·0	9·0	mA/V
μ	20	—	

30F5/PF818 H.F. Screened Pentode (pa max. = 3W)

B9A

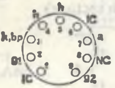
Ih	300	mA
Vh	7·3	V
Va	170	V
Vg3	0	V
Vg2	170	V
Vg1	−1·9	V
Ia	10	mA
Ig2	2·6	mA
Rk	150	Ω
gm	8·8	mA/V

30L15/PCC805 R.F. Cascode Double Triode


B9A

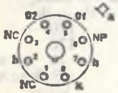
Ih	300	mA
Vh	7·0	V
Characteristics (each section)		
Va	90	V
Vg	−1·2	V
Ia	15	mA
gm	9·0	mA/V
μ	27	

30P12/PL801 Beam Tetrode (A.F. or field output, pa max. = 6W)


B9A

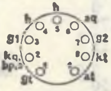
Ih	300	mA
Vh	12·6	V
Va	170	V
Vg2	180	V
Vg1	−10·3	V
Ia	31	mA
Ig2	7·3	mA
Ra	5·0	kΩ
Pout	2·25	W

30P19/PL302 Line Output Beam Tetrode (pa max. = 10W)


Octal

Ih	300	mA
Vh	25	V
Va max.	400	V
va(pk) max.	7·0	kV
Vg2 max.	250	V
vg2(pk) max.	2·0	kV
Ik max.	200	mA
Rg1–k max.	1·0	MΩ
Vh–k(r.m.s.) max.	200	V

30PL14/PCL88 Triode Output Beam Tetrode


B9A

Ih	300		mA
Vh	16		V
	Triode	Tetrode	
Va	100	170	V
Vg2	—	170	V
Ia	10	50	mA
gm	4·3	7·3 m	A/V
μ	18	—	

Index

Admittance nomograph, 32
Aerials, 116
Af amplifiers, 57
Alternating current, 30
Amateur prefixes, 79
Amplifier by G.E.C., 130
Amplifier, typical transistor, 131
Amplitude modulation, 84
Anode modulation, 85
Attenuator
- between matched impedances, 44
- design, 41
- tables, 42, 43

Audio frequency, 54
Automatic gain control, 60, 90

Balanced modulator, 86
Bandpass filter, 47
Bands III, IV & V, 119
Broadside array aerial, 123

Capacitance, 27
- table of values, 28

Capacitative reactance, Xc, 30
Capture area, 123
Carbon resistors, 22
Carrier wave production, 82
Cathode decoupling, 59
C.C.I.R. equalisation, 72
Circuit
- formulae, 21
- for velocity sensitive pick-up, 68

Clapp oscillator, 83
Colour
- locking, 110
- receiver, 112
- television, 104

Colpitts oscillator, 83
Compatability
- of PAL system, 115
- of stereo transmission, 66

Cone aerial, 120
Constant resistance networks, 44
Copper wire tables, 18
Coupling of a f amplifiers, 57
Critical frequency, 74
Cutting a stereo disc, 69

Decibel calculations, 38
Decoupling of af amplifiers, 57
Delta/star transform, 21
Demodulator, balanced, 86
Diatonic scale, 37
Directors & reflectors, for channels 118-123

Electro-magnetic waves, 74
Equal-tempered scale, 37

Feeders & aerials, 116
Filters, 45
fm detectors, 91
fm/vhf band, 119
Form factor, 30
Foster-Seeley discriminator, 92
Four-tube colour cameras 104, 112
Frequency
- bands, 74
- modulation, 86
- tables, channels 1-5, vhf/fm & bands III, IV & V 119

Gain at Lf & hf, 57
Gamma correction, 115
Grid modulation, 85

Half-wave dipole aerial, 123
Hartley oscillator, 83
Heising modulation, 85

Image operation of networks, 48
Image orthicon television camera, 98
Impedance, 30
- nomograph, 32

Induced voltage expressions, 23
Inductance, 23
- calculation of, 24

Inductive reactance, X_L, 24
Infinite impedance detector, 89
Intensity, 36
Intensity vs loudness, 40
International amateur prefixes, 79
Invertor, 53
Ionosphere, 74
Isotropic aerial, 123

Klystron, 94

Leaky-grid detector, 89
Lens aerial, 123

Loudness, 36
Loudspeaker operation, 64
Magnetic recording, 70
Maser, 96
Meissner oscillator, 83
Microphones, 54
Microwaves, 94
Musical scales, 36
Mutual inductance, 23
Offset waveguide, 123
Optimum wavebands, for short wave reception, 77
Oscillator circuits, 83
PAL
  decoder, block schematic, 111
  video stages, block schematic, 113
Parabolic aerial, 122
Pianoforte frequencies, 37
Pitch, 36
Power
  factor, 30
  supplies, 49
  supply circuits, 52
Pre-amplifier, transistor, 130
Preferred values of resistors, 22
Prefixes, international amateur, 79
Pressure, sound, 36
Propagation, of radio wave, 74
Q, calculation of, 35
Q factor, 31
Quarter wave rod aerial for vhf, 121
Radio Society of Great Britain, 78
Ratio detector, 93
Reactance, 30
  chart, af, 33
  chart, rf, 34
RC coupling response curves, 58
Receiving aerials for uhf, 124
Reception, 88
Reflector type aerials, 122
Resistance
  circuit formulae, 21
  ohm's law, 20
  nomograph, 20
  of materials, 21
  wire table, 19
  colour code, 22
  preferred values, 22
Resonance, 31
RCA shadow mask colour tube, 114
RM S value, 30
Screen decoupling, 57
Semi-conductor
  data, 132
  devices, 127
  replacement hints, 140
Series modulation, 85
Shadow-mask colour tube, RCA, 114
Short wave
  broacasting, 75
  reception, 77
Sinpo code, 78
Skin effect, 22
Sound, 36
  intervals, 37
  pressure, 36
Stage gain, 57
Star/delta transform, 21
Stereophonic brodcasting, 66
  pickups, 69
Stereophony, 64
Stop filter, 47
Superhet, 88
Television, 97
  cameras, 98
  colour, 104
  receivers, 101
  test card, 102
  waveforms, 98, 99, 108
Timbre, of sound, 36
Time constant, 29
Tone control circuits, 61
Transformer
  coupling, 59
  design, 49
  equivalent circuit, 59
Transmission, 82
Transistor data, 128, 132
Transistorised amplifiers, 130, 131
Travelling-wave tube, 95
Ultra high frequencies, 124
  receiving aerials, 124
  television stations, 124
Valve type & bases, 143
Variable-mu valves, 60
Velocity sensitive pickup, 68
Vhf/fm band, 119
Vidicon television camera, 100
Wave guides, 94
Wire tables, 19