THE RADIO AND TELECOMMUNICATIONS ENGINEER'S DESIGN MANUAL

BY

R. E. BLAKEY, D.Sc.

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PREFACE

The Author believes a demand exists for a book dealing exclusively with the basic design principles of radio components and radio test gear. At the same time a highly mathematical and academic treatment of the subject would completely invalidate the purpose of this book, which the Author has tried to make concise in every respect and at the same time a manual of reference which enables a given component or piece of apparatus to be designed in the most efficient and easiest manner. In addition to basic calculations, many tables and practical examples are given, these being included for the express purpose of facilitating design. Many examples of commercial test gear are given, care having been taken to select examples of common interest and of well-known makes. If this book assists the working radio engineer in his job, the Author will feel recompensed for his labours.

R. E. BLAKEY

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THE RADIO AND TELECOMMUNICATIONS ENGINEER'S DESIGN MANUAL

SECTION I

VOLUME CONTROLS AND ATTENUATION NETWORKS

DECIBEL RATIOS

Logarithmic Volume Controls. The decibel is defined by the equation

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

when E_1 = volts across the input terminals
and E_2 = volts across the output terminals.

Thus in the case of a potentiometer, since the current is equal in each section, we may write

$$N \text{ (db.)} = 20 \log_{10} (R_1/R_2)$$

where R_1 = resistance across input terminals
and R_2 = resistance across output terminals.

TABLE I

Multiplying Constants	1 db. Steps
R_1	10 R_0
R_2	11·22 R_0
R_3	12·599 R_0
R_4	14·126 R_0
R_5	15·849 R_0
R_6	17·793 R_0
R_7	19·953 R_0
R_8	22·387 R_0
R_9	25·119 R_0
R_{10}	28·184 R_0

Tables I and II give the multiplying factors of R_0 for each element in steps of 1 db. and 10 db. respectively, whilst Tables III and IV give resistance values for commonly used values of volume controls. Figs. 1 and 2 show the theoretical diagram.

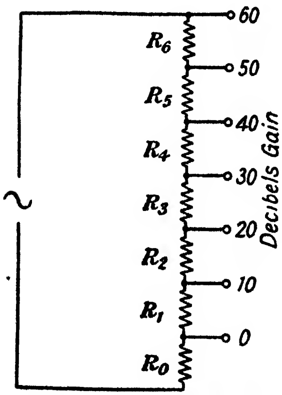


FIG. 1. TEN DECIBEL STEPS

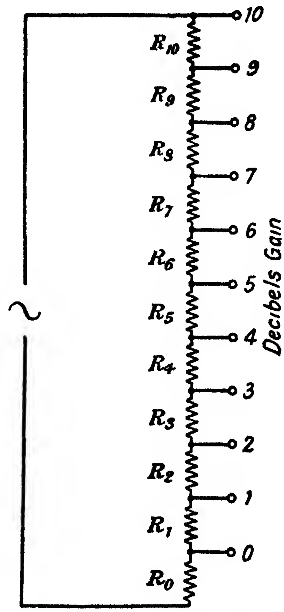


FIG. 2. ONE DECIBEL STEPS

TABLE II

Multiplying Constants	10 db. Steps	Voltage or Current Ratio	Power Ratio
R_1	2·1623 R_0		
R_2	6·8368 R_0 = 20 db.	10	10^2
R_3	21·623 R_0		
R_4	68·368 R_0 = 40 db.	10^2	10^4
R_5	216·23 R_0		
R_6	683·68 R_0 = 60 db.	10^3	10^6

TABLE III

RESISTANCE VALUES CORRESPONDING TO TABLE I AND FIG. 2

There is a Constant Error of 0·76 per cent in the totals corresponding to the nominal values, and is negative in sign

R	Nominal Value of Volume Control in Ohms								Db.
	5 000	10 000	25 000	50 000	100 000	250 000	500 000	1 MΩ	
R_0	28·05	56·10	140·25	280·50	561·00	1 402·50	2 805·00	5 610·00	0
R_1	280·50	561·00	1 402·50	2 805·00	5 610·00	14 025·00	28 050·00	56 100·00	1
R_2	314·72	629·44	1 573·60	3 147·21	6 294·42	15 736·05	31 472·10	62 944·20	2
R_3	353·48	706·96	1 767·42	3 534·84	7 069·69	17 674·23	35 348·47	70 696·94	3
R_4	396·23	792·46	1 981·17	3 962·34	7 924·68	19 811·71	39 623·42	79 246·84	4
R_5	445·56	891·12	2 227·82	4 455·64	8 911·28	22 278·22	44 556·44	89 112·88	5
R_6	499·09	998·18	2 495·46	4 990·93	9 981·86	24 954·65	49 909·31	99 818·62	6
R_7	559·67	1 119·34	2 798·35	5 596·71	11 193·43	27 983·58	55 967·16	111 934·32	7
R_8	627·95	1 255·90	3 139·77	6 279·54	12 559·08	31 397·71	62 795·42	125 590·84	8
R_9	704·58	1 409·17	3 522·93	7 045·87	14 091·75	35 229·39	70 458·78	140 917·56	9
R_{10}	789·76	1 579·52	2 948·80	7·897·61	15 795·22	39 488·05	78 976·11	157 952·22	10

TABLE IV

RESISTANCE VALUES CORRESPONDING TO TABLE II AND FIG. 1

There is a constant error of 0·001 per cent in the totals corresponding to the nominal values, and is negative in sign

R	Nominal Value of Volume Control in Ohms								Db.
	5 000	10 000	25 000	50 000	100 000	250 000	500 000	1 MΩ	
R_0	5·00	10·00	25·00	50·00	100·00	250·00	500·0	1 000··00	0
R_1	10·81	21·62	54·05	108·11	216·23	540·575	10 81·15	2 162 3	10
R_2	34·18	68·36	170·92	341·84	683·68	1 709·200	3 418·40	6 836·8	20
R_3	108·11	216·23	540·57	1 081·15	2 162·30	5 405·750	10 811·50	21 623·0	30
R_4	341·84	683·68	1 709·2	3 418·4	6 836·80	17 092·00	34 184·0	68 360·0	40
R_5	1 081·15	2 162·30	5 405·75	10 811·5	21 623·00	54 057·50	108 115·0	216 230·0	50
R_6	3 418·40	6 836·80	17 092·00	34 184·0	68 368·00	170 920·00	341 840·0	683 680·0	0

ATTENUATORS

Constant Impedance Attenuators. The attenuator sections shown in Fig. 3 (*a*) to (*e*) contain resistance elements whose values for any desired degree of attenuation can be easily calculated by reference to Table V. Multiply the desired characteristic resistance R_0 (resistance looking into either end of the section) by the factor in Column 1 to obtain the value of resistance R_1. Similarly multiply R_0 by the factor in Column 2 to obtain the value of R_2, by the factor in Column 3 for R_3 and by the factor in Column 4 for R_4. These values of R_1, R_2, R_3, and R_4 are the required resistance values to use wherever

TABLE V

Multiplying Constants for Constant Impedance Attenuators of Figs. 3–7

Attenuation Desired (Db.)	Multiplying Factors				Attenuation Desired (Db.)
	1	2	3	4	
0·25	0·01470	68·03	0·02955	33·85	0·25
0·50	0·02874	34·79	0·0576	17·361	0·50
1·0	0·0575	17·39	0·1153	8·669	1·0
2·0	0·1146	8·726	0·2323	4·305	2·0
3·0	0·1710	5·848	0·3524	2·838	3·0
4·0	0·2260	4·425	0·4776	2·094	4·0
5·0	0·2802	3·569	0·6080	1·645	5·0
6·0	0·3325	3·007	0·7469	1·339	6·0
7·0	0·3824	2·614	0·8961	1·116	7·0
8·0	0·4305	2·323	1·0575	0·9452	8·0
9·0	0·4760	2·101	1·2316	0·8117	9·0
10·0	0·5194	1·925	1·4229	0·7027	10·0
15·0	0·6980	1·432	2·720	0·3675	15·0
20·0	0·8183	1·222	4·95	0·2020	20·0
25·0	0·8940	1·119	8·876	0·1127	25·0
30·0	0·9389	1·065	15·8	0·06332	30·0
35·0	0·9651	1·036	28·131	0·03555	35·0
40·0	0·9804	1·020	50·00	0·0200	40·0

1 = tanh – db./(2 × 8·69). 2 = coth – db./(2 × 8·69).
3 = sinh – db./8·69. 4 = 1 ÷ factor 3.

TABLE VI

Decibel Ratios

Decibels	Current or Voltage Ratio	Power Ratio	Decibels	Current or Voltage Ratio	Power Ratio
0	1·00	1·00	20	10·0	100
1	1·12	1·26	22	12·6	160
2	1·26	1·59	24	16·0	250
3	1·41	2·0	26	20·0	400
4	1·59	2·5	28	25	630
5	1·78	3·2	30	32	1 000
6	2·00	4·0	32	40	1 590
7	2·24	5·0	34	50	2 510
8	2·51	6·3	36	63	3 980
9	2·82	7·9	38	79	6 310
10	3·16	10·0	40	100	10 000
12	4·0	16·0	45	178	31 600
14	5·0	25·0	50	316	100 000
16	6·3	40·0	60	1 000	1 000 000
18	7·9	63·0			

indicated in the diagrams of Fig. 3 (*a*) to (*e*) to obtain an attenuator section having the desired attenuation and characteristic resistance.

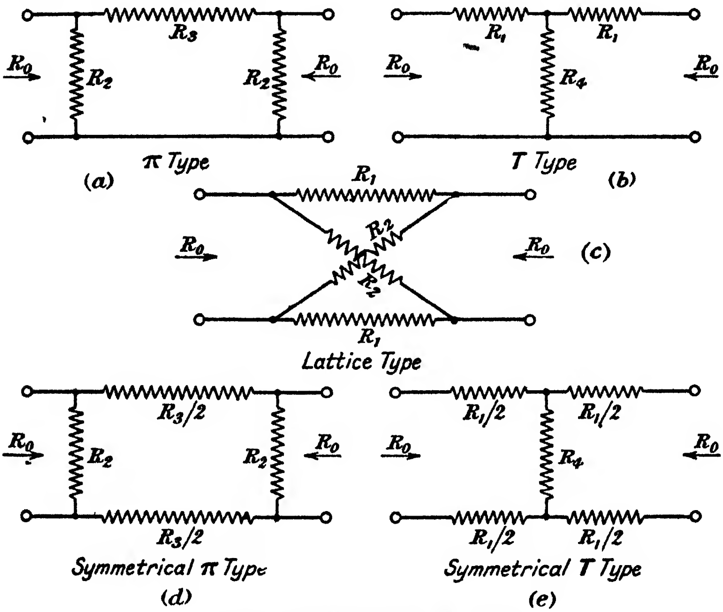


FIG. 3. ATTENUATOR SECTIONS

Radio-frequency Attenuators. These may be divided into two broad classes—

(*a*) For use with low power signal generators.

(*b*) For use with high power signal generators.

The "ladder" type of attenuator is most commonly used, therefore details are given of both low and high power types.

The basic equations are as follows—

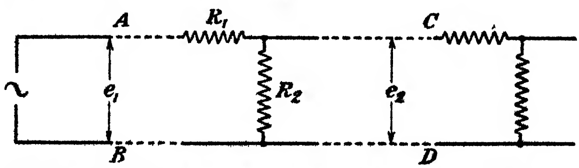


FIG. 4. "L" TYPE ATTENUATOR, FUNDAMENTAL DIAGRAM

Input resistance between AB, CD, and so on $= R$: then

$$e_1/e_2 = \text{attenuation ratio}$$

$$= \frac{RR_2}{RR_1 + RR_2 + RR_2} \quad . \quad . \quad . \quad (1)$$

and

$$R_2 = \frac{R^2 - RR_1}{R_1} \quad . \quad . \quad . \quad . \quad . \quad (2)$$

When Eq. (2) is fulfilled

$$e_1/e_2 = (R - R_1)/R = A \text{ (attenuation)} \quad . \quad (3)$$

$$\therefore \quad R_1 = R(1 - A) \quad . \quad . \quad . \quad . \quad . \quad (4)$$

Figs. 5, 6 and 7 show typical designs of r.f. attenuators for high, medium, and low power oscillators, and of these types the one of Fig. 7 is most commonly used.

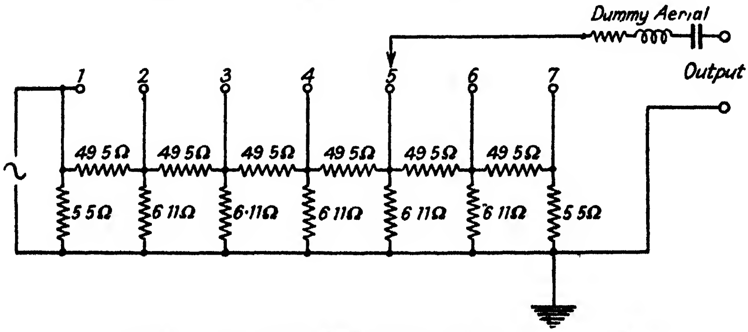


FIG. 5. "L" TYPE ATTENUATOR, FIXED INPUT TYPE

TABLE VII

ATTENUATION VALUES FOR NETWORK OF FIG. 5

Assuming 1 volt at Point 1, the power required from the oscillator is 200 mW. The effective resistance is 5 ohms.

Point	Output (μV.)	Attenuation
1	1 000 000	—
2	100 000	10^{-1}
3	10 000	10^{-2}
4	1 000	10^{-3}
5	100	10^{-4}
6	10	10^{-5}
7	1	10^{-6}

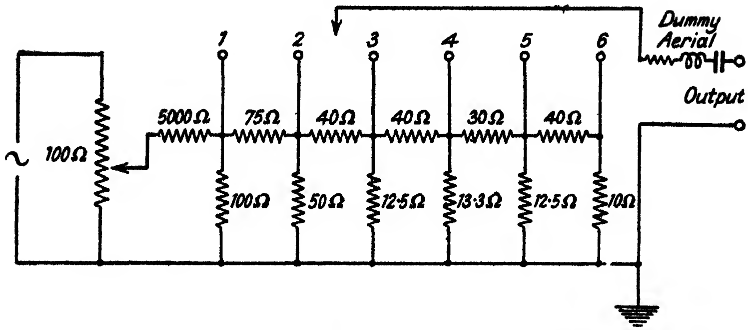


Fig. 6. "L" Type Attenuator, Variable Input Type

TABLE VIII

Attenuation Values of Network of Fig. 6

The input voltage for this attenuator requires to be 2 volts if the figures in the Table are to be fulfilled. The power required is 60 mW.

Point	Output (μV.)	Attenuation
1	10^5–10^6	*nil*–10
2	10^4–10^5	10 –10^2
3	10^3–10^4	10^2–10^3
4	10^2–10^3	10^3–10^4
5	10 –10^2	10^4–10^5
6	1 –10	10^5–10^6

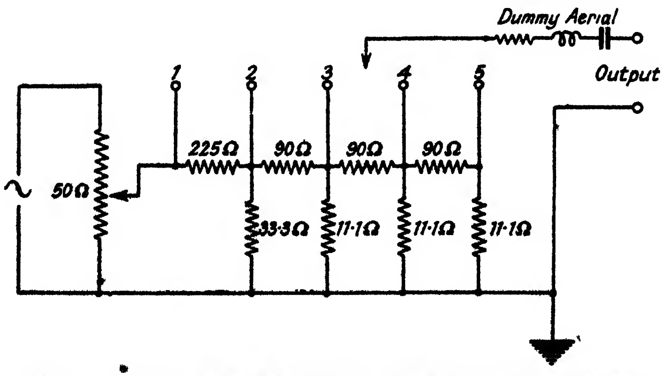


Fig. 7. "L" Type Attenuator of Economical Structure

TABLE IX

ATTENUATION VALUES OF NETWORK OF FIG. 7

Assuming an input voltage to the attenuator of 0·1 volt, the power required from the oscillator is 2·5 mW.

Point	Output (μV.)	Attenuation
1	10^4–10^5	*nil*–10
2	10^3–10^4	10 –10^2
3	10^2–10^3	10^2–10^3
4	10 –10^2	10^3–10^4
5	1 –10	10^4–10^5

SECTION II

INDUCTANCES

AIR- AND IRON-CORED

Air-cored Coils. The inductance of a single layer coil is given by the formula

$$L = 0{\cdot}00987dN^2(Kd/l)\ \mu\text{H.} \quad . \quad . \quad . \quad . \quad . \quad (5)$$

where d = diameter in cm.;
l = length in cm.;
N = total turns;
K = Nagaoka's constant and depends upon d/l.

Tables X and XI greatly facilitate the task of designing an inductance to given requirements.

Another form of calculation is as follows

$$L = \frac{0{\cdot}2A^2N^2}{3A + 9B + 10C}\ \mu\text{H.} \quad . \quad . \quad . \quad . \quad . \quad (6)$$

where A = mean diameter of coil in inches;
B = length of winding in inches;
C = radial depth of winding in inches;
N = total number of turns.

In the case of a single layer coil, C may be neglected.

To Calculate the Number of Turns of a Single Layer Coil.

$$N = \sqrt{\frac{3A + 9B}{0{\cdot}2A^2}L} \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

Multi-layer Air-cored Coils. It is very difficult to conceive a useful set of tables for such inductances due to the number of variable quantities involved. A useful approximate formula is that due to Mr. J. H. Reyner, thus—

$$L = \frac{0{\cdot}023N^2D\,(l - 2{\cdot}25c/D)}{1 + 2{\cdot}3b/D}\ \mu\text{H.,} \quad . \quad . \quad . \quad (8)$$

where D = outside diameter in cm.;
c = radial length in cm.;
b = length of winding in cm.

TABLE X

Inductance of Air-cored Coils of Circular Section

Inductance of a coil wound with 10 turns per cm. Values given in μH

Length (cm.)	Diameter (cm.) 4	5	6	7	8	9	10	12	14	16	18	Length (cm.)
1	5·78	7·89	10·14	12·49	14·94	17·47	20·06	—	—	—	—	1
2	16·59	23·28	30·5	38·1	46·2	54·5	63·1	81·12	99·7	119·5	139·7	2
3	29·5	42·25	56·02	70·87	86·57	97·15	120·0	155·8	196·0	233·0	273·7	3
4	43·4	63·0	84·57	107·9	132·8	159·0	186·3	243·5	305·2	369·5	436·0	4
5	58·0	84·92	115·0	147·6	179·6	220·3	259·2	342·2	430·7	502·3	620·5	5
6	72·9	107·6	146·7	190·2	236·2	300·2	337·5	448·2	567·5	698·7	823·7	6
7	87·9	130·7	179·3	233·1	291·2	353·7	420·2	560·7	711·7	873·2	1 046	7
8	103·3	154·2	212·5	277·2	348·0	423·7	504·0	676·7	863·5	1 062	1 278	8
9	118·7	177·8	246·1	322·7	406·0	470·5	590·5	800·7	1 020	1 256	1 512	9
10	134·2	201·8	280·0	368·0	464·5	568·5	679·5	920·2	1 181	1 464	1 762	10
12	165·2	249·3	348·7	460·5	586·0	717·5	860·7	1 175	1 518	1 886	2 283	12
14	196·4	298·2	418·2	554·0	702·5	868·7	1 046	1 435	1 862	2 329	2 822	14
16	227·7	347·2	487·8	648·2	825·2	1 023	1 234	1 701	2 219	2 785	3 400	16
18	259·0	395·7	557·8	743·0	950·0	1 178	1 424	1 970	2 577	3 245	3 962	18
20	290·5	444·7	628·0	838·0	1 074	1 333	1 615	2 241	2 945	3 715	4 547	20
22	321·7	493·7	698·5	933·5	1 205	1 490	1 807	2 515	3 312	4 187	5 140	22
24	353·5	543·0	751·2	1 038·7	1 322	1 647	2 000	2 790	3 682	4 667	5 735	24
26	384·0	591·7	889·2	1 125	1 447	1 845	2 194	3 067	4 057	5 152	6 342	26
28	416·5	641·2	910·5	1 221	1 572	1 961	2 421	3 347	4 432	5 637	6 952	28
30	448·0	690·2	980·5	1 316	1 697	2 120	2 582	3 625	4 807	6 125	7 565	30

TABLE XI

Data for Air-cored Single Layer Coils

The columns designated "Relative Inductance" give the ratio of inductance of a coil wound with a given wire to that of a similar coil wound with 10 turns per cm. as in Table X

Wire		Single Silk-covered			Double Silk-covered			S.W.G.
S.W.G.	Bare Wire Diameter (mm.)	Turns per cm.	Turns per in.	Relative Inductance	Turns per cm.	Turns per in.	Relative Inductance	
40	0·122	62·2	159	38·68	55·9	142	31·24	40
38	0·152	52·3	133	27·36	47·6	121	22·64	38
36	0·193	43·3	110	18·76	40·0	102	16	36
34	0·234	36·8	93·4	13·56	33·6	85·5	11·28	34
32	0·274	32·0	81·3	10·24	29·6	75·2	8·76	32
30	0·315	28·2	72	7·96	25·5	67·1	6·72	30
28	0·376	23·9	60·4	5·72	22·1	56·2	4·88	28
26	0·457	20·3	50·6	4·00	18·7	47·6	3·496	26
24	0·559	16·7	42·1	2·788	15·7	40	2·464	24
22	0·711	13·2	33·3	1·744	12·8	31·8	1·64	22
20	0·914	10·35	26·3	1·072	9·9	25·3	0·98	20
18	1·219	7·87	20	0·620	7·64	19·6	0·584	18
16	1·626	5·87	15	0·344	5·75	14·7	0·332	16
14	2·032	4·76	12	0·228	4 65	11·8	0·216	14
12	2·642	3·68	9·2	0·136	3·66	9·39	0·132	12

The formula of (6) is also applicable provided the factor C is taken into account.

A formula by Mr. P. R. Coursey is of the same form as Nagaoka's formula thus—

$$L = 0{\cdot}001\pi^2(N^2d^2/l)K' \ \mu\text{H.}, \qquad . \quad . \quad . \quad . \quad (9)$$

where K is a ratio of diameter to length and length to thickness.

Dr. F. W. Grover, in *Scientific Papers*,* gave the factor K of Mr. Coursey's formula in tabular form, which is reproduced here in Tables XII and XIII. In Eq. (9) and Tables XII and XIII the following notation is used.

$d = 2a =$ mean diameter in cm.;
$l =$ axial length;
$c =$ radial depth or thickness of coil in cm.

Table XIV gives the turns and physical dimensions of air-cored coils often used for filter circuits in gramophone reproduction, etc.

Table XVI gives winding data for unscreened coils, of 2 100 μH. for use on the long wave tuning range of broadcast receivers.

Table XV gives similar data for coils of 175 μH. for use on the medium wave band.

* Bureau of Standards, No. 455.

TABLE XII

VALUES OF K' FOR USE IN EQUATION 6

c/d	*l/c*										*c/d*
	0·1	0·2	0·3	0·4	0·5	0·6	0·7	0·8	0·9	1·0	
0·025	0·0071	0·0140	0·0206	0·0270	0·0332	0·0392	0·0450	0·0508	0·0568	0·0618	0·025
0·05	0·0121	0·0235	0·0346	0·0451	0·0553	0·0651	0·0747	0·0839	0·0928	0·1015	0·05
0·1	0·0197	0·0383	0·0559	0·0727	0·0887	0·1040	0·1186	0·1327	0·1426	0·1593	0·1
0·2	0·0307	0·0592	0·0857	0·1106	0·1339	0·1559	0·1767	0·1963	0·2149	0·2326	0·2
0·3	0·0386	0·0737	0·1061	0·1360	0·1637	0·1894	0·2134	0·2538	0·2568	0·2764	0·3
0·4	0·0445	0·0845	0·1209	0·1541	0·1844	0·2122	0·2379	0·2615	0·2834	0·3036	0·4
0·5	0·0491	0·0928	0·1320	0·1673	0·1991	0·2281	0·2544	0·2785	0·3005	0·3208	0·5
0·6	0·0529	0·0994	0·1406	0·1772	0·2100	0·2394	0·2660	0·2901	0·3119	0·3319	0·6
0·7	0·0561	0·1049	0·1476	0·1851	0·2185	0·2481	0·2747	0·2986	0·3201	0·3395	0·7
0·8	0·0590	0·1097	0·1536	0·1920	0·2257	0·2555	0·2820	0·3056	0·3268	0·3457	0·8
0·9	0·0617	0·1142	0·1594	0·1986	0·2328	0·2627	0·2892	0·3126	0·3336	0·3524	0·9
1·0	0·0645	0·1189	0·1654	0·2055	0·2403	0·2707	0·2972	0·3207	0·3416	0·3603	1·0

TABLE XIII

VALUES OF K' FOR USE IN EQUATION (4)

c/d	*c/l*										*c/d*
	0·1	0·2	0·3	0·4	0·5	0·6	0·7	0·8	0·9	1·0	
0·025	0·3498	0·2219	0·1650	0·1322	0·1107	0·0953	0·0838	0·0749	0·0677	0·0618	0·025
0·05	0·4947	0·3361	0·2574	0·2097	0·1774	0·1540	0·1362	0·1222	0·1109	0·1015	0·05
0·1	0·6276	0·4674	0·3735	0·3119	0·2682	0·2356	0·2102	0·1899	0·1732	0·1593	0·1
0·2	0·7012	0·5788	0·4863	0·4204	0·3703	0·3310	0·2992	0·2731	0·2512	0·2326	0·2
0·3	0·7051	0·6069	0·5301	0·4697	0·4210	0·3813	0·3483	0·3206	0·2969	0·2764	0·3
0·4	0·6907	0·6110	0·5453	0·4912	0·4465	0·4084	0·3761	0·3485	0·3246	0·3036	0·4
0·5	0·6717	0·6045	0·5775	0·4990	0·4576	0·4223	0·3916	0·3649	0·3415	0·3208	0·5
0·6	0·6533	0·5949	0·5443	0·5004	0·4624	0·4293	0·4003	0·3747	0·3520	0·3319	0·6
0·7	0·6379	0·5868	0·5399	0·4996	0·4642	0·4330	0·4054	0·3809	0·3591	0·3395	0·7
0·8	0·6268	0·5791	0·5368	0·4992	0·4658	0·4362	0·4098	0·3862	0·3650	0·3457	0·8
0·9	0·6207	0·5762	0·5364	0·5007	0·4689	0·4404	0·4149	0·3919	0·3712	0·3524	0·9
1·0	0·6200	0·5778	0·5397	0·5054	0·4746	0·4469	0·4219	0·3993	0·3789	0·3603	1·0

TABLE XIV

WINDING DATE: MULTI-LAYER COILS FROM 0·066 HENRY TO 0·5 HENRY

The former for use with these coils is shown in Fig. 8

Inductance (H.)	36 D.S.C. Turns	Inductance (H.)	36 D.S.C. Turns
0·066	2 000	0·3	3 850
0·075	2 250	0·325	4 000
0·1	2 400	0·35	4 150
0·125	2 620	0·375	4 250
0·15	2 870	0·4	4 350
0·175	3 120	0·425	4 450
0·2	3 250	0·45	4 550
0·225	3 370	0·475	4 700
0·25	3 500	0·5	4 840
0·275	3 700		

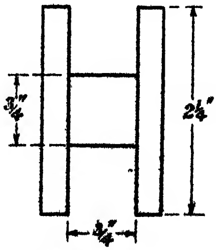


FIG. 8. COIL FORMER FOR INDUCTANCES OF TABLE XIV

TABLE XV

INDUCTANCES: 2 200 μH.

Wound on slotted hexagonal formers, slots being $\frac{1}{16}$ in. wide, $\frac{1}{8}$ in. deep, and spaced $\frac{1}{8}$ in.

Diameter (in.)	Slots	Wire (S.W.G.)	Turns per Slot
1·5	3	36 Enam.	81
1·5	4	36 D.S.C	65
2·0	3	36 D.S.C.	65
2·0	4	34 D.S.C.	51

TABLE XVI

INDUCTANCES: 175 μH.

Diameter (in.)	Turns	Wire (S.W.G.)	Winding Length (in.)
1·5	74	30 D.S.C.	1·1
2·0	58	28 D.S.C.	1·0
2·25	52	28 D.S.C.	0·9
2·5	58	24 D.C.C.	1·8
3·0	50	22 D.C.C.	1·9

Iron-cored Coils. In the case of a coil where the iron circuit is almost completely closed as in a choke coil a convenient formula is

$$L = 4N^2\mu A/l10^9 \text{ henries.} \qquad (10)$$

where μ = permeability of the air gap ($\mu = 1$);

l = length of air gap in cm.;

A = cross-section normal to the flux at the gap (i.e. the area of the iron surface at the gap);

N = total number of turns.

An approximate rule is that inductance varies as the square of the number of turns. Thus one-half of the number of turns gives about one-fourth the inductance.

In Tables XVII and XVIII the flux density is that with pure d.c. alone.

The actual gap can only be an approximation but is close enough for all normal purposes.

TABLE XVII

Inductance of Iron-cored Chokes

Inductance (H.)	Cross-section Core (in. × in.)	Equivalent Gap (in.)	Actual Gap (in.)	Wire* (S.W.G.)	Turns	Flux Density (Gauss)	D.C. Resistance (Ω)
0·5	½ × ½	0·040	0·017	36	1 600	6 500	83
1·0	„	0·041	0·019	36	2 300	9 000	129
5·0	„	0·043	0·023	36	5 200	20 000	351
10·0	„	0·046	0·030	36	7 600	27 000	553
15·0	„	0·048	0·035	36	9 500	32 000	737
5·0	¾ × ¾	0·043	0·023	36	3 500	13 000	275
10·0	„	0·046	0·030	36	5 000	18 000	418
15·0	„	0·048	0·035	36	6 300	21 000	544
20·0	„	0·052	0·044	36	7 600	24 000	689
50·0	„	0·070	0·100	36	14 000	33 000	1 459
100·0	2 × 2	0·100	0·250	36	8 900	14 000	1 614
10·0	1 × 1	0·046	0·030	36	3 800	14 000	371
15·0	„	0·048	0·035	36	4 800	16 000	487
20·0	„	0·052	0·044	36	5 700	18 000	592
50·0	„	0·070	0·100	36	11 000	25 000	1 296
100·0	„	0·100	0·250	36	18 000	29 000	2 302

* Current-carrying capacity = 50 mA.

TABLE XVIII

Inductance of Iron-cored Chokes

Inductance (H.)	Cross-section Core (in. × in.)	Equivalent Gap (in.)	Actual Gap	Wire* (S.W.G.)	Turns	Flux Density (Gauss)	D.C. Resistance (Ω)
0·5	½ × ½	0·040	0·017	33	1 600	13 000	46
1·0	„	0·041	0·019	33	2 300	18 000	72
5·0	„	0·043	0·023	33	5 200	39 000	200
1·0	¾ × ¾	0·041	0·019	33	1 500	12 000	56
5·0	„	0·043	0·023	33	3 500	26 000	151
10·0	„	0·046	0·030	33	5 000	35 000	230
5·0	1 × 1	0·043	0·023	33	2 600	20 000	130
10·0	„	0·046	0·030	33	3 800	27 000	200
15·0	„	0·048	0·035	33	4 800	32 000	260
10·0	2 × 2	0·046	0·030	33	1 900	13 000	160
15·0	„	0·048	0·035	33	2 400	16 000	200
20·0	„	0·052	0·044	33	2 900	18 000	250
50·0	„	0·070	0·100	33	5 300	24 000	480
100·0	„	0·100	0·250	33	4 800	28 000	860

* Current-carrying capacity = 100 mA.

SINGLE TURN LOOPS AND INDUCTANCE OF STRAIGHT CONDUCTORS

The inductance of a straight wire of length l cm. and radius r cm. is given by

$$L = 0{\cdot}002l[2{\cdot}30 \log_{10} (2l/r) - k_1] \,\mu\text{H.}, \qquad (11)$$

where k_1 is 0·75 for a non-magnetic material at low frequencies, and 1·0 at high frequencies. For a magnetic material $k_1 = 1 - \mu/4$ where μ is the permeability.

The inductance of the above wire with a parallel return at a distance d cm. is

$$L = 0{\cdot}004l[2{\cdot}30 \log_{10}(d/r) + k_2] \,\mu\text{H.}, \qquad (12)$$

where k_2 is 0·25 for low frequencies and decreases to zero at very high frequencies.

Tables XVIIIA and XVIIIB give the inductance of a single turn loop, and straight wire respectively.

TABLE XVIIIA

SINGLE TURN LOOP
Inductance in Centimetres

Loop Radius a (cm.)	$r =$ 0·05 cm.	$r =$ 0·1 cm.	$r =$ 0·15 cm.	$r =$ 0·2 cm.
5	310·4	266·5	241	223
10	707	618	567	531
15	1 137	1 006	930	875
20	1 590	1 414	1 312	1 238
25	2 057	1 840	1 710	1 622
30	2 536	2 275	2 123	2 015
35	3 030	2 722	2 548	2 420
40	3 528	3 175	2 975	2 830
45	4 040	3 640	3 410	3 250
50	4 547	4 110	3 850	3 680

Note. 1 000 cm. = 1μH.

$$L = 4\pi a\left(\log_e \frac{8a}{r} - 1{\cdot}75\right) \text{cm. (Kirchoff)}$$

where a = radius of loop in cm.,

r = radius of wire in cm.

TABLE XVIIIb

INDUCTANCE OF A STRAIGHT WIRE

Inductance in Centimetres

Length of Wire (cm.)	r = 0·025 cm.	r = 0·05 cm.	r = 0·1 cm.
10	113·7	99·8	86
20	255·5	227·5	200
30	407	365·5	324
40	565	510	455
50	729	660	591
60	898	815	731
70	1 068	970	874
80	1 242	1 130	1 020
90	1 418	1 293	1 170
100	1 597	1 459	1 320
125	2 053	1 880	1 706
150	2 515	2 310	2 102
175	2 990	2 750	2 510
200	3 476	3 192	2 916
300	5 450	5 030	4 620
400	7 500	6 945	6 400
500	9 596	8 903	8 210

Note. 1 000 cm. = 1μH.

$$L = 2l\left(\log_e \frac{2l}{r} - 1\right) \text{ cm.} \quad \text{(Neumann's formula)}$$

where l = length of wire in cm.;
r = radius of wire in cm.

SHIELDED COIL INDUCTANCES

In most radio receivers of current design it is usual to find most if not all of the tuning circuit inductances shielded.

When an inductance is shielded it is necessary to make a correction factor in its effective inductance. The problem was tackled by the R.C.A. engineers and the full details were given in *Radio Engineering.**

The solution is as follows. If the shield is considered as a single turn of wire round the coil, the decrease in reactance of the coil is shown to be

$$\omega^2 M^2/\omega L_s \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

where M = the mutual inductance between coil and shield;
and L_s = the inductance of the shield.

The resistance of the shield is assumed to be small compared with its reactance.

* July, 1935.

The coefficient of coupling between the coil and shield is

$$K = M/\sqrt{(L_a L_s)} \quad . \quad . \quad . \quad . \quad . \quad (14)$$

where L_a is the inductance of the coil without shield.

Substituting K for M in Eq. (13), we obtain for the decrease in reactance of L_a—

decrease in

$$X_a = \omega L_a K^2; \quad . \quad . \quad . \quad . \quad . \quad (15)$$

decrease in

$$L_a = K^2 L_a \quad . \quad . \quad . \quad . \quad . \quad . \quad (16)$$

The inductance of the coil when shielded is therefore

$$L = L_a(1 - K^2) \quad . \quad . \quad . \quad . \quad . \quad (17)$$

MUTUAL INDUCTANCE

This section is an abridged digest of Dr. F. W. Grover's paper in the *Proceedings of the Institute of Radio Engineers.**

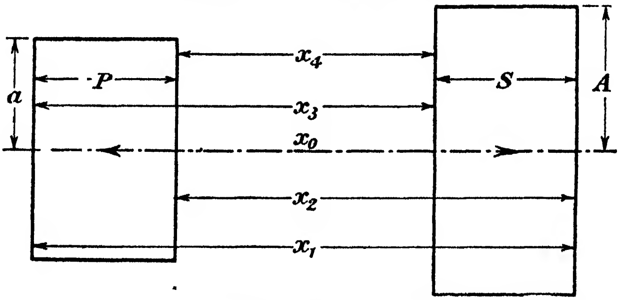


FIG. 9. BASIC ARRANGEMENT FOR CALCULATION OF MUTUAL INDUCTANCE

The tables, used in conjunction with a single formula, will give the mutual inductance with an accuracy sufficient for all purposes excepting those of refined measurements. The formulae apply strictly to steady current values of the mutual inductance: also inductances of the single-layer type are assumed.

The absolute elliptic integral formulae give the mutual inductance of coaxial solenoids as a combination of four integrations, each of which depends upon one of the four axial distances between the end planes of one of the coils and those of the other.

* Vol. XXI. 1933, No. 7.

The formula used by J. Clem* is useful for such routine calculations. Assume that the coils have axial lengths P and S (Fig. 9), radii a and A (A being the larger), winding densities n_1 and n_2 turns per centimetre, and that the distance between their centres is x_0.

Then the distance between their ends

$$x_1 = x_0 + \tfrac{1}{2}(P + S),$$
$$x_2 = x_0 + \tfrac{1}{2}(P - S),$$
$$x_3 = x_0 - \tfrac{1}{2}(P - S),$$
$$x_4 = x_0 - \tfrac{1}{2}(P + S),$$

are to be calculated, and the four diagonals

$$r_n = \sqrt{(x_n^2 + A^2)}.$$

corresponding to the four values of x_n.

Taking Clem's formula, mutual inductance M is given by

$$M = \frac{2\pi^2 a^2 n_1 n_2}{10^9}[r_1B_1 - r_2B_2 - r_3B_3 + r_4B_4] \text{ henries} \quad (18)$$

in which B_n quantities are obtained from the tables. All dimensions are in centimetres. Eq. (18) applies to coaxial coils.

In the case of concentric coils $x_0 = 0$, and, since only the numerical values of the x_n are required,

$$x_4 = x_1 = \tfrac{1}{2}(P + S)$$

and

$$x_3 = x_2 = [(P - S)/2],$$

so that for this special case

$$M = \frac{4\pi^2 a^2 n_1 n_2}{10^9}[r_1B_1 - r_2B_2] \quad . \quad . \quad . \quad (19)$$

The quantities B_n are functions of the radii and the spacing of the coils, and require for their specification two parameters only. For Dr. Grover's tables (Tables XIX, XX, and XXI) the parameters chosen are $\alpha = a/A$ and $r_n^2 = A^2/r_n^2$, each of which must lie between the limits of zero and unity. Table XIX gives the values of B_n to five figures in steps of 0·05 in the values of α and r_n^2. From this interpolations may be made by using the Newton interpolation formula.

Table XX covers the range of values of α and r_n^2 from 0·9 to 1·0 where interpolation in Table XIX is difficult. Table XXI is for calculations with coaxial coils of equal radii $\alpha = 1$, with a tabular interval in r_n^2 of 0·01.

* *Jour. Am. I.E.E.*, Vol. XLVI, 1927, No.

TABLE XIX

Values of B_n as a Function of α and β^2

Part I

α	r^2 = 1·00	0·95	0·90	0·85	0·80	0·75	0·70	0·65	0·60	0·55	0·50
1·00	0·84883	0·87727	0·89552	0·91020	0·92264	0·93345	0·942975	0·95144	0·95900	0·96576	0·97180
0·95	0·86783	0·889825	0·90561	0·91859	0·92971	0·939435	0·94805	0·95573	0·96261	0·96877	0·97428
0·90	0·88418	0·90175	0·91531	0·926665	0·93655	0·945245	0·95298	0·95990	0·96612	0·97169	0·97668
0·85	0·89870	0·91296	0·924555	0·93444	0·94314	0·95085	0·95774	0·96393	0·96951	0·974515	0·97901
0·80	0·91176	0·923445	0·933285	0·94185	0·94944	0·95622	0·96231	0·96781	0·922765	0·97723	0·98124
0·75	0·923565	0·93318	0·94150	0·94885	0·95542	0·96132	0·96668	0·97151	0·97588	0·97983	0·98338
0·70	0·93426	0·94217	0·94917	0·95543	0·96107	0·96618	0·97082	0·97503	0·97884	0·98230	0·98541
0·65	0·94394	0·95045	0·95629	0·96157	0·96637	0·97074	0·97472	0·97835	0·981645	0·984635	0·98732
0·60	0·95270	0·95803	0·96286	0·96727	0·97130	0·97499	0·97837	0·98146	0·98427	0·98683	0·98913
0·55	0·96060	0·96492	0·96888	0·972515	0·97586	0·978935	0·98176	0·98435	0·98672	0·98887	0·99082
0·50	0·96769	0·97115	0·97434	0·97730	0·98003	0·982555	0·98488	0·98702	0·98897	0·99076	0·99237
0·45	0·97400	0·97673	0·97927	0·98163	0·98382	0·985845	0·98772	0·98945	0·99103	0·99248	0·99379
0·40	0·97958	0·98169	0·98366	0·985495	0·98721	0·98880	0·990275	0·99164	0·99289	0·994035	0·99508
0·35	0·98444	0·98603	0·98751	0·98890	0·99020	0·991415	0·99254	0·99358	0·99454	0·99542	0·99622
0·30	0·98862	0·989765	0·99084	0·99186	0·99280	0·99369	0·99451	0·99527	0·99598	0·99662	0·99721
0·25	0·992125	0·99291	0·99365	0·99435	0·99500	0·99561	0·99618	0·99671	0·99720	0·99765	0·99806
0·20	0·994975	0·99547	0·99594	0·99638	0·99680	0·99719	0·99755	0·99789	0·99821	0·99849	0·99875
0·15	0·99718	0·99746	0·99772	0·99797	0·99820	0·99842	0·99862	0·99881	0·99899	0·99915	0·99930
0·10	0·99875	0·99887	0·99899	0·99910	0·99920	0·99930	0·99939	0·99947	0·99955	0·99962	0·99969
0·05	0·99969	0·99972	0·99910	0·99977	0·99980	0·99982	0·99985	0·99987	0·99989	0·99995	0·99992
0·00	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000

Part II

α	$r^2 = 0{\cdot}50$	0·45	0·40	0·35	0·30	0·25	0·20	0·15	0·10	0·05	0·00
1·00	0·97180	0·97718	0·98194	0·98612	0·98974	0·992815	0·99535	0·99735	0·99880	0·999695	1·000
0·95	0·97428	0·97919	0·98354	0·98736	0·99066	0·99346	0·99577	0·99759	0·99891	0·99972	1·000
0·90	0·97668	0·98114	0·98509	0·98855	0·99155	0·99409	0·99618	0·99783	0·99902	0·99975	1·000
0·85	0·97901	0·98302	0·98658	0·98970	0·99240	0·99469	0·99657	0·99805	0·99912	0·99978	1·000
0·80	0·98124	0·98483	0·98801	0·99080	0·99322	0·99526	0·99695	0·99827	0·99922	0·99980	1·000
0·75	0·98338	0·986555	0·989375	0·99185	0·99399	0·99581	0·99730	0·99847	0·99931	0·99983	1·000
0·70	0·98541	0·98820	0·99068	0·99285	0·99437	0·99633	0·99764	0·99866	0·99940	0·99985	1·000
0·65	0·98732	0·98975	0·99190	0·993795	0·99543	0·996815	0·99795	0·99884	0·99948	0·99987	1·000
0·60	0·98913	0·99121	0·99306	0·99467	0·99608	0·99727	0·99825	0·99901	0·99956	0·99989	1·000
0·55	0·99082	0·99257	0·99413	0·995505	0·99669	0·99770	0·99852	0·999165	0·99963	0·999905	1·000
0·50	0·99237	0·99383	0·99512	0·996285	0·99725	0·99809	0·99877	0·99931	0·99969	0·99992	1·000
0·45	0·99379	0·99498	0·99603	0·99696	0·99776	0·998445	0·99900	0·99944	0·99975	0·99994	1·00
0·40	0·99508	0·99601	0·99685	0·99759	0·99823	0·99877	0·99921	0·99955	0·99980	0·99995	1·00
0·35	0·99622	0·99694	0·99758	0·99815	0·99864	0·99905	0·99939	0·99966	0·99985	0·99996	1·00
0·30	0·99721	0·99774	0·99822	0·998635	0·99900	0·99930	0·99955	0·99975	0·99989	0·99997	1·00
0·25	0·99806	0·99843	0·99876	0·99905	0·99930	0·999515	0·99969	0·999825	0·99992	0·99998	1·00
0·20	0·99875	0·99899	0·99920	0·99939	0·99955	0·99969	0·99980	0·99989	0·99995	0·99999	1·00
0·15	0·99930	0·99943	0·99955	0·99966	0·99975	0·999825	0·99989	0·99994	0·99997	0·99999	1·00
0·10	0·99969	0·99975	0·99980	0·99985	0·99989	0·99992	0·99995	0·99997	0·99999	1·00000	1·00
0·05	0·99992	0·99994	0·99995	0·99996	0·99997	0·99998	0·99999	0·99999	1·00000	1·00000	1·00
1·00	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·000	1·00

TABLE XX

Auxiliary Table for Larger Values of a and r^2

a	$r^2 = 1{\cdot}00$	0·99	0·98	0·97	0·96	0·95	0·94	0·93	0·92	0·91	0·90
1·00	0·84883	0·85698	0·86298	0·86820	0·87292	0·87727	0·88133	0·88515	0·88877	0·892225	0·89552
0·99	0·85294	0·86035	0·86606	0·87107	0·87562	0·879825	0·88376	0·88747	0·89100	0·89346	0·89757
0·98	0·85686	0·863665	0·86910	0·87391	0·87829	0·88236	0·88617	0·88978	0·893205	0·89647	0·89960
0·97	0·86063	0·86693	0·87210	0·876715	0·88094	0·88487	0·88857	0·89207	0·89539	0·898575	0·90162
0·96	0·86428	0·87014	0·875065	0·87949	0·88356	0·88736	0·89094	0·89433	0·89757	0·90066	0·903625
0·95	0·86783	0·87329	0·877985	0·88223	0·88615	0·889825	0·89329	0·89658	0·89972	0·90273	0·90561
0·94	0·87127	0·87639	0·88086	0·88494	0·88872	0·89226	0·89562	0·89881	0·90186	0·90478	0·90759
0·93	0·87462	0·879435	0·88370	0·88761	0·89125	0·894675	0·89792	0·901015	0·90397	0·90681	0·90954
0·92	0·87788	0·88242	0·88649	0·890245	0·89375	0·89706	0·90020	0·90320	0·90607	0·90883	0·91148
0·91	0·88107	0·88536	0·889245	0·89285	0·89622	0·89942	0·90246	0·90536	0·90815	0·91082	0·91340
0·90	0·88418	0·88824	0·89195	0·89541	0·89866	0·90175	0·90469	0·907505	0·91020	0·91280	0·91531

TABLE XXI

VALUES FOR B^n FOR COILS OF EQUAL RADII ($a = 1$)

r^2	B_n	r^2	B_n	r^2	B_n	r^2	B_n
0	1·00	0·25	0·992815	0·50	0·971802	0·75	0·933448
0·01	0·999987	0·26	0·2244	0·51	0·970649	0·76	0·931397
0·02	0·999950	0·27	0·1650	0·52	0·969469	0·77	0·929294
0·03	0·99889	0·28	0·7035	0·53	0·8262	0·78	0·7135
0·04	0·99804	0·29	0·990399	0·54	0·7027	0·79	0·4918
0·05	0·999695	0·30	0·989742	0·55	0·965763	0·80	0·922639
0·06	0·99562	0·31	0·9062	0·56	0·4471	0·81	0·920297
0·07	0·99407	0·32	0·8360	0·57	0·3149	0·82	0·917886
0·08	0·99228	0·33	0·7637	0·58	0·1798	0·83	0·5403
0·09	0·99026	0·34	0·6891	0·59	0·960416	0·84	0·2843
0·10	0·998802	0·35	0·986123	0·60	0·959002	0·85	0·910202
0·11	0·8556	0·36	0·5332	0·61	0·7558	0·86	0·907472
0·12	0·8287	0·37	0·4520	0·62	0·6080	0·87	0·4648
0·13	0·7996	0·38	0·3684	0·63	0·4570	0·88	0·901721
0·14	0·7684	0·39	0·2826	0·64	0·3024	0·89	0·898683
0·15	0·997349	0·40	0·981944	0·65	0·951443	0·90	0·895522
0·16	0·6992	0·41	0·1039	0·66	0·949826	0·91	0·892225
0·17	0·6614	0·42	0·980110	0·67	0·8172	0·92	0·888774
0·18	0·6214	0·43	0·979158	0·68	0·6480	0·93	0·5154
0·19	0·5793	0·44	0·8182	0·69	0·4748	0·94	0·881327
0·20	0·995351	0·45	0·977181	0·70	0·942975	0·95	0·877266
0·21	0·4886	0·46	0·6156	0·71	0·941161	0·96	0·872917
0·22	0·4401	0·47	0·5106	0·72	0·939302	0·97	0·868201
0·23	0·3894	0·48	0·4031	0·73	0·7398	0·98	0·862983
0·24	0·3366	0·49	0·2930	0·74	0·5448	0·99	0·856980
						1·00	0·848826

There are numerous other formulae for obtaining mutual inductance and these are given below.

The mutual inductance of two circles is given by

$$M = 0{\cdot}001 M_0 \sqrt{(Aa)}\ \mu\text{H.}, \qquad . \quad . \quad . \quad (20)$$

where M_0 is a function of R_2/R_1, R_2 and R_1 being the least and greatest distances between circles.

A full exposition of this formula is to be found in Mr. Nottage's book, *The Calculation and Measurement of Inductance and Capacity*.*

When two circles are near together and of nearly the same diameter,

$$M = 0{\cdot}02895 \log_{10} 8a/R - 0{\cdot}109a\ \mu\text{H.} \qquad . \quad . \quad (21)$$

* Iliffe & Sons Ltd. Second Edition, 1925.

where R = the distance $\sqrt{(c^2 + d^2)}$;
d = axial distance between centres;
$c = A - a$;
A = radius of the larger circle;
a= radius of the smaller circle;

all dimensions being in centimetres.

The mutual inductance between two parallel wires of cylindrical section is given by

$$M = 0{\cdot}002\left[l 2{\cdot}303 \log_{10}\frac{l + \sqrt{(l^2 + d^2)}}{d} - \sqrt{(l^2 + d^2)} + d\right] \quad (22)$$

where l = length of either wire;
d = distance between them;

all dimensions being in centimetres.

Professor Lyle has given a very convenient method for calculating M between two coaxial coils when the coils are some distance apart. Each coil is replaced by two equivalent circles. The mutual inductances for the four pairs of circles AC, AD, BC, BD are calculated, and the arithmetic mean taken. This gives the mutual inductance of the coils. Thus

$$M = 0{\cdot}25[M(A/C) + M(AD) + M(BC) + M(BD)] \quad (23)$$

The student desirous of making an intensive study of mutual inductance of coils of various shapes is recommended to read Mr. Nottage's book.

SECTION III

CONDENSERS

FIXED CONDENSERS

The formulae adopted in the calculation of capacitance are usually based on the electrostatic systems, therefore the answers are in electrostatic units, that is, *jars* or *centimetres.* Table XXII shows the relation between the electromagnetic unit, the *Farad*, and sub-multiples thereof, and the electrostatic units. Throughout this section all dimensions will be given in centimetres, except where otherwise stated.

The basic calculations for the commonly used parallel plate condenser are of such a simple order that it would be superfluous to prepare a table of physical dimensions for the wide range of capacitances used in radio work.

The capacitance of a parallel plate condenser consisting of two plates separated by a single sheet of dielectric, which may be air, is given by the formula—

$$C = \kappa A/4\pi d \text{ cm.,}$$
$$= 0{\cdot}07958(\kappa A/d) \text{ cm.,}$$
$$= 0{\cdot}08841(\kappa A/d)\ \mu\mu\text{F. (or picofarads)} \quad . \quad (24)$$

where A = area of one metal plate in cm.^{2};

d = distance between plates, or thickness of the dielectric, in cm.;

κ = dielectric constant of the dielectric.

Table XXIII gives the dielectric constants of many substances. This constant is often known as the *S.I.C.*—*specific inductive capacity.*

The capacitance of a condenser containing a number of plates is obtained from the following formula—

$$C = n\kappa A/4\pi d \text{ cm.,}$$
$$= 0{\cdot}7958 n\kappa A/d \text{ cm.,}$$
$$= 0{\cdot}08841 n\kappa A/d\ \mu\mu\text{F.} \quad . \quad . \quad . \quad (25)$$

where n = number of sheets of dielectric enclosed by metal plates.

TABLE XXII

CAPACITANCE RELATIONSHIPS

Farad (F.)	Microfarad (μF.)	Picofarad ($\mu\mu$F.)	Centimetre
1	10^{6}	10^{12}	9×10^{11}
10^{-6}	1	10^{5}	9×10^{5}
10^{-12}	10^{-6}	1	0·9
$1·111 \times 10^{-12}$	$1·111 \times 10^{-6}$	1·111	1

It should be noted that 1 centimetre is equal to 10^{-3} jars, i.e. 1 jar = 10^{3} cm.

TABLE XXIII

DIELECTRIC CONSTANTS (S.I.C.)

Dielectric	κ	Puncture Voltage	
		kV. per cm.	kV. per in.
Air (normal pressure) .	1·00	7·8–9·0	19·8–22·8
Flint Glass . . .	6·6–10	900	2 280
Mica (Indian ruby) . .	4·6–8	1 500	3 810
Paraffin Wax (solid) .	2·0–2·5	400	1 017
Sulphur	3·9–4·2	—	—
Castor Oil . . .	4·7	150	381
Porcelain . . .	4·4	—	—
Quartz	4·5	—	—
Common Glass . .	3·1–4	300–1 500	762–3 810
Bakelite . . .	5–6	—	—

A very informative book on the design and manufacture of fixed condensers has been written by Mr. Coursey of the Dubilier Condenser Company. It is called *Electrical Condensers** and is recommended to those desirous of making a complete study of this subject.

VARIABLE AIR CONDENSERS

Straight-Line Capacitance Law. The general mechanical construction consists of a set of metal vanes fixed to a spindle, by the rotation of which they are interleaved with a fixed set of vanes. The simplest type is where each set of vanes is of

* First Edition (Pitman).

semicircular pattern. The capacitance is given by the same formula as that for a multi-plate fixed condenser, thus—

$$C = n\kappa A/4\pi d \text{ cm.} \qquad (26)$$

which reduces to

$$C = \kappa_n \frac{(r_1^2 - r_0^2)}{4d} \frac{\theta}{360} \text{ cm.} \qquad (27)$$

$$= 6{\cdot}944\kappa_n/d(r_1^2 - r_0^2)\theta \times 10^{-4} \text{ cm.} \qquad (28)$$

$$= 7{\cdot}715\kappa_n d(r_1^2 - r_0^2)\theta \times 10^{-4}\ \mu\mu\text{F.} \qquad (29)$$

where r_1 = radius of the outside edge of the rotor plates, in cm.;
r_2 = radius of the inside edge of stator plates, in cm.;
n = number of air spaces or sheets of dielectric between the plates when meshed;
θ = angle in degrees through which the plates have been turned from minimum capacitance;
d = distance between plates, or thickness of dielectric in cm.

Square Law. Mr. Dudell in the *Electrician** gives a formula for a condenser of this type whereby the variation in capacitance is proportional to the square of θ. In Mr. Dudell's design the inside edge of the stator vanes is a semicircle concentric with the rotor vanes spindle, and the outer edge of the stator vanes is such that the moving plate never projects beyond it. The curve of the outside edge of the rotor plates is obtained as follows—

Let r be the radius of the inside edge of the stator vanes and x be one of the radii of the curve forming the outside edge of the rotor vanes. Then the formula for making the area bounded between the curve and the circle of radius r proportional to the square of the angle θ is

$$x^2 = 4\kappa\theta + r^2 \qquad (30)$$

where κ is such that

$$\kappa\theta^2 = A \qquad (31)$$

where A is the area of the rotor vanes meshed within the stator vanes.

The radius of the rotor vane at any given position of θ is found in the following manner. For small angular increments

* Vol. 74, 6th February, 1914.

$\partial\theta$ the incremental areas may be considered as sectors of circles of radius R. Since the area of a circle is

$$[a/(2 \times 57{\cdot}3)]R^2 \quad . \quad . \quad . \quad . \quad . \quad (32)$$

the incremental area

$$\partial A_0 = [\partial\theta/(2 \times 57{\cdot}3)]R_0^2 \quad . \quad . \quad . \quad (33)$$

where R_0 = radius of plate in cm.; and
A_0 = area of plate in cm.^{2}

Then

$$R_0 = \sqrt{(114{\cdot}6\partial A_0/\partial\theta)} \quad . \quad . \quad . \quad . \quad (34)$$

By considering $\partial\theta$ infinitely small, the exact value of R_0 is

$$R_0 = \sqrt{[114{\cdot}6(dA_0/d\theta)]} \quad . \quad . \quad . \quad (35)$$

and from

$$A_0 = 0{\cdot}000615\theta^2$$

$$dA_0/d\theta = 0{\cdot}00123\theta$$

$$\therefore \quad R_0 = \sqrt{(114{\cdot}6 + 0{\cdot}00123\theta)}$$

$$= 0{\cdot}376\sqrt{\theta}\text{ cm.} \quad . \quad . \quad . \quad . \quad (36)$$

This type of condenser is often referred to as a *straight line wavelength* condenser. However, this is hardly correct as the residual capacitance of the condenser at C min. prevents this being fully achieved.

Mr. Griffiths in *Experimental Wireless* (Vol. III, No. 28) of January, 1926, gave a formula for a corrected form of square law condenser which overcame this difficulty. This is given hereunder. Mr. Griffiths's own notation is used throughout when referring to his designs. It should be noted particularly that the distinguishing suffixes such as in C_4, R_4, C_2, R_2, etc., have no special significance: they are reproduced in the exact form employed by the original author.

$$C_2 = (a_2 + b_2)^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (37)$$

where C_2 = capacitance.

$$A_2 = k[a(\theta + b_2)^2 - \text{residual capacitance}] + K\theta \quad . \quad (38)$$

where A_2 = area in cm.^{2}

$$R_2 = \{114{\cdot}6[2ka_2(a_2\theta + b_2) + K]\}^{\frac{1}{2}} \quad . \quad . \quad . \quad (39)$$

where R_2 = radius in cm.;

$$a_2 = \frac{\sqrt{(\text{max. capacitance})} - \sqrt{(\text{residual capacitance})}}{180},$$

$$b_2 = \sqrt{(\text{residual capacitance})},$$

$$k = \frac{\text{total plate area} - 180K}{\text{maximum capacitance} - \text{residual capacitance}},$$

$$K = r^2/114{\cdot}6.$$

With Mr. Griffiths's corrected plate shape the relation that $d\lambda/d\theta$ shall be constant is almost perfectly fulfilled. The self-capacitance of the associated tuning coil should also be included in the factor b_2 for the highest degree of accuracy.

Straight Line Frequency. It is quite logical to assume that if it is possible to make a straight line wavelength condenser it is also possible to design one giving a constant change in frequency, since frequency is directly related to wavelength by the factor

$$\lambda \text{ (wavelength in metres)} = 300\,000/f \quad . \quad (40)$$

and

$$f\text{(kilocycles)} = 300\,000/\lambda \quad . \quad (41)$$

Such a condenser requires an inverse square law plate and may be developed from the corrected square law plate just described. From Mr. Griffiths's paper previously mentioned the following equations summarize the design formulae and constants.

$$C_3 = 1/(a_3\theta + b_3)^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (42)$$

C_3 = capacitance.

$$A_3 = k\left[\frac{1}{(a_3\theta + b_3)^2} - \left\{\begin{matrix}\text{residual}\\ \text{capacitance}\end{matrix}\right\}\right] + K(180 - \theta) \quad (43)$$

where A_3 = area.

$$R_3 = \left\{114{\cdot}6\left[\frac{2ka_3}{(a_3\theta + b_3)^3} + K\right]\right\}^{\frac{1}{2}} \quad . \quad . \quad . \quad (44)$$

where R_3 = radius

$$a_3 = \frac{1}{180}\left[\frac{1}{\sqrt{(\text{residual capacitance})}} - b_3\right]$$

$$b_3 = \frac{1}{\sqrt{(\text{maximum capacitance})}}.$$

k and K are derived in the same manner as for the square law plate. (See above.)

Logarithmic Law. The earliest example the Author can trace of a variable condenser following this law is in the *Bulletin of the Bureau of Standards*,* and an early instance of its use is in Dr. Kolster's Decremeter. The requirements are that

$$d\lambda/d\theta \text{ must be proportional to } \lambda.$$

The equation given for the radius from the centre to the edge of the moving vane is

$$R = \sqrt{(2bm\varepsilon^{m\theta} + r^2)} \quad . \quad . \quad . \quad . \quad (45)$$

where r is the radius of the circle space in which is situated the condenser.

$$\left.\begin{matrix} b \\ m \end{matrix}\right\} \text{are constants depending upon } C_{max.} \text{ and } C_{min.}$$

Mr. Griffiths has prepared a formula for such a condenser, and this is given herewith.

$$C_4 = a_4\varepsilon^{b_4\theta} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (46)$$

where C_4 = capacitance.

$$R_4 = \{114{\cdot}6[ka_4b_4\varepsilon^{b_4\theta} + K]\}^{\frac{1}{2}} \quad . \quad . \quad . \quad . \quad (47)$$

where R_4 = radius;

a_4 = residual capacity;

$$b_4 = \frac{\log\left(\begin{matrix}\text{maximum}\\ \text{capacitance}\end{matrix}\right) - \log\left(\begin{matrix}\text{residual}\\ \text{capacitance}\end{matrix}\right)}{78{\cdot}174};$$

k and K are as in previous examples. (See p. 28.)

This plate shape gives a constant percentage change of wavelength or frequency.

Eq. (46) thus can be written

$$C = a\varepsilon^{b\theta} \quad . \quad . \quad . \quad . \quad (48)$$

and this same remark applies to other such equations. Needless to say, the powers of the quantities remain unchanged.

CYLINDRICAL TUBE CONDENSER

This type of condenser is formed by one cylinder sliding over another.

* Vol. XI, p. 433, 1919.

When the distance separating the cylinders is small compared with the radius, the capacitance per unit length is

$$C = \frac{\frac{1}{2}\kappa}{\log \varepsilon(r_2/r_1)} \text{ cm.} \quad . \quad . \quad . \quad . \quad (49)$$

$$= \frac{0{\cdot}2413\kappa}{\log_{10}(r_2/r_1)} \quad . \quad . \quad . \quad . \quad . \quad (50)$$

where κ = dielectric constant of the dielectric between cylinders;

r_1 = radius of inner cylinder in cm.;

r_2 = radius of outer cylinder in cm.

Condensers of this type are often used for neutralizing the grid-anode capacitance of a triode valve when operating as a high-frequency amplifier. In this type of condenser the capacitance is proportional to the length of the inner cylinder projecting inside the outer cylinder; thus the relationship between capacitance change and scale reading is linear.

If the dielectric is air ($\kappa = 1$), Eq. (51) gives the design formula.

$$C = \frac{1{\cdot}112}{2 \log_e (r_2/r_1)} \quad . \quad . \quad . \quad . \quad . \quad (51)$$

where r_2 = radius of inner cylinder in cm.;

r_1 = radius of outer cylinder in cm.

SUPERSONIC HETERODYNE CONDENSERS

When the supersonic heterodyne receiver returned into favour in 1930 or thereabouts, the problem of designing a suitable plate shape for the local oscillator circuit arose since single knob control was demanded. The Author evolved a mathematical solution in January, 1931, and although many papers have been written since that date, some being in the form of graphical solution, and others of the trial and error type, the Author still considers his pure mathematical approach to be the best method after five years' use.

The design requires a knowledge of the plate shape of the signal frequency tuning condenser, and particularly its area in polar co-ordinates. This is given by

$$A = \tfrac{1}{2}\int \rho_1^2 d\theta - \tfrac{1}{2}\int r^2 d\theta \quad . \quad . \quad . \quad . \quad (52)$$

$$\therefore \quad C_1 = K_1\left(\tfrac{1}{2}\int \rho_1^2 d\theta - \tfrac{1}{2}\int r^2 d\theta\right) \quad . \quad . \quad . \quad (53)$$

where C_1 = capacitance of condenser referred to zero;
K_1 = a constant.

Differentiating Eq. (53) gives

$$\partial C_1/\partial C_2 = (K_1/2)\,(\rho_1{}^2 - r^2) \quad . \quad . \quad . \quad . \quad (54)$$

Now in the design of the oscillator condenser plate all we require to know is the radius and capacitance of the signal frequency tuning plate to obtain a solution which will give perfect tracking.

We may now obtain the oscillator plate capacitance by

$$C_2 = K_2(\tfrac{1}{2}\int \rho_2{}^2 d\theta - \tfrac{1}{2}\int r^2 d\theta) \quad . \quad . \quad . \quad . \quad (55)$$

where C_2 = oscillator capacitance in $\mu\mu$F. referred to zero.
Dividing Eq. (53) by Eq. (55) we get

$$\frac{C_1}{C_2} = B\frac{\int \rho_1{}^2 d\theta - \int r^2 d\theta}{\int \rho_2{}^2 d\theta - \int r^2 d\theta} \quad . \quad . \quad . \quad . \quad . \quad (56)$$

where $B = K_1/K_2$.

Or

$$\int \rho_2{}^2 d\theta - \int r^2 d\theta = (BC_2/C_1)\left(\int \rho_1{}^2 d\theta - r^2 d\theta\right) \quad . \quad (57)$$

Differentiating and particularly noting C_2/C_1 is a variable quantity

$$\rho_2{}^2 - r^2 = B\left[\frac{C_2}{C_1}(\rho_1{}^2 - r^2) + 2\frac{d}{d\theta}\left(\frac{C_2}{C_1}\right)(\tfrac{1}{2}\int \rho_1{}^2 d\theta - \tfrac{1}{2}\int r^2 d\theta)\right] \quad (58)$$

But from Eq. (53) we saw that

$$\tfrac{1}{2}\int \rho_1{}^2 d\theta - \tfrac{1}{2}\int r^2 d\theta = C_1/K_1.$$

$$\therefore \qquad \rho_2{}^2 - r^2 = B\left[\frac{C_2}{C_1}(\rho_1{}^2 - r^2) + \frac{2C_1}{K} + \frac{d}{d\theta}\left(\frac{C_2}{C_1}\right)\right] \quad . \quad (59)$$

The term $\frac{d}{d\theta}\left(\frac{C_2}{C_1}\right)$ is eliminated as follows.

Let f_2 = oscillator frequency in kc.;
f_1 = signal frequency in kc.;
L_2 = oscillator inductance in μH.;
L_1 = signal frequency inductance in μH.;
C_2 = oscillator condenser capacitance in $\mu\mu$F referred to zero;

C_1 = signal frequency condenser capacitance in $\mu\mu$F. referred to zero.

f_i = intermediate frequency in kc.;

$C_{osc.\,min.}$ = oscillator minimum capacitance in $\mu\mu$F.;

$C_{s\,f.\,min.}$ = signal frequency minimum capacitance in $\mu\mu$F.;

$C_0 = C_2 + C_{osc.\,min.}$ = total oscillator capacitance in circuit;

$C_R = C_1 + C_{s\,f.\,min.}$ = total signal frequency circuit capacitance.

Then $$f_2 - f_1 = f_i$$

which may be written

$$\frac{159\,150}{\sqrt{[L_2(C_2 + C_{osc.\,min.})]}} - \frac{159\,150}{\sqrt{[L_1(C_1 + C_{s.f.\,min.})]}} = f_i \quad . \quad (60)$$

This may be simplified thus

$$\frac{C_2 + C_{osc.\,min.}}{C_1 + C_{s.f.\,min.}} = \frac{1}{(D\sqrt{[C_1 + C_{s\,f.\,min.}]} + P)^2} \quad . \quad (61)$$

where $D = (f_i\sqrt{L_2})/159\,150$;

and $P = \sqrt{(L_2/L_1)}$.

Differentiating

$$d\left(\frac{C_0}{C_R}\right) = \frac{-DC_0^{\frac{3}{2}}}{C_R^2}\,dC_1 \quad . \quad . \quad . \quad (62)$$

Eq. (60) can be transformed thus

$$C_2/C_1 = C_0/C_R + [(C_{s.f.\,min.}/C_1) \times (C_0/C_R)] - C_{osc.\,min.}/C_1 \quad (63)$$

Differentiating and substituting Eq. (62)

$$C_1^2\frac{d}{d\theta}\left(\frac{C_2}{C_1}\right) = \left[C_{osc.\,min.} - \frac{C_{s.f.\,min.}C_0 + DC_1C_0^{\frac{3}{2}}}{C_R}\right]\frac{dC_1}{d\theta} \quad (64)$$

But from Eq. (53)

$$dC/d\theta = (K_1/2)\,(\rho_1^2 - r^2) \quad . \quad . \quad . \quad (65)$$

therefore

$$C_1^2\frac{d}{d\theta}\left(\frac{C_2}{C_1}\right) = \left[C_{osc.\,min.} - C_{s.f.\,min.}\frac{C_0}{C_R} - \frac{DC_1C_0^{\frac{3}{2}}}{C_R}\right]\frac{K_1}{2}(\rho_1^2 - r^2) \quad . \quad . \quad . \quad (66)$$

Substituting this for $(d/d\theta)\,(C_2/C_1)$ in Eq. (59)

$$\rho_2^2 - r^2 = B(\rho_1^2 - r^2)\left[\frac{C_2}{C_1} + \frac{C_{osc.\,min.}}{C_1} - \frac{C_{s.f.\,min.}(C_0/C_R)}{C_1} - \frac{DC_0^{\frac{3}{2}}}{C_1}\right] \quad . \quad . \quad . \quad . \quad (67)$$

then

$$\rho_2^2 - r^2 = B\sqrt{[(L_2/L_1)(C_0^3/C_R^3)]}(\rho_1^3 - r^2) \quad . \quad . \quad (68)$$

where $B = K_1/K_2 = (N_R - 1)/(N_0 - 1)$;

where N_0 = number of plates in oscillator condenser;

N_R = number of plates in signal frequency condenser.

Thus

$$\frac{\rho_2^2 - r^2}{\rho_1^2 - r^2} = \frac{N_R - 1}{N_0 - 1}\sqrt{\left[\frac{L_2}{L_1} \times \frac{C_0^3}{C_R^3}\right]} = \frac{N_R - 1}{N_0 - 1} \times \frac{L_1}{L_2} \times \left(\frac{f_1}{f_2}\right)^3 \quad . \quad . \quad . \quad . \quad . \quad (69)$$

This, then, is the formula for the curve of the oscillator plate. It is well to remember that this formula was derived with the understanding that four important constants of the signal frequency condenser were known, viz.—

(*a*) the radius of the rotor plate (ρ);

(*b*) the radius of the cutaway portion of the stator plate (the space occupied by the rotor plate spindle) (r);

(*c*) the frequency to which the signal frequency circuit tunes (f_1);

(*d*) the number of plates in the signal frequency condenser.

These quantities may be readily computed from previous sections dealing with plate shapes of other types of condensers; also the simple functions of the relation between frequency, inductance, and capacitance, thus—

$$f = [1/2\pi\sqrt{(LC)}] \cdot 10^6 \quad . \quad . \quad . \quad . \quad (70)$$

where L = inductance in μH.;

C = capacitance in $\mu\mu$F.

EFFICIENCY OF AIR DIELECTRIC CONDENSERS

A common method of expressing the figure of merit of a condenser is by the expression $R\omega C^2$ which is the product of the power factor and the capacitance.

The derivation is as follows. When an air condenser C has an equivalent series resistance R, it may be represented by a condenser C_0 formed of the solid dielectric circuits and supports and having all the losses R_0 in parallel with a loss-free condenser C'. By working out the energy relations on the assumption that the losses were very small, it is found that

$$R = R_0\omega C_0^2/\omega(C' + C_0)^2 = R_0\omega C_0^2\lambda\omega C^2 \quad . \quad (71)$$

and $$R\omega C^2 = R_0\omega C_0^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (72)$$

However, it has been pointed out that this expression is not a true representation of the goodness factor of an air condenser at radio frequencies.*

A modified form of expression, derived by Mr. Griffiths to give the effective resistance of an air condenser, is as follows—

$$R = R_s + a/\omega C^2 + 1/\beta\omega^2 C^2 \quad . \quad . \quad . \quad (73)$$

where R_s = conductor resistance of the plate systems;

$a/\omega C^2$ = the resistance equivalent to the inherent power loss in the insulation supports;

a = a constant proportional to the power loss factor of the insulating material;

$1/\beta\omega^2 C^2$ = the series resistance equivalent to the loss in the parallel resistance which is the insulation resistance β.

ELECTROLYTIC CONDENSERS

For this highly specialized electro-chemical subject the reader is referred to the following two articles, each of considerable length and wherein will be found extensive references—

"The Aluminium Electrolytic Condenser," Dr. Maddison. *Electrical Communication*, October, 1929.

"A.C. Electrolytic Capacitor." *Electrical Engineering*, October, 1935.

* Mr. Griffiths. *Wireless Engineer*, Vol. VIII, 1931, No. 90.

SECTION IV

TRANSFORMERS

Audio-frequency Transformers. From the outset it is desired to recommend readers who wish to make a complete study of this subject to a series of articles entitled "The Performance and Properties of Telephonic Frequency Intervalve Transformers," by the late Dr. D. W. Dye, which appeared in *Experimental Wireless*.* Although these papers are now twelve years old, the Author is not aware of any other such complete analysis of basic transformer design principles.

In this section the basic design principles will not be rigorously analysed, but the more salient items of design explained.

In modern radio practice much emphasis is laid upon the fidelity of reproduction, and it is desirable that the low frequency amplifier should have an even response from 30 to 15 000 cycles per sec., although 50 to 10 000 cycles is considered very good. Thus it is necessary, if such response is to be achieved, that the intervalve and output transformers shall also have such a response. Now the frequency response can be predicted quite accurately by calculation, and this will now be considered.

The diagrams of Fig. 10 are the equivalent circuits of a transformer working between load resistances, the actual circuit being shown at (*a*), whilst the equivalent circuits, where the valve is replaced by a generator acting in series with the plate resistance, are shown at (*b*) and (*c*). A push-pull stage shown in Fig. 11 (*a*) has an equivalent circuit as in Fig. 11 (*b*) from which the likeness to Fig. 10 (*a*) will be observed, the principal difference being that the plate resistance is twice that of a single valve, in the case of triodes, but is usually the same in the case of pentodes. It therefore follows that a separate analysis for push-pull operation is not required. In the equivalent circuits the transformer is taken into account by the primary inductance L_p, the leakage inductances L_1 and L_2, the primary and secondary d.c. resistances, and the transformation ratio η. For a given voltage developed across the primary at a given frequency the flux density of the core must not

* Vol. II, 1924, Nos. 12, 13, and 14.

be allowed to exceed a certain figure if amplitude distortion is to be avoided, and this is a function of the quantity of iron and number of turns. Expressed in another way, there is a minimum quantity of iron and number of turns to be used at

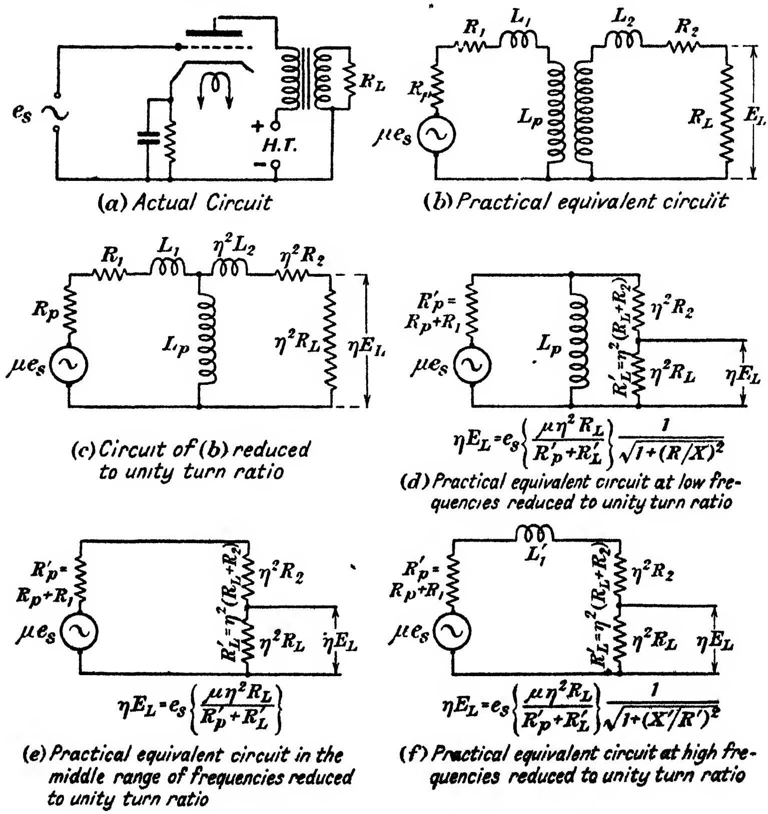
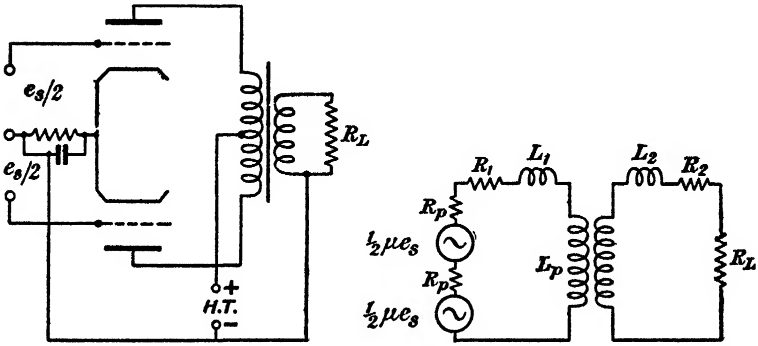


Fig. 10. Practical and Equivalent Transformer Circuit Diagrams at Audio-frequencies

a given frequency and voltage if distortion is to be avoided. These items are rigorously analysed in Dr. Dye's paper, whilst here it is proposed to give some practical examples of successful designs and omit the basic magnetic calculations. The equivalent circuit of Fig. 10 (c) may be simplified by considering only a limited frequency range. Fig. 10 (d) is applicable at low frequencies where the leakage reactance has a negligible effect, and Fig. 10 (*f*) is applicable where the reactance of the primary inductance at high frequencies has a negligible shunting effect. Between these extreme cases the frequency is low enough for the leakage inductance to have but small effect, and high enough so that the shunting action of the primary inductance can be ignored, resulting in the circuit of Fig. 10 (*e*).

In an article in *Electronics*,* Professor F. E. Terman described how the frequency response of a transformer may be computed by the manipulation of the equivalent circuits of Fig. 10 (*d*),



Push-pull amplifier (*a*) Practical equivalent circuit (*b*)

FIG. 11. PUSH-PULL AMPLIFIER

(*e*), and (*f*), and his methods will now be summarized. Taking initially the middle frequencies, the equation attached to Fig. 10 (*e*) gives the output voltage to be expected, whilst for the low frequencies the equation of Fig. 10 (*d*) shows that the response is the output of Fig. 10 (*e*), multiplied by the factor

$$1/\sqrt{[1 + (R + X)^2]},$$

where R is the resistance formed by the effective plate and load resistance, R_p' and R_L' in parallel, whilst X is the reactance of the primary inductance.

Similarly, the response at high frequencies is also a function of Fig. 10 (*c*), multiplied by the factor

$$1/\sqrt{[1 + (X'/R')^2]},$$

where X' is the reactance of the leakage inductance and R' is

* January, 1936.

the sum of R_p' and R_L' representing the effective plate and plate load resistances.

According to Professor Terman, results obtained in this way are accurate to a few per cent. The Author, in checking the calculated against measured response found this error does not exceed 6 per cent for a well-designed transformer.

With this method it is necessary to know the d.c. resistance of the primary and secondary windings, and the primary and leakage inductances, also the turns or voltage ratio. Then, knowing the plate and load resistances, an accurate frequency response curve can easily be calculated.

The leakage inductance reduced to unity turn ratio (1 : 1) is the inductance measured across the primary with the secondary short-circuited, and the primary inductance is similarly obtained but with the secondary open-circuited. These measurements are usually made on an ordinary bridge at 1 000 cycles. If a d.c. component flows through the primary winding under operating conditions, this should also be used in measurements.

It is now possible to make a summary of requirements to produce a really good transformer having so-called *high fidelity* characteristics.

A high primary and low leakage inductance is the main requirement, and it should be noted that the leakage co-efficient L_1'/L_p determines the maximum to minimum frequency ratio that can be covered with a given load to plate resistance. The leakage co-efficient depends upon transformer proportions and design rather than size; and the most suitable ways of reducing it to a low value are by interleaving the primary and secondary windings and properly proportioning the winding space; by the elimination of the d.c. component in the primary winding by the use of either a parallel feed arrangement or push-pull operation, and by the use of a core material having a high permeability such as Mumetal, etc. Although replacing normal iron with a high permeability iron greatly increases the primary inductance, it does not alter the leakage inductance, due to the reluctance of the leakage paths being largely in the winding space. It does, however, reduce the leakage factor L_1'/L_p, and therefore materially improves the transformer response. By using Table XXIV in conjunction with the curves of Figs. 12A and 12B, the response characteristic calculations are materially simplified. It will be noticed that by making the load resistance greater than the conventional twice the plate resistance, the high frequency response may be somewhat improved without materially affecting the low frequency response. Professor Terman explains this as shown on next page, Table XXIV.

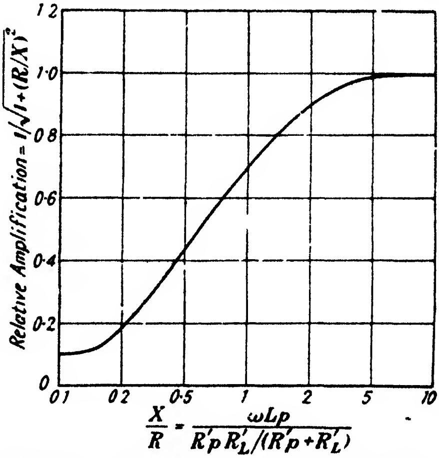


Fig. 12a. Factor for Low Frequencies

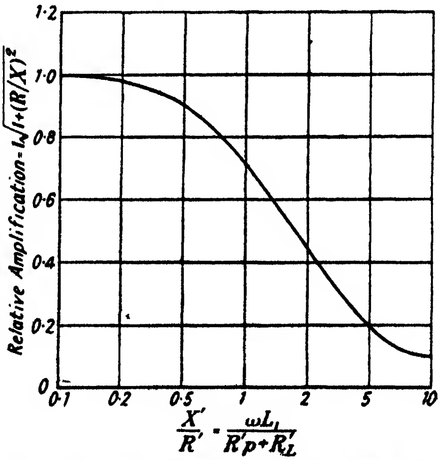


Fig. 12b. Factor for High Frequencies

The number of octaves which a transformer can cover is determined by the ratio of leakage inductance to primary inductance, and is increased somewhat by making the load resistance considerably greater than twice the plate resistance.

TABLE XXIV

RATIO OF MAXIMUM TO MINIMUM FREQUENCY FOR 3 DB. LOSS PLOTTED AGAINST LEAKAGE CO-EFFICIENT

$$\frac{\text{Maximum efficiency}}{\text{Minimum efficiency}} = \frac{(1 + R_L'/R_p')^2}{(R_L'/R_p')} \cdot \frac{L_p}{L_1'}$$

Leakage Co-efficient L_1'/L_p	R_L'/R'_p				
	1	2	4	6	8
0·001	4 000	4 500	6 000	8 000	10 000
0·002	2 000	2 250	3 000	4 000	5 000
0·003	1 333	1 500	2 000	2 666	3 333
0·004	1 000	1 125	1 500	2 000	2 500
0·005	800	900	1 200	1 600	2 000
0·006	666	750	1 000	1 333	1 666
0·007	571	643	857	1 143	1 428
0·008	500	562	750	1 000	1 250
0·009	444	500	744	888	1 111
0·01	400	450	600	800	1 000
0·02	200	225	300	400	500
0·03	133	150	200	266	333
0·04	100	112	150	200	250
0·05	80	90	120	160	200
0·06	66	75	100	133	166
0·07	55	64	85	114	143
0·08	50	56	75	100	125
0·09	44	50	66	77	111
0·1	40	45	60	80	100

An interesting example of a high grade audio-frequency transformer is the "Ferranti A.F.5." (See Fig. 13.) The

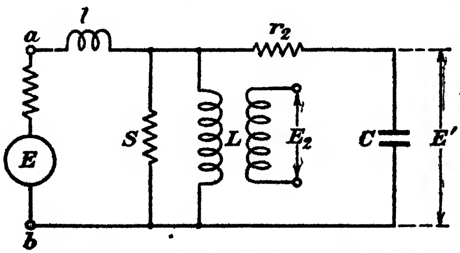


FIG. 13. SIMPLIFIED EQUIVALENT CIRCUIT OF FERRANTI A.F.5. TRANSFORMER

primary inductance is 100 henries, turns ratio 3·5 : 1, leakage inductance 0·8 henry, whilst the self-capacitance is approximately 0·00062 μF.; thus the leakage factor = 0·8/100 = 0·008.

At 50 cycles the impedance of the leakage inductance is negligible if an input resistance of 10 000 ohms is assumed. The main effect is that due to the shunting of L, whose impedance is $j31\ 400$ ohms. Then E' is given by

$$E' = \frac{j\omega L}{R_1 + j\omega L} E = \frac{j31\ 400}{10\ 000 + j31\ 400} E$$

$$= \frac{31\ 400 \underline{|90^\circ}}{33\ 000 \underline{|72{\cdot}5^\circ}}$$

$$= 0{\cdot}95 \underline{|17{\cdot}5} E \quad . \quad . \quad . \quad . \quad . \quad (74)$$

so that

$$E_2 = 3{\cdot}33 \underline{|17{\cdot}5^\circ} E \quad . \quad . \quad . \quad . \quad (75)$$

At a frequency of 640 cycles L and C resonate: thus the impedance of the LC, r_2 circuit is very high and the transformer impedance is then S, giving

$$E' = E \frac{S}{R_1 + S}$$

$$= E \, . \, (500\ 000/513\ 000)$$

$$= 0{\cdot}973E$$

and

$$E_2 = 3{\cdot}41 \underline{|0^\circ} E \quad . \quad . \quad . \quad . \quad . \quad (76)$$

At the higher frequencies the shunting effect of S and L becomes negligible, and the loss is due to the series drop l, so that

$$E' = \frac{E(1/j\omega C)}{r_2 + 1/j\omega C + j\omega L + R_1}$$

$$\doteq \frac{E}{1 - \omega^2 lC + j\omega C(R_1 + r_2)} \quad . \quad . \quad (77)$$

The resonant frequency of l and C is important and is 7 200 cycles. At this frequency

$$E' = E/j\omega C(R_1 + r_2)$$

$$= E\overline{|90^\circ}\ (35\ 900)/(R_1 + r_2) \quad . \quad . \quad (78)$$

The value of E' is dependent upon the value of the input and secondary resistances. If $R_1 = 10\ 000$ ohms and r_2 is small, $E' = 3{\cdot}59 \overline{|90^\circ} E$, and $E_2 = 12{\cdot}5 \overline{|90^\circ} E$, which is too high. This resonance is damped by using a higher input resistance and connecting a resistance across the secondary. If an additional resistance of 10 000 ohms is put in the primary, the value E_2 becomes $6{\cdot}2\,\underline{|90^\circ}E$. By placing 0·25 MΩ. across the secondary, it is equivalent to a resistance $250\,000/3{\cdot}5^2 = 20\,400$ ohms across C, or approximately 16 000 ohms in series with C, and then $E' \doteqdot 1{\cdot}35E$. If the extra 10 000 ohms is placed in series and the 250 000 ohms across the secondary, $E' \doteqdot E$ to a close approximation.

The reader will find a full discussion of equivalent transformer circuits in Dr. Starr's *Electric Circuits and Wave Filters**, from which this summary of the "A.F.5" transformer is given.

PRACTICAL TRANSFORMER DESIGN

The preceding part of this section dealt with the more technical aspects of the subject, so it is now opportune to consider practical design problems for manufacturing purposes.

Intervalve Transformers. Since intervalve transformers usually operate between two high impedances, it is very necessary that the distributed capacitance and leakage reactance should be reduced as much as possible, whilst the primary inductance should be as high as practicable. The usual procedure is to select a suitable core stamping and design the transformer simply on a voltage ratio basis. If no d.c. flows in the windings a small wire, such as 42 S.W.G. is used, and as many turns placed on the primary as the window space will permit after allowing for the space occupied by the secondary winding. If the same gauge of wire is used for both primary and secondary windings, the following formula gives to a close approximation the maximum number of turns of a given size of wire that will fit in a given space—

$$N_{max} = \left[\frac{0{\cdot}87(L - 2M)}{D}\right]\left[\frac{(W - Q) \times 0{\cdot}9}{D + 2T}\right] \quad . \quad (79)$$

or if the desired number of turns is known the wire size may be determined thus

Let $L - M = A$

and $W - Q = B$.

Then $$D = T \pm \sqrt{[T^2 + 0{\cdot}78(AB/N)]} \quad . \quad . \quad (80)$$

* Second Edition (Pitman).

where L = window length;
W = window width;
M = paper margin;
T = paper thickness.
D = diameter of insulated wire;
Q = thickness allowed for former and interwinding insulations;
N = number of turns.

(in inches or centimetres)

A sample transformer should be wound to the specification prepared from the preceding data, and should be checked for voltage ratio and frequency response. If the transformer is for push-pull operation, the secondary balance and phase relationships should also be checked. When there is a d.c. component in the primary circuit, allow 600 to 1 200 circular mils per ampere in determining the wire gauge, and when initially testing, the air gap should be adjusted to optimum position. If these initial tests show the voltage ratio to be correct, it is quite likely the frequency response, especially at high frequencies, will not be satisfactory. This must be dealt with later. If the voltage ratio is not correct, adjustments should be made to the secondary windings only, provided the primary inductance is of sufficiently high value. A good practical value is about 70–100 henries, whilst 150–200 henries is extremely good.

Now to deal with frequency response. Poor low note response means insufficient primary inductance, whilst poor high note response means that the leakage inductance is too great. The cure for the first case is obvious, but the second needs some attention to detail. The first step is to sectionalize the primary winding, and initially this may be in two sections with the secondary sandwiched between. Further frequency response checks should now be made, and it is almost certain that the high note response will have considerably improved when compared with the first sample. Further sectionalizing usually improves matters still more, but a practical limit is usually reached when the primary is in four sections and the secondary in three sections between the primary windings. The primary and leakage inductances may be checked at each stage in the manner described in the previous section, and the decrease of leakage inductance is an index to the "goodness" of the transformer, at high frequencies. It should also be noted that sectionalizing the windings decreases the self-capacitance, which is very desirable.

Output Transformers. The remarks made in reference to primary inductance and leakage reactance in connection with intervalve transformers are equally applicable to output transformers. In general, if output transformers are made with wire sufficiently large for high efficiency operation, no air gap is needed for push-pull operation. For single valve output stages an air gap is always required. The wire chosen for the primary winding should give a current density between 600 and 1 000 circular mils per ampere, whilst the secondary wire should be such that the d.c. resistance of the secondary winding does not exceed 5 per cent of the d.c. resistance of the voice coil of the loudspeaker with which it is to be used. In practice it is often found that this figure is nearer 10 per cent. During the preliminary stages of design, the d.c. resistance of the primary winding, and also the core losses, are not very accurately known. Under these circumstances the minimum inductance may be computed thus. (See Fig. 14.)

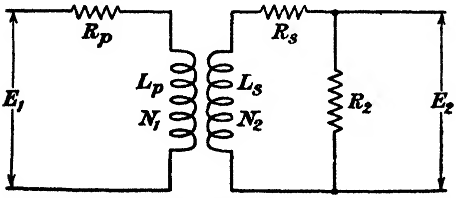


FIG. 14. BASIS FOR PRELIMINARY CALCULATIONS IN OUTPUT TRANSFORMER DESIGN

$$L_p = \frac{1}{\omega\sqrt{[(1/K^2) - 1]}}\left[\frac{R_p(R_s + R_2)}{a^2R_p + R_s + R_2}\right] \quad . \quad (81)$$

where L_p = primary inductance in henries;

$\omega = 2\pi f$ (f being the lowest desired frequency);

R_p = plate impedance of output valve, R_s = d.c. resistance of transformer secondary, R_2 = impedance of speech coil of loudspeaker } in ohms;

N_1 = primary turns;

N_2 = secondary turns;

$a = N_2/N_1$;

K = ratio (expressed as a decimal) of the response at the lowest frequency of interest to the highest frequency.

The ratio E_2/E_1 at high frequencies can be shown to be

$$\frac{E_2}{E_1} = B = \frac{-j\omega M R_2}{(a^2 R_{pL} + R_s + R_2) j\omega L_p} = \frac{-aR_2}{a^2 R_{pL} + R_s + R_2} \quad (82)$$

where R_{pL} is the optimum load of the output valve.

Taking $\partial B/\partial a$ of the former equation and equating to zero gives—

$$a^2 \frac{R_s + R_2}{R_{pL}} \text{ or } a = \sqrt{\frac{R_s + R_2}{R_{pL}}} \quad . \quad : \quad (83)$$

for the most efficient impedance match at this stage.

Most commercially produced output transformers use this as the basis of design, but as shown in the preceding section, the d.c. resistance of the primary winding should be taken

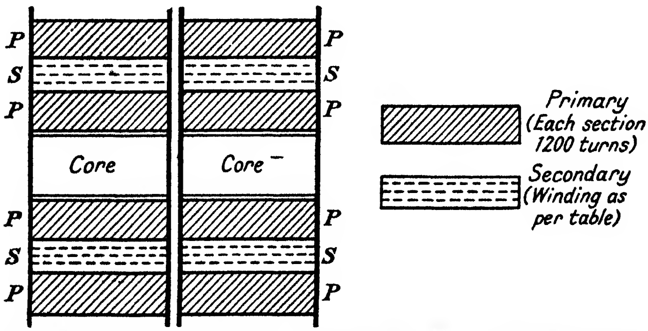


FIG. 15. WINDING SPACE ALLOCATION FOR MULTI-RANGE OUTPUT TRANSFORMER

into account for the most accurate results. This can well be seen in a case where the optimum load of a valve may be about 800 ohms, and the d.c. resistance of the primary may be about 200 ohms. Such would introduce an appreciable error in impedance matching. It is safe to assume that the primary d.c. resistance can only justifiably be ignored when it does not exceed 5 per cent of the optimum load resistance of the output valve, but in the case of pentode or class B output stages, due to the critical matching requirements, it should always be taken into account.

The *Wireless World** gives details of a push-pull, multi-ratio output transformer for use with the "Monodial" receiver, which has two PX4 valves in the output stage. The primary consists

* 8th September, 1933.

of 4 800 turns of 36 S.W.G. wound in four sections of 1 200 turns on two half-size No. 4 formers. The secondary windings, details of which are given in Table XXV, are sandwiched between each quarter primary on each former as shown in Fig. 15. The core consists of 100 pairs No. 4 Stalloy T. & U. stampings (Sankey), and the primary inductance is 69 henries. Since the transformer is of liberal design, no air gap is required. The total primary resistance is 300 ohms, and the frequency response is level within 2 db. from 50–8 000 cycles.

TABLE XXV

OUTPUT TRANSFORMER WINDINGS (SECONDARY) FOR *Wireless World* PUSH-PULL DESIGN

Loudspeaker	Speech Coil Impedance (ohms)	Ratio	Secondary Turns per Bobbin	Wire	Connection of Two Secondary Coils
Magnavox Dual .	0·8	112–1	42	18 d.c.c.	Parallel
Rola Dual . .	1·0	100–1	48	18 ,,	,,
Magnavox Single .	1·6	79–1	62	20 ,,	,,
Rola Single . .	2	70·7–1	68	20 ,,	,,
B.T.H. R.K. Senior & Ferranti M.2 & D.3	15	25·8–1	93	20 d.s.c.	Series
Ferranti M.1 . .	20	22·3–1	108	22 ,,	,,

An improvement in the upper register may be effected by omitting the multi-ratio characteristic, and using only the desired single secondary winding to give correct impedance matching. Further improvement may then be made by further subdividing the primary and secondary with appropriate sandwiching of the secondary winding.

Some experiments undertaken by the Author using the *Wireless World* design as a basis showed that dividing the primary into eight parts and the secondary into six parts resulted in the upper register being extended to 11 400 cyc. with but 2 db. deviation. Replacing the Stalloy core by an equivalent size of Mumetal (an expensive job) increased the primary inductance enormously, and resulted in but a 2 db. deviation at 20 cyc. This modified design was eventually used in a beat frequency oscillator, which gave a level response within 2 db. between 20 and 11 000 cyc.

A smaller output transformer to handle 3 watts a.c. and carry up to 60 mA., d.c. in the primary, such as may be required in a single-valve output stage, may be constructed by using a 3 000 turn primary (two halves of 1 500 turns each) of 34 S.W.G. enamelled wire wound on a No. 30 former: this will give a d.c. resistance of a fraction over 200 ohms. The secondary winding is sandwiched between the two half-primaries, and details will be found in Table XXVI. The core consists of 100 pairs No. 30 Stalloy T. & U. stampings (Sankey), and an air gap of 0·005 in. should be allowed at each butt joint. The electrical details are as follows—

Frequency response .	Level from 50 to 3 000 cyc.
	1 db. down at 4 000 cyc.
	2 db. ,, ,, 8 000 cyc.
Primary inductance .	25 henries with no d.c.
	22·5 ,, ,, 30 mA.
	20 ,, ,, 50 ,,
	18 ,, ,, 60 ,,

As in the previous example, improved high-note response is obtained by further sectionalizing.

TABLE XXVI

SECONDARY WINDINGS FOR 3-WATT A.C. SINGLE RATIO OUTPUT TRANSFORMER

Output Value Optimum Load (watts)	Speech Coil Impedance	Transformer Ratio	Secondary Turns	Wire Enamelled (S.W.G.)
2 000	2·0	1 : 31·6	95	20
2 000	7·5	1 : 16·3	184	22
3 000	15·0	1 : 11·5	261	24
3 000	2·0	1 : 38·7	72	20
3 000	7·5	1 : 20·0	150	22
3 000	15·0	1 : 14·1	212	24
4 000	2·0	1 : 44·8	67	20
4 000	7·5	1 : 23·1	130	22
4 000	15·0	1 : 16·3	184	24
5 000	2·0	1 : 50·0	60	20
5 000	7·5	1 : 25·8	116	22
5 000	15·0	1 : 18·2	165	24

Power Type Transformers. Viewed from purely an electrical power engineering aspect, a great number of power transformers used in current radio receiver practice are very poor in efficiency and design. The power engineer's axiom is "to get as much out as is put in," and we find small power transformers designed with this idea having efficiencies anywhere between 90 per cent and 98 per cent. Contrasting with this we find many transformers designed for purely radio purposes having efficiencies much nearer 60 per cent. It therefore appears desirable to

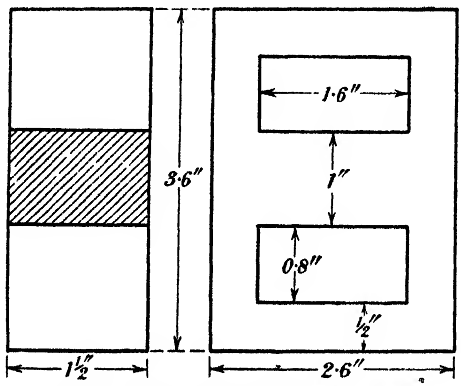


FIG. 16. STAMPING FOR TRANSFORMER
Stalloy: 0·014 in. thick

review small power transformer design from the electrical power engineer's viewpoint in the first instance, and then to proceed to the consideration of the technique of mass-produced "radio" transformers.

The basis of design will be a small power transformer which the Author has found particularly useful: its specification is given herewith.

Primary	240 V., 50 cyc.
Secondary, h.t. . . .	250–0–250 V., 100 mA. rectified a.c.
,, l.t. . . .	2–0–2 V., 2 A.
,, rectifier filament	2–0–2 V., 2 A.

From this specification it will be realized that each plate of the full-wave rectifier must supply 50 mA. The transformer is wound on a core of the dimensions given in Fig. 16, this being a standard Stalloy stamping regularly used by the Author. The fundamental formula for design is

$$\frac{1}{\text{turns per volt}} = 4{\cdot}44 \times 10^{-8} f \times \text{core section} \times \text{dux density} \qquad . \quad . \quad . \quad . \quad . \quad (84)$$

Since the core section is 1·5 in.² and 0·014 in. Stalloy is operated at 50 000 lines per in.², we find the turns per volt to be 6; thus we have

Primary	240 × 6	=	1 440 turns
Secondary, h.t. . . .	250 × 6 × 2	=	3 000 ,,
,, l.t. . . .	4 × 6	=	24 ,,
,, rectifier filament	4 × 6	=	24 ,,

TABLE XXVII

TURNS PER VOLT IN A WELL-DESIGNED TRANSFORMER HAVING CORE DENSITY OF 50 000 LINES PER IN.²

Cross-section of Core (in.²)	Turns per Volt		
	25 cyc.	50 cyc.	60 cyc.
1	18	9	7·5
1¼	14·5	7·25	6
1½	12	6	5
1¾	10·20	5·10	4·3
2	9	4·5	3·9
2½	7·2	3·6	3
3	6	3	2·6
3½	5·2	2·6	2·2
4	4·5	2·25	2·0

Table XXVII gives the requisite turns per volt for various core cross-sectional areas and frequencies when Stalloy is used at a flux density of 50 000 lines per in.². It is now necessary to determine the primary current when the transformer is fully burdened, also to make an estimate of the efficiency of the computed design. This estimate should always be conservative, and for the present purpose we will estimate it at 80 per cent, although it will be seen later the overall efficiency is considerably higher. The total secondary load is thus—

$$
\begin{aligned}
250\text{ V.} \times 100\text{ mA.} &= 25\text{ W.}\\
4\text{ V.} \times 2\text{ A.} &= 8\text{ W.}\\
4\text{ V} \times 2{\cdot}5\text{ A.} &= 10\text{ W.}\\
\hline
&\ 43\text{ W.}
\end{aligned}
$$

Thus at 80 per cent efficiency the primary would have to supply 43 × 100/80 = 54 W. at 240 V. = 225 mA.

Knowing the currents in each winding the correct wire gauge is chosen, the current density in each case being taken as 1 200 A. per in.^{2}. A standard wire table (as, for example, London Electric Wire Co. & Smiths Ltd.) shows the requisite gauges to be as follows—

Primary	. . . . .	28 S.W.G	
Secondary, h.t.	. . .	36 ,,	
,, l.t.	. . .	18 ,,	(See Note)
,, rectifier filament	.	16 ,,	

Note. Since only 24 turns are required for each 4-volt winding, it will be more convenient to use 16 S.W.G. in each case.

In a new design it is also necessary to determine the winding area available. This may easily be found by simple arithmetic when the size of the stampings is known. Table XXIX gives the number of turns per square inch for common gauges of wire, but allowance for insulation must always be made in these calculations. In the design now being considered, we have determined the turns required for each winding and the gauge thereof; thus we have now to find the loss due to the resistance of windings. This is commonly known as *copper loss.* One quantity still to be determined is the mean length of each turn. In the present case this can be shown to be 2(1·8 + 2·3) = 8·2 in. —the dimensions being obtained from the size of the stampings. Thus the copper loss for each winding is given in Table XXVIII, from which it will be seen the total copper loss is 3·66 watts.

TABLE XXVIII

SUMMARY OF CALCULATED DATA FOR SMALL POWER TRANSFORMER

Winding	Mean Turn Length (in.)	No. of Turns	Total Length (yd.)	S.W.G. Wire	Resistance (Ω.)	Current (A.)	Loss (W.)
Primary . .	8·2	1 440	328	28	46	0·225	2·34
Secondary—							
H.T. . .	8·2	3 000	684	36	361	0·05	0·9
L.T. . .	8·2	24	5·47	16	0·0408	2	0·1632
Rect. Fil. .	8·2	24	5·47	16	0·0408	2·5	0·255
						TOTAL	3·6582

TABLE XXIX

MAXIMUM TURNS IN IN.² WINDING AREA ALLOWING FOR ADEQUATE INSULATION

Turns per In.²					Wire	
Enamel	S.C.C.	D.C.C.	S.S.C.	D.S.C.	Diameter (in.)	S.W.G
—	54·1	49·6	58·3	57·0	0·128	10
—	79·7	71·8	87·3	85·0	0·104	12
—	129	113	145	140	0·080	14
226	198	173	223	213	0·064	16
392	343	297	400	377	0·048	18
685	567	472	692	641	0·036	20
1 110	865	692	1 110	1 010	0·028	22
1 770	1 280	977	1 770	1 600	0·022	24
2 560	1 740	1 280	2 560	2 270	0·018	26
3 760	2 310	1 630	3 650	3 160	0·0148	28
5 370	2 950	1 990	5 180	4 500	0·0124	30
6 890	4 010	2 550	6 610	5 650	0·0108	32
9 610	4 960	3 020	8 730	7 310	0·0092	34
13 500	7 430	4 110	12 100	10 300	0·0076	36
20 400	10 000	5 100	17 800	14 700	0·006	38
32 500	12 900	6 100	25 200	20 100	0·0048	40
44 300	—	—	36 300	27 800	0·0040	42
64 100	—	—	50 500	37 000	0·0032	44
—	—	—	75 100	51 600	0·0024	46

The iron loss is determined from a knowledge of the volume of iron in the transformer and its quality. As one side of the stampings is coated with an insulating substance, the actual volume should only be taken as 90 per cent of the physical volume; thus we have

$$1{\cdot}5 \times 0{\cdot}9\,(2{\cdot}6 \times 3{\cdot}6 - 2 \times 1{\cdot}6 \times 0{\cdot}8) = 9{\cdot}15\ \text{in.}^3$$

The expression for iron loss is watts per lb. × volume of iron × density of iron.

The iron loss of Stalloy at 50 000 lines per in.² is 0·39 W. per lb., and the density of the iron is 0·28. Thus we have a total iron loss of 1 watt. It is now possible to make a summary of the transformer.

Total secondary burden	=	43 W.
Copper loss	=	3·66 W.
Iron loss	=	1·0 W.
		47·66 W.

which must be supplied by the primary. It will be noted that this shows a much higher efficiency than the estimate of 80 per cent. Since the primary is not required to supply the current estimated for, it is obvious this will cause a further decrease in copper loss, and reverting to the formula for calculating this loss, it will be seen this now amounts to 1·7 watts for the primary as against 2·34 watts in the original estimate. Thus the total copper loss is now 3·02 watts plus iron loss of 1 watt = 4·02 watts. The primary is therefore required to supply 43 + 4·02 watts = 47 watts to a close approximation. The efficiency of the transformer is therefore

$$43/47 \times 100 = 91{\cdot}5 \text{ per cent,}$$

which may be considered quite satisfactory. This transformer operates with but the slightest temperature rise after a continuous operating period of 24 hr., and after three months' continuous operation is still quite cool. Electrical power engineers may feel disposed to be critical of the copper loss being greater than the iron loss, as they prefer to balance both losses. This could be done by the use of thicker gauge wire in the primary, but would present some difficulties with the existing stampings as the winding area is fairly full already when liberal insulation is used between windings. The transformer shows infinite insulation resistance between windings and from windings to core at 10 000 volts alternating.

MASS PRODUCTION TYPES. It has never been the Author's lot to design such transformers: for the information in this subsection he is therefore indebted to a number of designing engineers engaged on such work, supplemented by careful examination and measurements he himself has made of a number of typical transformers. The items to be considered in such designs are set out below, placed in order of importance.

1. Manufacturing cost.
2. Size.
3. Operating characteristics—

   (*a*) Regulation.
   (*b*) Temperature.
   (*c*) Operating life.
   (*d*) Freedom from mechanical noise.

To meet the conditions of Item 1, a minimum quantity of copper is used, whilst the iron is cut to as fine a limit as possible. Enamelled wire is used throughout, and each layer of wire is insulated from the next by thin Kraft paper or Glassine material. Under these circumstances it is important to know the maximum permissible current and flux densities. These items are given in Tables XXX and XXXI. It is very likely that two or three standard stampings are available; thus it is necessary to determine the size of core. A working formula for this is

$$A = c\sqrt{(E_l)} \qquad \qquad (85)$$

where A = cross-section of core, in.^{2};
c = empirical design constant;
E_l = power output of transformer in watts.

TABLE XXX

MAXIMUM PERMISSIBLE CURRENT DENSITIES
in Circular Mils per Ampere

Winding	Current Density	
	Average	Maximum
Filaments . . .	600	450
H.t. Windings (secondary)	600	500
Primary . . .	600	500

The quantity *c* is usually expressed thus—

50 cyc. 0·18 to 0·28 } The lowest figure corresponds to the
25 ,, 0·28 to 0·38 } highest flux density.

TABLE XXXI

AVERAGE VALUES OF FLUX DENSITY
Measured on Ten Transformers

Rated Output (watts)	Flux Density (Lines per in.²)	
	25 cyc.	50 cyc.
20	105 000	75 000
30	97 000	70 000
40	94 000	69 000
50	92 000	67 000
60	90 000	66 000
70	88 000	65 500
80	86 000	65 000
90	85 000	64 000
100	84 000	63 500

The turns per volt may be determined from Table XXVII, but as this is for a flux density of 50 000 lines per in.² only, it will be necessary on most occasions to determine this item from the formula

$$N_p = \frac{E_p}{4{\cdot}44 \times f \times \phi \times A} 10^8 \quad . \quad . \quad (86)$$

where E_p = primary voltage;
f = supply frequency;
ϕ = flux density in lines per in.²;
A = cross-sectional area of core, in.².

The various secondary windings are obtained by the voltage ratios of the primary to secondary winding, and about 5 per cent is added to allow for the voltage drop in the windings.

In a mass production factory automatic coil winding machines are available; thus it is possible to determine a winding spacing factor. As enamelled wires are used it is usual not to wind each turn tightly against its neighbour. In the primary and h.t. secondary windings the winding factor is usually 85 to 87 per cent, which means that each turn is just slightly separated from its neighbour. Filament windings are usually wound with a

93 to 95 per cent winding factor. A certain amount of consideration is given to the potential difference between adjacent layers, and Table XXXII gives general values for different thicknesses of wire in r.m.s. volts per layer. Also, in connection with insulation, general practice appears to allow $\frac{1}{4}$ in. paper margin for primary windings, and about $\frac{1}{4}$ in.–$\frac{5}{16}$ in. for h.t. secondaries of the 1 000-volt type, whilst 2 000 volt secondaries have $\frac{5}{16}$ in.–$\frac{3}{8}$ in. margins. Filament windings have a paper margin of $\frac{1}{4}$ in.–$\frac{3}{8}$ in. A temperature rise of 40° C. appears to be quite common, and an electrical efficiency of 75 to 80 per cent was the best figure which could be obtained from the samples examined. It is desired to emphasize that generated heat is absolutely wasted watts.

TABLE XXXII

RECOMMENDED MAXIMUM R.M.S. VOLTS PER WINDING LAYER

Paper Thickness (in.)		Max. Size of Wire (S.W.G.)	R.M.S. Volts per Layer	
			Double Paper	Single Paper
Kraft paper	0·002	26	150	75
	0·003	24	200	100
	0·005	17	350	175
	0·005	14	Filament	Filament
Tissue paper	0·0005	40	50	25
	0·001	38	75	37·5
	0·0015	31	140	70

Intermediate Frequency Transformers for Supersonic Heterodyne Receivers. The normal intermediate frequency (i.f.) transformer usually consists of two coils tuned by two air or mica dielectric condensers. The goodness factor of a coil-condenser combination is known as the Q of the circuit, and is equal to the ratio of the inductive or capacitative reactance to the resistance. Thus

$$Q = \omega L/R = 1/\omega CR. \qquad (87)$$

A simple manner for the determination of the Q of a coil is to time it to resonance by means of a condenser possessing negligible losses, and shunt a valve voltmeter having a high effective input resistance across the coil. The capacitance is changed to values above and below resonance until the valve voltmeter reads 0·71 of the resonant value, when the Q of the coil is

$$Q = 2C_0/\Delta C \quad . \quad . \quad . \quad . \quad . \quad . \quad (88)$$

where C_0 = total capacitance at resonance;

ΔC = total change between $\pm$ 0·71 from resonance.

Intermediate frequency transformers fall into two broad sections: first, those between 110 and 175 kc.; and second, those between 430 and 470 kc. In both classes, multi-layer self-supporting coils are used, and these are sometimes subdivided into "pies." Several narrow "pies" give a higher Q at the high frequency end, but very often space requirements determine the form of winding which can be used, and thus limit the Q. All forms of coils suffer greatly from losses due to the proximity of the shield can. Table XXXIII illustrates this effect from some measured values. The tuning condenser should have as high a Q as is reasonably possible, or to use a more general expression, it should have negligible loss at high frequencies. The Section on "Condensers" should be referred to for the formula giving the figure of merit of condensers. The field linkage between the two coils is both electromagnetic and electrostatic.

TABLE XXXIII

SHIELD LOSSES IN I.F. TRANSFORMERS

	Type of Winding	Wire	Diameter of Core (in.)	Inductance (mH.)	Q Shield: 1 3/8 in. square	Q Shield: 1 7/8 in. circular
Air Core	1 Pie	40 S.W.G. S.S.C.	1/2	1·24	46	—
	1 „	3/40 Litz	3/8	1·15	75	—
	3 „	7/41 „	1/2	1·7	94	103
	3 „	7/41 „	3/8	1·8	87	108
	4 „	7/41 „	1/2	1·5	99	106
Iron Dust Core	1 „	7/41 „	3/8	1·5	136	—

By winding the coils in the same direction and connecting either the starting or finishing leads to the grid and plate, the magnetic coupling is made to oppose the capacitative coupling.

Reversing the connections or direction of winding of either coil causes these couplings to become additive. Commercial practice is to keep the capacitance coupling low in value to make coil spacing non-critical. Fig. 17 shows a typical i.f. amplifier circuit, from which it will be seen that high-Q anti-resonant circuits, formed by inductance and capacitance in parallel, result in a high impedance at the desired frequency. This high

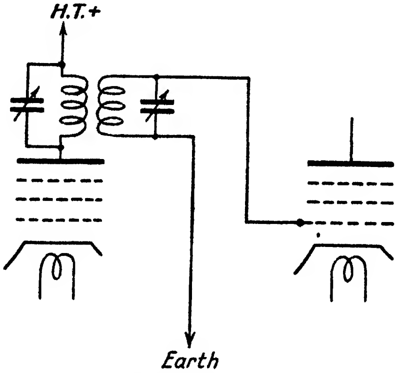


FIG. 17. TYPICAL I.F. AMPLIFIER CIRCUIT

impedance in the plate circuit of a pentode results in high voltage amplification. If only the plate circuit is tuned, an approximate formula for the stage gain is

$$\text{Gain} = G_m \left[\frac{1.}{1/r_p + 1/Q\omega L}\right] \quad . \quad . \quad (89)$$

where r_p = plate resistance of the valve.

And when both coils are tuned the maximum gain is approximately given by

$$\text{Maximum gain} = \frac{G_m}{2}\sqrt{\left[\frac{1}{1/r_p + 1/Q_1\omega L}\sqrt{(Q_2\omega L_2)}\right]} \quad . \quad (90)$$

The selectivity of a single tuned circuit can be approximately given by Eq. (91)

$$\Delta f \doteq (f_r/Q)\sqrt{[(E_r/E) - 1]} \quad . \quad . \quad . \quad (91)$$

and when loaded by the plate resistance of the valve becomes

$$\Delta f = f_r(1/Q + \omega L/r_p)\sqrt{[(E_r/E)^2 - 1]} \quad . \quad (92)$$

where Δf = band width corresponding to the ratio of resonant voltage, E_r to off-resonant voltage E;

f_r = resonant frequency.

The selectivity of two loosely coupled tuned circuits is given by Eq. (93).

$$\Delta f = \frac{f_r}{\sqrt{(Q_1 Q_2)}}\sqrt{\left[\frac{E_r}{E} - 1\right]} \quad . \quad . \quad (93)$$

If the circuits have critical coupling and the primary is loaded by the plate resistance of a valve, the selectivity is given by

$$\Delta f = \frac{f_r}{\sqrt{Q_2}}\sqrt{\left[2\left(\frac{1}{Q_1} + \frac{\omega L}{r_p}\right)\right]}\sqrt{\left[\frac{E_r}{E} - 1\right]} . \quad . \quad (94)$$

Thus band width is seen to vary inversely with the Q of the circuit, and directly with resonant frequency. From this it is easy to see that the coils and condensers should have as high a Q value as possible to attain maximum gain and selectivity from the i.f. amplifier.

In the so-called *high fidelity* receivers, it is usual to find the coupling between the two coils is variable by mechanical means, thus enabling different degrees of selectivity and band width to be obtained.

The inductances may be calculated from data given in the section on "Inductances," whilst the mutual inductance may be calculated from the details given in the same section.

SECTION V

AUDIO-FREQUENCY SOURCES

ONE of the principal requirements of an a.f. oscillator is purity of waveform; thus simple feed-back oscillators utilizing a single valve are generally unsuitable. A.f. oscillators can be conveniently divided into two classes, viz. the *tuned circuit* type, and the *heterodyne beat* type. The former is more easily constructed, but is not so convenient when a continuously variable frequency range is desired, this being the main advantage of the heterodyne beat type.

Tuned Circuit Oscillators. The arrangement of Fig. 18 is particularly recommended for purity of waveform when correctly designed and constructed. It is possible to make it so that the total second harmonic content does not exceed 2 per cent over the range 30–15 000 cyc. Air-core coils are used for the range 6 000–15 000 cyc. and iron-core coils employing Gecalloy telephone-type dust cores are utilized for the range 30–6 000 cyc. High-grade intervalve and output transformers are used to maintain full response at all frequencies. Care should be taken to balance electrically each half of the centre-tapped tuned winding, likewise the intervalve and output transformer split windings. High "C" circuits should be used in the tuned circuit, and the anode currents of each pair of valves should be accurately matched. Using two Mazda A.C.P.1 valves in the oscillator stage and two Marconi-Osram P.X.4 valves in the output stage, 3 watts output is obtained with the second harmonic distortion previously stated.

The output valves should be biased well up the straight portion of their curves, whilst a d.c. supply to all the valve filaments is a desirable feature. The power supply unit should be liberally filtered for 50 cyc., 100 cyc., and 150 cyc. This is best accomplished by tuned trap circuits. The output transformer should be of the single-ratio type, and a "T" type attenuator should be connected to the secondary winding for control of output.

An inexpensive arrangement is shown in Fig. 19. A dynatron oscillator is used and a buffer stage isolates it from the load. By careful adjustments of the oscillator operating conditions

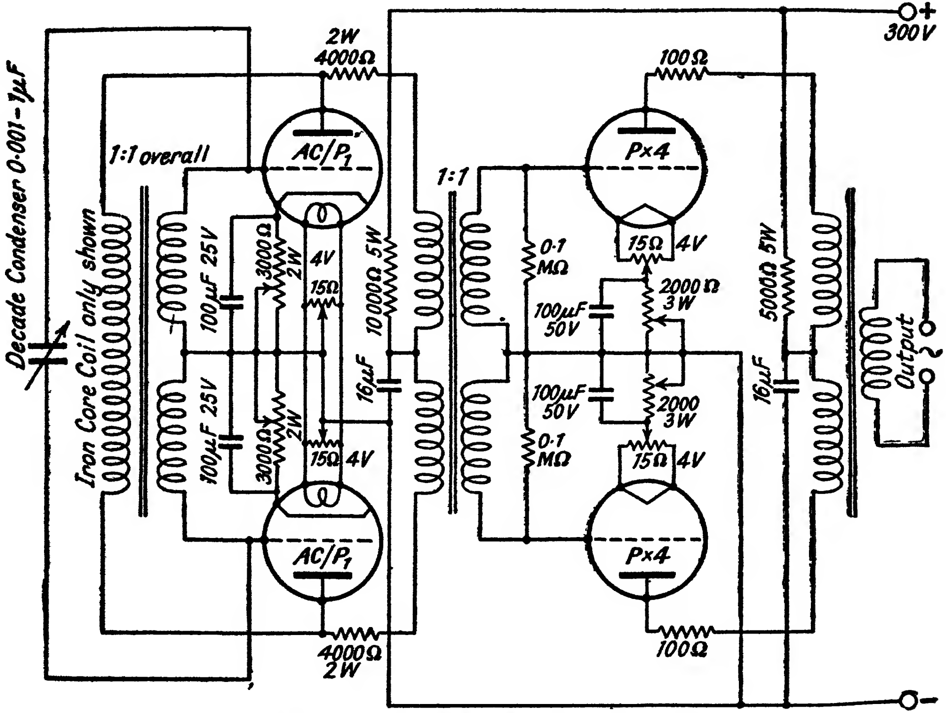


FIG. 18. AUDIO-FREQUENCY OSCILLATOR

a good waveform is obtained. In general the best operating condition for the dynatron is almost to the "step in the knee." This requires careful adjustment—yet not critical—of the bias potentiometer.

Beat Frequency Oscillators. The arrangement of Fig. 20 is perhaps the cheapest and simplest form of beat frequency

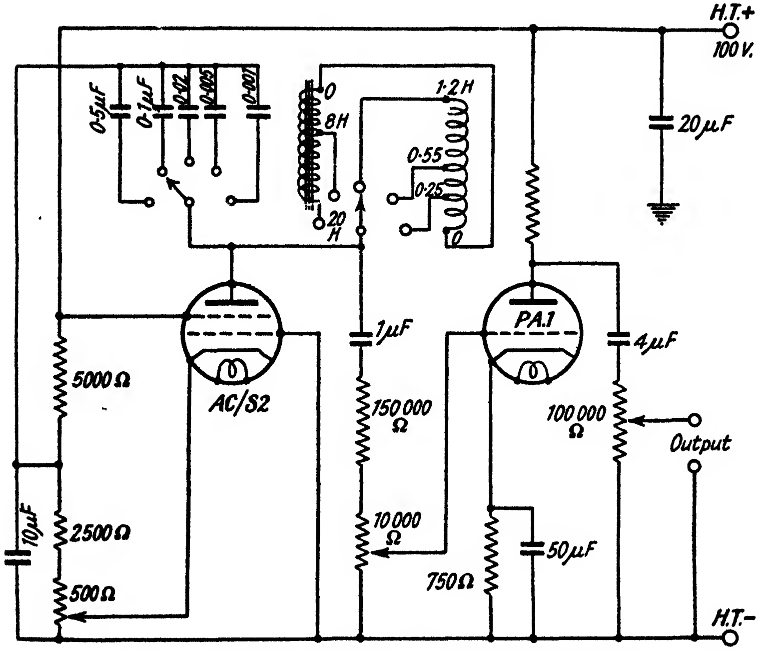


Fig. 19. Dynatron Audio-frequency Oscillator

oscillator. A twin triode, such as a Class B amplifier valve, serves the purpose of fixed and variable frequency oscillators, and since these are of the self-rectifying type it becomes possible to pick off the resultant audio-frequency component in the common anode circuit. Needless to say, this circuit cannot be regarded as a source of pure audio-frequency signals due to its basic principles of design.

Fig. 21 shows another inexpensive arrangement, but this is of considerably greater merit than Fig. 20, and in this case good waveform is claimed by the original designer, Mr. C. P.

Edwards, who described the instrument in detail in the *Wireless World.**

The two previous arrangements are departures from conventional beat frequency oscillator design practice, but Figs. 22–25 may be regarded as typical examples of standard practice. Before describing the salient features of these designs it is

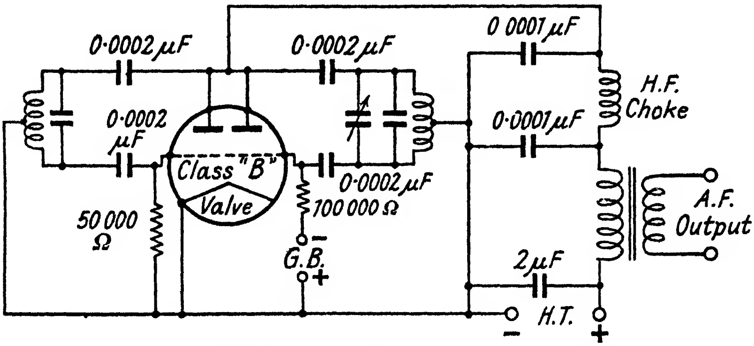


FIG. 20. INEXPENSIVE BEAT FREQUENCY OSCILLATOR

desirable to consider the most important fundamental requirements of conventional arrangements. The required performance can be tabulated thus—

(*a*) Maximum a.c. output . . .	3–4 watts
(*b*) Frequency range	50–10 000 cyc.
,, ,, preferably . .	30–15 000 cyc.
(*c*) Tolerable harmonic content in output	5 per cent below 100 cyc. 2 per cent above 100 cyc.
(*d*) Output level	± 2db. over complete frequency range
(*e*) Frequency stability . . .	± 1 cycle per hour after initial warming up is very desirable

The principle employed is to heterodyne two radio-frequency sources and rectify the resultant beat frequency, which is afterwards passed through a suitable audio-frequency amplifier. Some form of aperiodic coupling is used between the h.f. oscillators and the detector in order that the amount of h.f. applied to the detector shall not change with frequency. Anode bend rectification is most commonly used, and providing there is no grid current in this detector it is quite free from harmonic production. It is vitally important that one of the h.f. oscillators should have a pure waveform output, and the

* 22nd January, 1937.

fixed oscillator is usually designed with this in mind. In this connection it should be particularly noted that harmonics can only occur in the resultant rectified current if both h.f. carriers have harmonics. Also, in heterodyne reception, if one current is stronger than the other, the resultant rectified current is proportional to the weaker, assuming a linear rectifier. This may be explained as follows. Let X and Y represent the amplitude of the strong and weak carriers respectively; then the peak amplitude of the combined carrier varies between $X + Y$ and $X - Y$. The peak value of the low (beat) frequency current will be proportional to the difference, which can conveniently be expressed—

$$(X + Y) - (X - Y) = 2Y,$$

from which it will be seen that for a linear detector the amplitude of the rectified component is unchanged by a change in amplitude of the strong carrier. This also applies to a non-linear detector provided X is sufficiently strong to sweep over the non-linear portion of the detector on to the major portion of the straight portion of the detector characteristic, and that Y is not sufficiently strong to cause the instantaneous value of the d.c. component (the resultant rectified current) to reach a non-linear part of the detector characteristic. A detector operating under these conditions will not introduce any harmonics into the a.f. component.

Another important item is the prevention of "pulling in step" when each h.f. oscillator is operating at very nearly the same frequency. This trouble usually occurs when the maximum frequency difference between each oscillator is about 200 cyc. This may be somewhat overcome by the careful shielding of each oscillator, but unless further precautions are taken, the trouble is likely to persist and is usually traceable to feeding each carrier into a common detector. The use of push-pull detection almost invariably completely overcomes this trouble, and permits frequency differences of 1 cycle or even less before "pulling in" occurs. A low-pass filter is generally employed between the detector and a.f. amplifier. This is usually arranged to cut off frequencies above about 20 000 cyc. When transformer coupling is utilized in the a.f. amplifier stages, great care must be taken in the design of the transformers to prevent distortion. In this respect, accurately balanced push-pull stages are very helpful. It seems distinctly

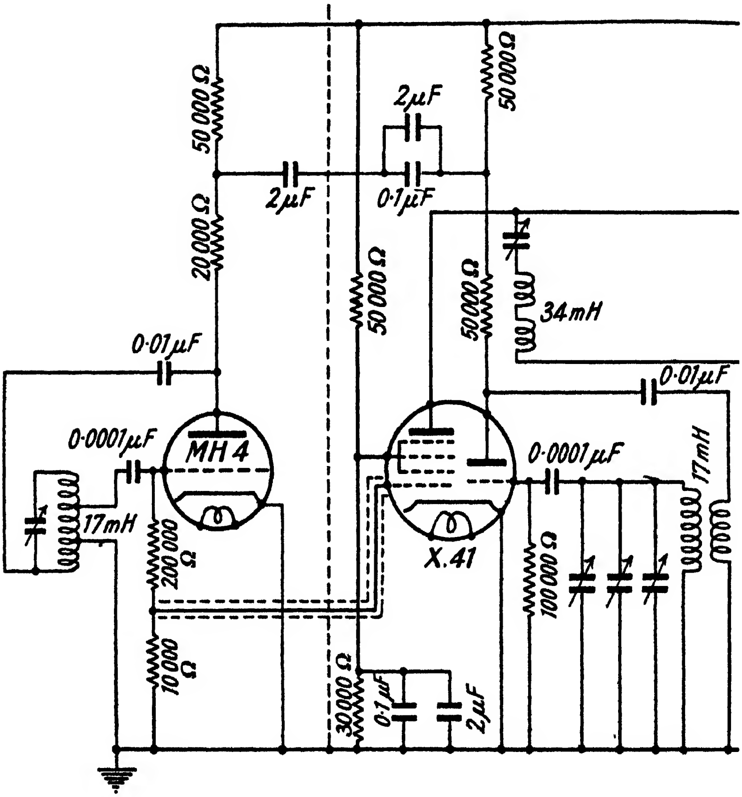
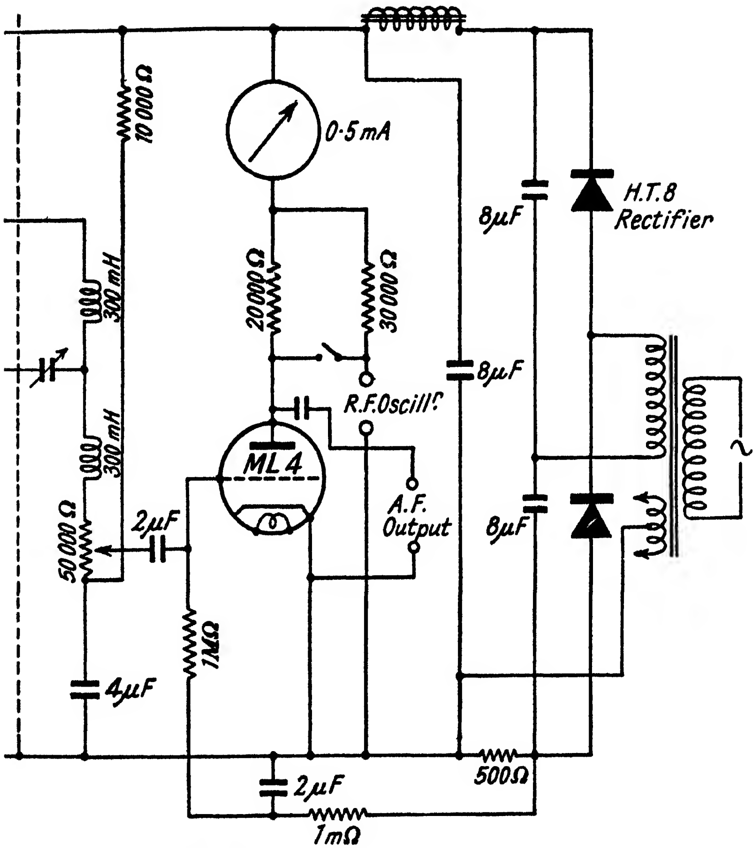


FIG. 21. SIMPLIFIED BEAT
(*Wireless World*,



FREQUENCY OSCILLATOR
22nd January, 1937)

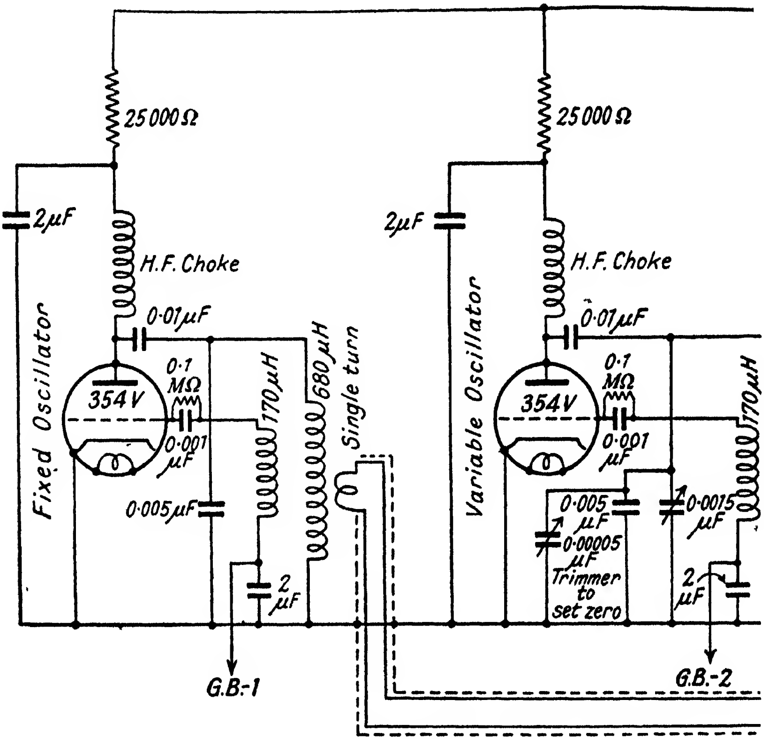
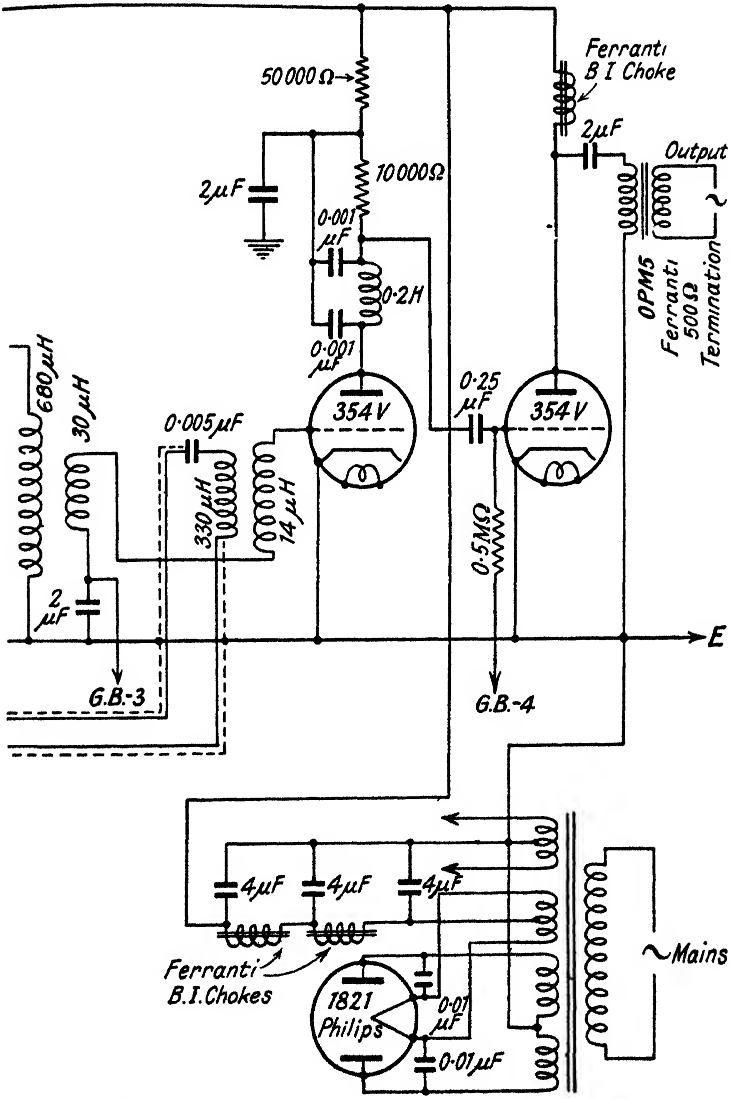


FIG. 22. BEAT FRE-
(Messrs. Cooper



QUENCY OSCILLATOR
and Page)

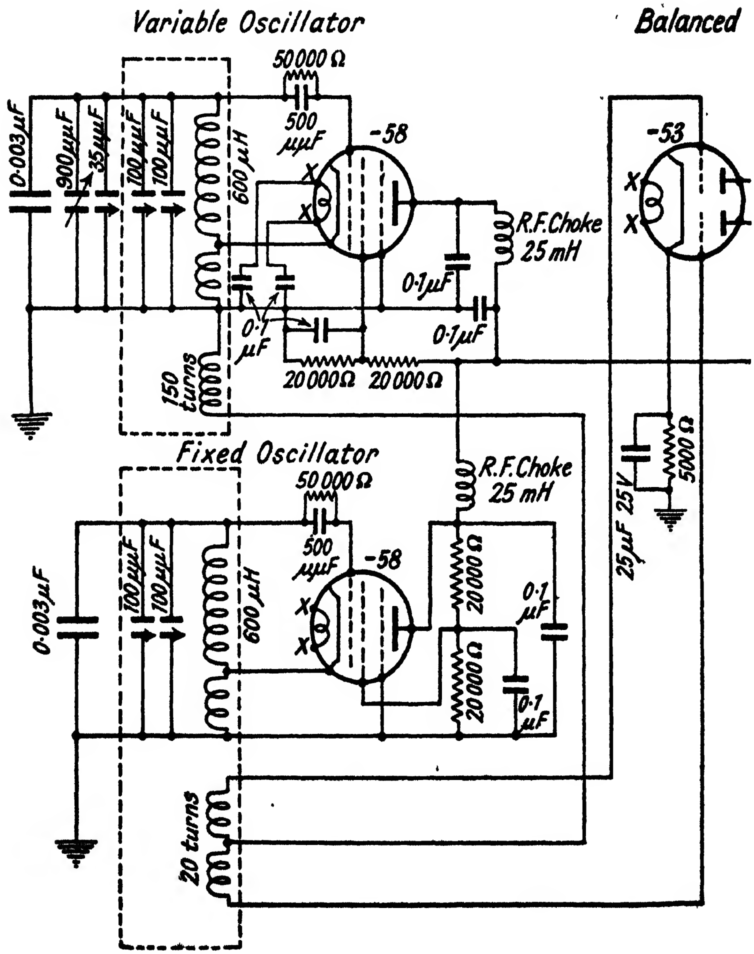
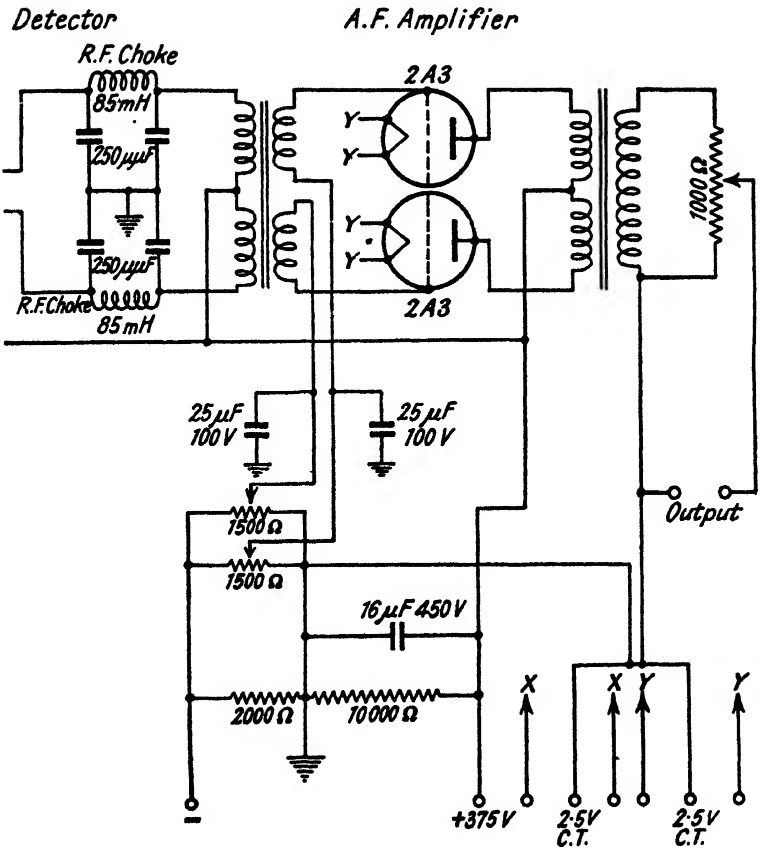


FIG. 23. BEAT FRE-
(Mr. de



QUENCY OSCILLATOR
Soto)

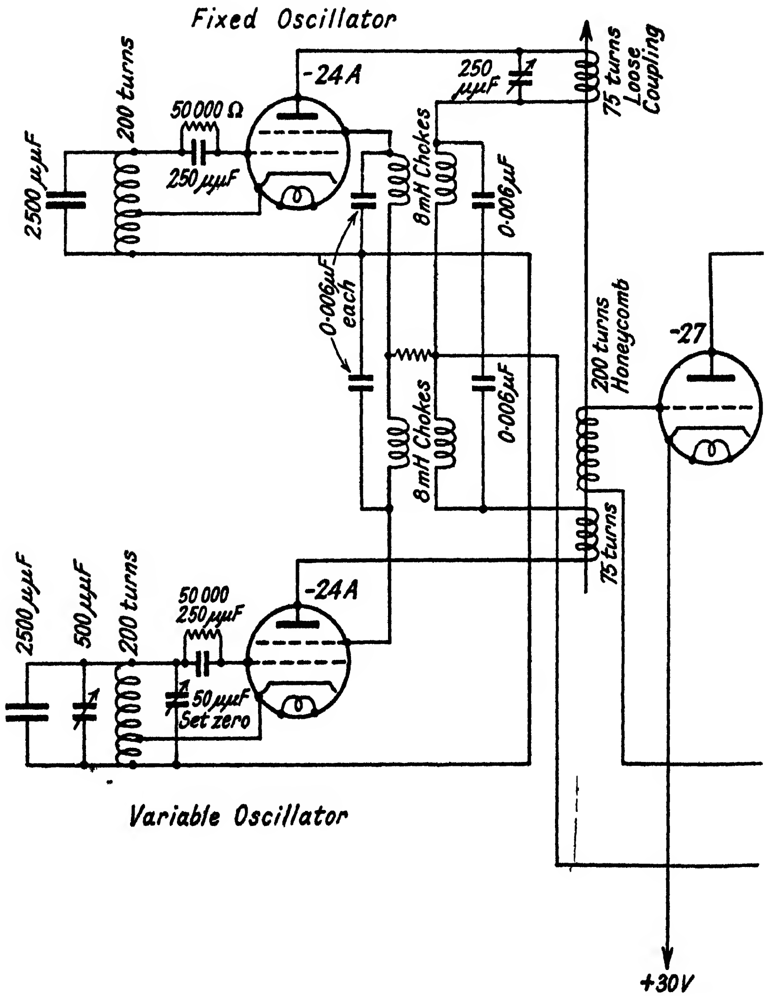
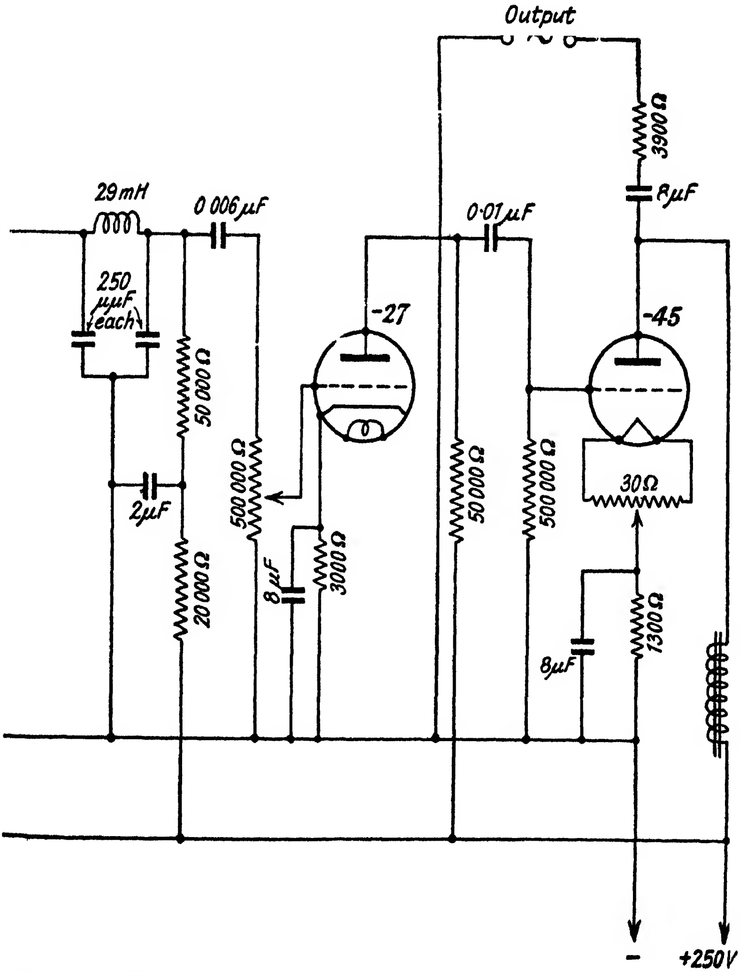


FIG. 24. BEAT FRE-
(Messrs. Haefner



QUENCY OSCILLATOR
and Hamlin)

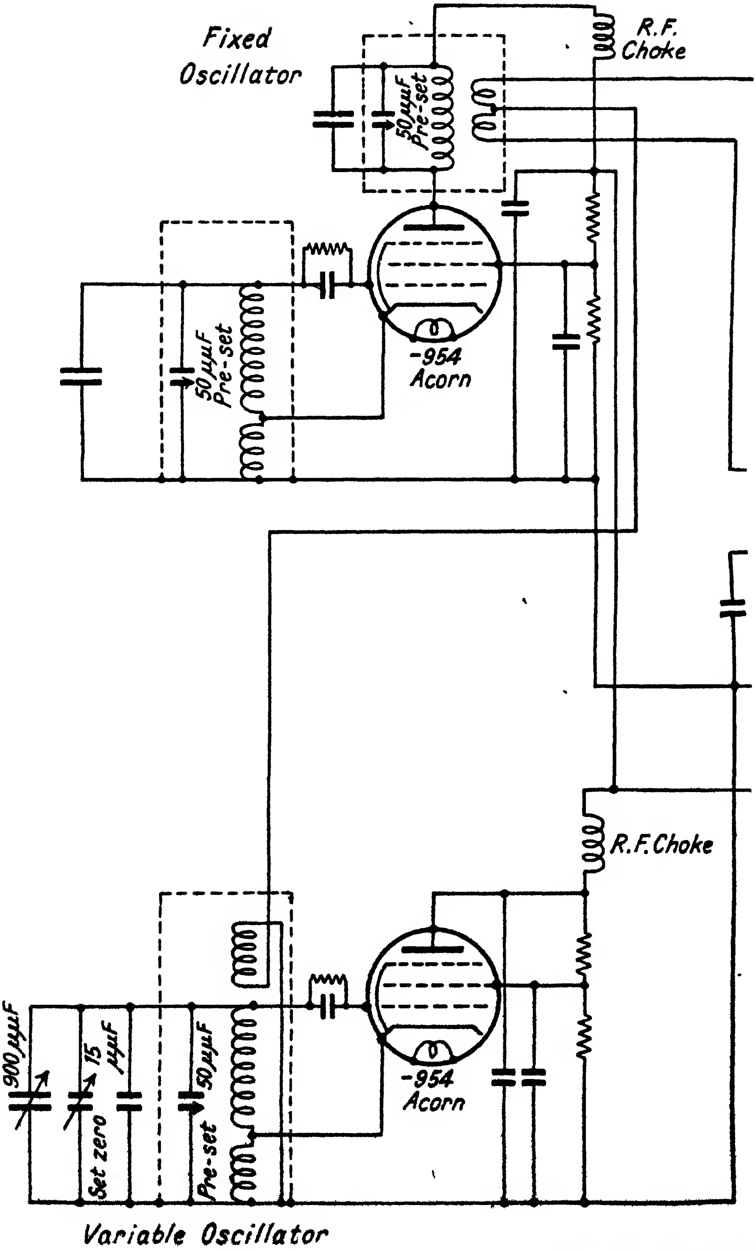
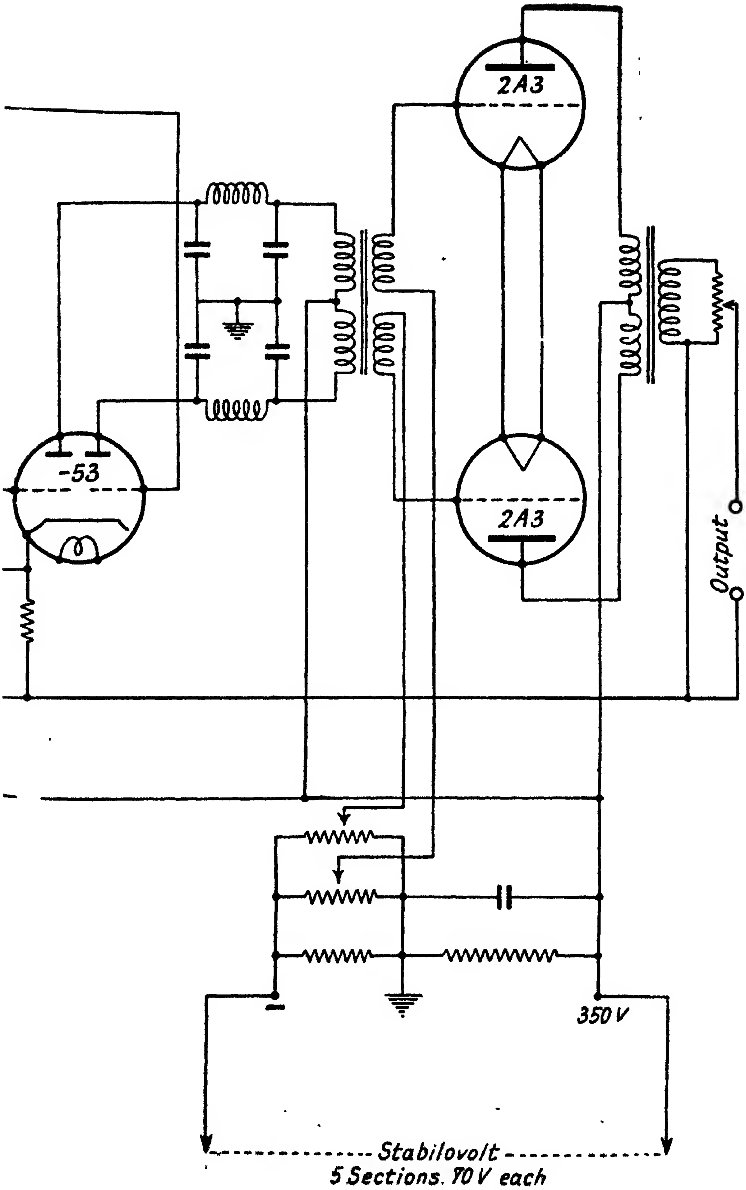


FIG. 25. BEAT FRE-



QUENCY OSCILLATOR

possible that a push-pull paraphase resistance-capacitance coupled a.f. amplifier would be admirable for beat frequency oscillators, and particularly if a push-pull beat detector was used such as in Figs. 23 and 25. The section on "H.F. Sources" deals with the merits and failings of common types of oscillators, and it will be seen that the Dow electron coupled oscillator and the dynatron oscillator are considered to be the most stable types. It is interesting to note that the designs of Figs. 23, 24, and 25 all employ electron coupled oscillators, but perhaps one of the best commercial products, the Ryall-Sullivan, uses the dynatron.

Triode oscillators have gone out of favour because their inherent stability is not so good as the electron coupled or dynatron types. It will be seen that Messrs. Cooper and Page's design of Fig. 22* employs tuned anode triode oscillators. However, the success of a beat frequency oscillator design does not entirely depend upon the choice of the most stable types of h.f. oscillators. Of comparable importance is the mechanical construction, and equally so, the physical properties, of the circuit components used in the h.f. oscillators.

Mechanical construction can perhaps be best dealt with by saying "rigidity" is the keynote of ultimate success, and at the same time due consideration should be given to thermal radiation and conduction caused principally by heat from the valves. From this viewpoint, battery type valves are certainly the best, since their temperature rise is negligible under operating conditions.

The temperature co-efficients of the coils and condensers used in the h.f. oscillators should be of a very low order, and in this respect the reader's attention is directed to the numerous papers of Griffiths, Groszkowski, and H. A. Thomas on this subject. The necessity for considering thermal conduction and radiation from the chassis assembly will now be realized. Stable voltage supplies are imperative if frequency stability of the order desired is to be obtained. The "Stabilovolt†" tubes make this a simple matter for the h.t. supplies, but great care must be taken to maintain the filament or heater voltages constant. Perhaps the most reliable method is to use a suitable filament heating transformer operated from a.c. mains. This, of course, demands mains type valves with their consequent disadvantage

* *Wireless Engineer*, Vol. X, No. 120, September, 1933.

† Stabilovolt Gesellschaft m.b.H. and Marconi W. T. Co. Ltd.

of increased heat radiation compared with battery types. In the Author's design of Fig. 25 the pros and cons of each type were carefully considered, and it was decided that a.c. filament heating and adequate provision for heat insulation of the h.f. circuits was the best method. It should be realized that the large thermal capacity of modern mains valves is such that momentary fluctuations in a.c. supply have little or no effect on the generated frequency. It would appear, therefore, that there is every justification for building a completely mains-driven instrument, although this is contrary to beliefs held four or five years ago. The description of basic requirements and principles earlier on will enable the reader to judge the merits and failings of the designs of Figs. 22–25. The Author's own design, Fig. 25, was produced after he had carefully considered each of the previous types as well as many others. Since this design may be regarded as having been produced from the cumulative experience gained from the arrangement shown in Figs. 22–24, a summary of its features compared with the previously stated fundamentals of design will not be out of place.

ELECTRICAL DESIGN

Oscillator Stability . .	Electron coupled oscillators.
Oscillator "Pull in" . .	Negligible due to push-pull detection, intensive screening and Mr. de Soto's special coupling system.
H.F. Oscillator Waveform .	The adoption of Haefner and Hamlin's double tuned fixed frequency oscillator results in an exceptionally pure waveform from this source.
Audio-frequency Amplifier .	Push-pull transformer coupling effectively cancels the second harmonic. Subsequent experience indicates improved response level may be obtained by substituting paraphase amplification.

PHYSICAL PROPERTIES OF H.F. CIRCUITS

	Frequency Drifts of Oscillators
(*a*) Valves	The acorn pentodes have minute inter-electrode capacitances, and thermal expansion of the electrodes causes a capacitance change so small that it cannot be detected by beat frequency drift.
(*b*) Coils	The use of aged loaded ebonite solid formers having a temperature coefficient equal to copper minimizes drifts in inductance value due to stress effects. (See Griffiths and Groszkowski, *loc. cit.*)

(*c*) Chassis assembly . . Thick sheet brass ($\frac{3}{8}$ in.) is used as a platform for the h.f. oscillators and maintains ambient temperature within close limits. This prevents the coils from becoming heated by the valves.

(*d*) Condensers . . . Either the Sullivan temperature coefficient-less type or silver sprayed Ardostan* types are suitable.

Full details of the designs of Figs. 23 and 24 are given in the following papers—

Fig. 24. Beat Frequency Oscillator: Haefner and Hamlin, *Electronics*, May, 1936.

Fig. 23. Beat Frequency Oscillator: Clinton B. de Soto, *Q.S.T.*, Vol. XX, No. 4, April, 1936.

* Ardostan has a temperature coefficient almost the same as quartz. Manufactured by Hermsdorf Schomburg Isolatorien Ges. Thuringia.

SECTION VI

HIGH-FREQUENCY SOURCES

Some form of high-frequency oscillator is required by even the smallest radio laboratory, and this piece of equipment will be put to so many uses (and most likely be so ill-treated throughout its life) that the choice of type justifies due consideration. Basic oscillator circuits in general use are shown in Figs. 26–32, and

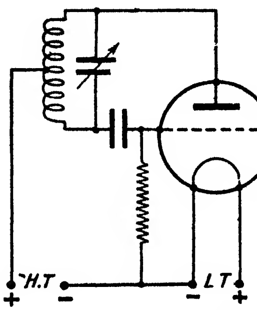


Fig. 26A. Series-fed Hartley Oscillator

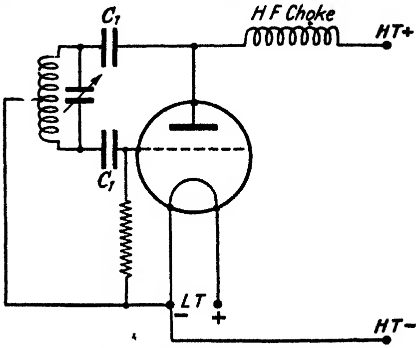


Fig. 26B. Parallel-fed Hartley Oscillator

of these the Hartley arrangements can be said to be the best known types. The principal difficulty with the Hartley arrangement is that each side of the tuning condenser is above earth potential and so presents certain body-capacitance problems. The Author's experience leads him to believe the parallel-fed arrangement of Fig. 26B to be more stable than the series-fed arrangement. Conditions for maximum stability are when the condensers C_1 are of equal value. Also since there is no d.c. flowing through the inductance, instability due to any heating effect from this source is removed. The h.f. choke in the plate circuit should present a high impedance to all frequencies it is desired to generate. A well-known make of iron-core choke having an inductance of approximately ½ henry fulfils this requirement from 140–3 500 metres satisfactorily.

Both Hartley arrangements produce an abundance of harmonics of the fundamental frequency.

The Colpitts arrangement of Fig. 27 overcomes the problem of each side of the condenser being at high potential, thus overcoming body-capacitance effects. The centre-tapped condenser can conveniently be a conventional 2-gang type with the rotor at earth potential. It will be seen that the condensers are virtually in series and therefore are effectively half the capacitance of one section. The Colpitts circuit shares the advantage of the dynatron in being a "two-terminal oscillator," and thus simplifies waveband switching when such is desired.

The necessity of a choke in the plate circuit calls for the same remarks as the Hartley circuit of Fig. 26B. Harmonic production

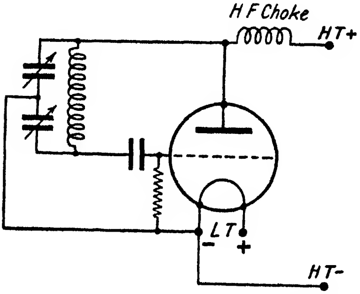


Fig. 27. Colpitts Oscillator

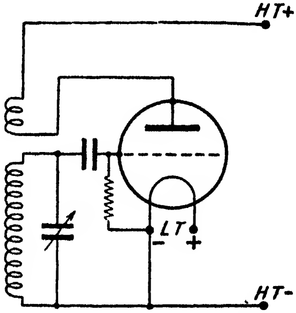


Fig. 28. Feed-back Oscillator

by the Colpitts circuit is usually appreciably less than with the Hartley arrangements, and this may be explained by the low impedance path presented to harmonic frequencies by each half of the tuning condenser. The normal feed-back circuit of Fig. 28 usually requires more initial attention in design than either of the previous examples. The magnetic coupling between the grid and plate coils requires careful adjustment, as does the number of turns on the plate coil. These items being correct, the general results are very similar to those of the series-fed Hartley of Fig. 26A. This circuit has the advantage of one side of the condenser being at earth potential.

For general laboratory work the circuits of Figs. 29 and 30 are not particularly suitable. In the T.G.T.P. (Fig. 29) arrangement the coils are not inductively coupled, the grid-plate capacitance of the valve being utilized to provide the coupling between the grid and plate circuits. The frequency of oscillation is principally controlled by the constants of the plate *LC* circuit. Although the grid *LC* circuit tuning affects the frequency of oscillation, its principal function is that of controlling the feed-back or excitation. The T.N.T. variation of Fig. 30 requires the grid coil plus stray capacitances to be broadly resonant at the desired operating frequency. It is a very simple

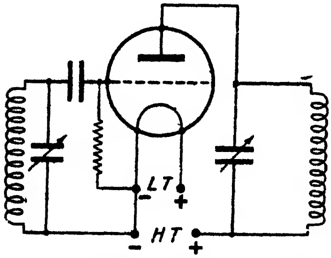


FIG. 29. ARMSTRONG OR TUNED GRID-TUNED PLATE OSCILLATOR

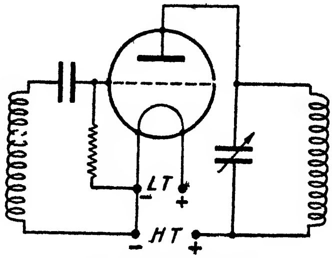


FIG. 30. SO-CALLED T.N.T. MODIFICATION OF THE OSCILLATOR IN FIG. 29

circuit to tune once the proper size for the grid coil has been determined.

The Author considers the electron-coupled circuit of Fig. 31 the very best arrangement for a general purpose laboratory oscillator. It is actually a development of the Hartley circuit

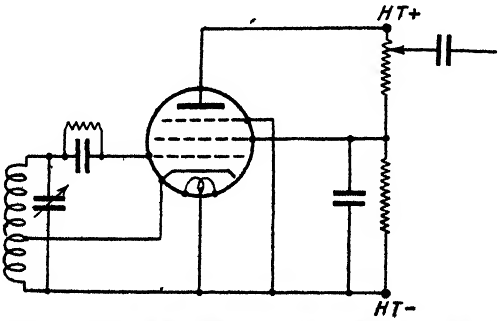


FIG. 31. DOW OR ELECTRON-COUPLED OSCILLATOR

made possible by the introduction of the screened-grid and pentode valves. The control grid, cathode, and heater are combined in a triode oscillating circuit with the screen at earth potential for radio-frequency voltage.

The output may be taken off the resistance as indicated, or a separate tank circuit in the proper anode circuit which may be tuned to the fundamental or harmonic frequency. It should be noted that harmonics above the third are very weak with this arrangement. With proper design the electron coupled oscillator provides frequency stability which is often compared with crystal control, whilst appreciable changes in h.t. voltage have but small effect on frequency. It is for these reasons that the Author pins his faith to this circuit.

Although the dynatron arrangement of Fig. 32 is an exceptionally stable oscillator, its critical operating requirements, such as the critical screen and plate voltages (note the screen is at higher potential than the plate), and also the necessity of selecting a suitable valve—not all tetrodes being good dynatron oscillators—are items which cannot be tolerated in view of the "sure fire" robust qualities demanded of the general purpose laboratory oscillator.

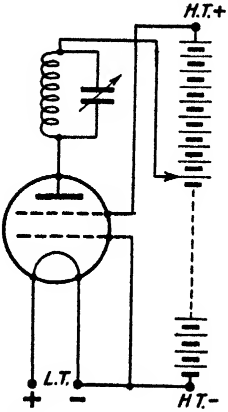


FIG. 32. DYNATRON OSCILLATOR

A few words as to the general make-up will not be out of place.

Unscreened solidly constructed coils with tags for connection under terminal heads are by far preferable to flimsy cardboard tube types with waveband switching. The condenser should be a good solid job, the ex-Government types enclosed in glass cases being recommended. These can be bought very cheaply and are far superior to the mass-produced types normally used in radio practice. A Marconi M.S.P.4 valve is an excellent oscillator with about 120 volts on the plate and 70 volts on the screen. The grid condenser is preferably of the air dielectric type and about 50–100 $\mu\mu$F. capacitance, whilst the parallel grid leak is about 100 000 ohms, 1 watt. By constructing the whole on a solid wooden board with ebonite terminal strips for battery connections and 16 S.W.G. wire throughout for connections, a really robust job results which will stand an enormous amount of knocking about. Such a piece of equipment is actually a small transmitter, and therefore should not be coupled to an aerial. The signals radiated from an electron coupled oscillator such as described, can be comfortably received on a simple detector without reaction 100 yd. distant.

WAVEMETERS, TEST OSCILLATORS, STANDARD SIGNAL GENERATORS

Wavemeters. If the high-frequency source just discussed is a well-made affair, it will most probably be calibrated and used as a heterodyne type wavemeter. However, this section would be incomplete without reference to really precision wavemeters, and for this purpose the Sullivan-Griffiths Type R.202 will be briefly described. A dynatron circuit is employed, and the associated coil and condenser are of special design to ensure a high degree of permanence of calibration. The condenser follows Mr. Griffiths's Corrected S.L.F. Law and the resultant accuracy may be judged when the accuracy of simple interpolation is one part in 10 000. The mathematical details of this special condenser will be found in Section I of this book, "Condensers." The dual range feature embodied in the R.202 wavemeter is the logical "two gang" development of the Corrected S.L.F. Law. The Author has used this wavemeter extensively, and considers it to be the most easily operated sub-standard he has yet encountered. A Table of Accuracies prepared by the makers and rigorously checked by the Author is given hereunder.

TABLE XXXIV

SULLIVAN-GRIFFITHS R.202 WAVEMETER

Long period accuracy and stability.	1 part in 10 000
Temperature coefficient of frequency	5 parts in 10^6 per ° F.
Reading accuracy	3 parts in 100 000
Accuracy of simple arithmetic interpolation	1 part in 10 000
Short period stability . . .	1 part in 100 000

A circuit diagram of the R.202 is given in Fig. 33, from which it will be seen a detector circuit is embodied in the design. The Author personally prefers to use a sensitive and fairly selective radio receiver for the purpose of detecting heterodyne beats, and so obviating the necessity of tight coupling to the built-in detector circuit. The two dynatron valves (A.C. S.2 and S.4 V.B.) are used for the wavebands 10–300 metres and 300–10 000 metres respectively, whilst a small precision variable condenser is connected in parallel with the tank circuit to compensate for the different inter-electrode capacitances of the valves. This valuable device enables a constant self-capacitance to be maintained in the circuit, thus

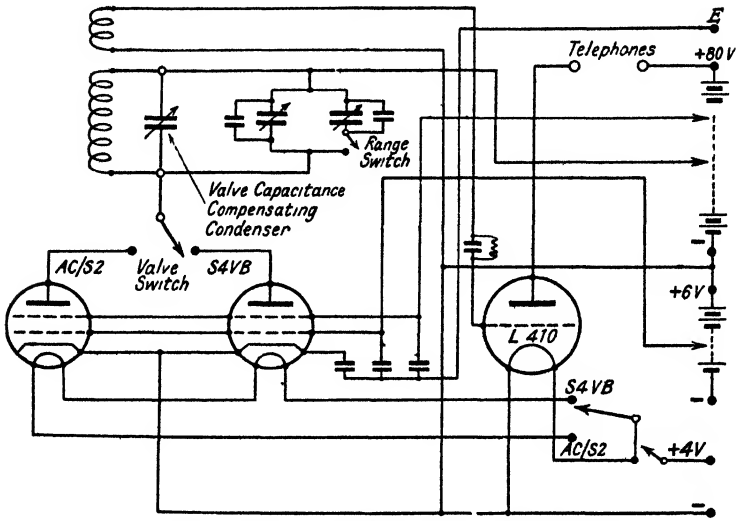


FIG. 33. SULLIVAN-GRIFFITHS R.202 WAVEMETER
Skeleton circuit diagram; meter switching, etc., omitted

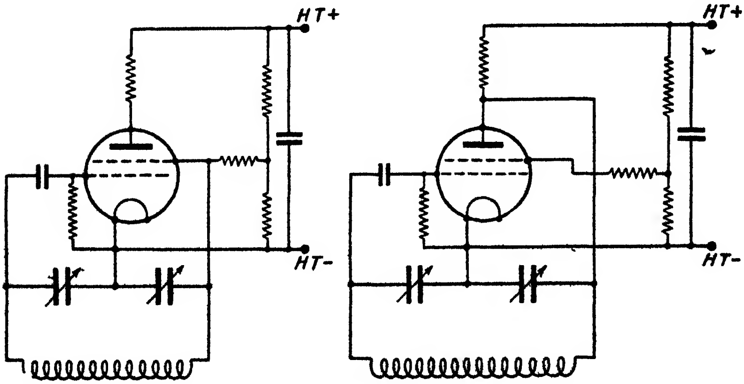


FIG. 34. BASIC ARRANGEMENTS USED IN G.R. FREQUENCY METER: U.S. COASTGUARD SERVICE
(a) Colpitts electron-coupled oscillator. Screen-grid used as plate
(b) Colpitts electron-coupled oscillator

preserving accuracy of calibration with change of valves. Turning now to instruments of more robust character but yet in the precision class, it is particularly interesting to refer to a design employed by the U.S. Coast Guard and manufactured by the General Radio Co., Cambridge, Mass., U.S.A. These

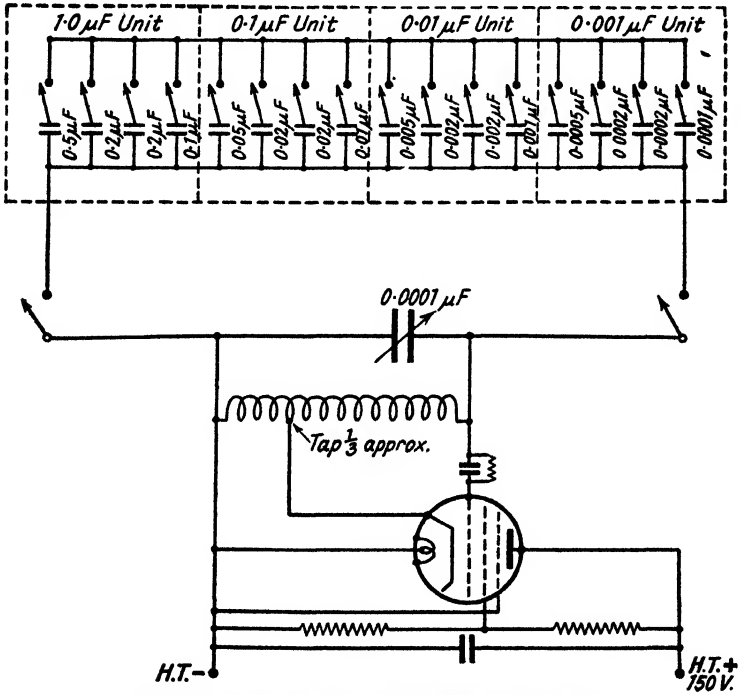


FIG. 35. WIDE RANGE WAVEMETER OSCILLATOR
Used in Author's own laboratory

are portable instruments employing the electron-coupled oscillator, but instead of the Hartley arrangement, a Colpitts circuit is used. With some valves it is sometimes found that for conditions of maximum stability, the plate must be operated at a lower potential then the screen. A skeleton circuit diagram of the Colpitts circuit applied to the electron coupled oscillator is shown in Fig. 34. This is the basic arrangement of the General Radio Co.'s design.

The Author uses a wavemeter in his own laboratory for general purposes which has an extremely wide range, actually covering the higher radio frequencies to upper audio frequencies. The range is 60 mc. to 10 kc. A Hartley circuit in an electron coupled oscillator is employed for reasons given previously, and also because it will oscillate with large amounts of capacitance in relation to inductance.

It will be seen that a decade condenser augmented by a variable condenser is used for tuning purposes. A series of plug-in coils is used with this combination. For purposes of economy the decade condenser departs from the conventional decimal steps arrangement, and instead the units 5, 2, 2, 1 are manipulated to give each unit to 10. The use of four condensers instead of the usual ten per decade results in simplification of lay-out and much reduced initial cost. The arrangement is shown in Fig. 35. Although there is provision for capacitance in excess of 1 μF., 0·25 μF. is the maximum used with the oscillator. The complete condenser unit is easily removable, whence the extra capacitance is often valuable in bridge measurements, etc.

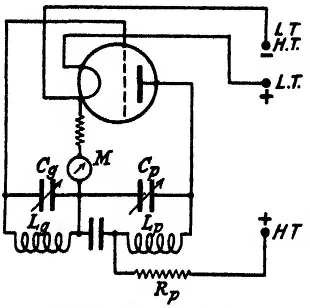


FIG. 36. SKELETON CIRCUIT DIAGRAM: OSCILLATOR OF BURGESS BATTERY CO. DESIGN

A particularly interesting design of oscillator-wavemeter is described in the American Burgess Battery Company's Circular No. 12. The designers have striven to maintain the h.f. output constant, irrespective of the setting of the tuning condensers. A simplified circuit diagram is shown in Fig. 36 and will be seen to take the form of a tuned plate, tuned grid oscillator employing series plate feed. The resistance R_p is used in place of a choke, but it is found that a good h.f. choke is more efficient in preventing high-frequency currents from flowing through the h.t. battery. It can be shown that the ratio of voltage drop across L_p to the drop across L_g is always the same, since $L_p = L_g$ and the halves of the tuning condenser C_p and C_g are equal. Thus, with the excitation of the grid constant, the h.f. output current should also be constant throughout a considerable tuning range. However, on the shorter wave ranges this is not true, as with the coils of but few turns required for these ranges, the current is, of course, not constant throughout the length of the coil. Also, the resistance of the coils increases with frequency which causes a decrease in output. The coils are wound on bakelite formers 3 in. diameter, 4½ in. long, whilst the pin connections for each coil are 1½ in. centres apart. General details are given alongside Fig. 36A, which gives the full circuit diagram.

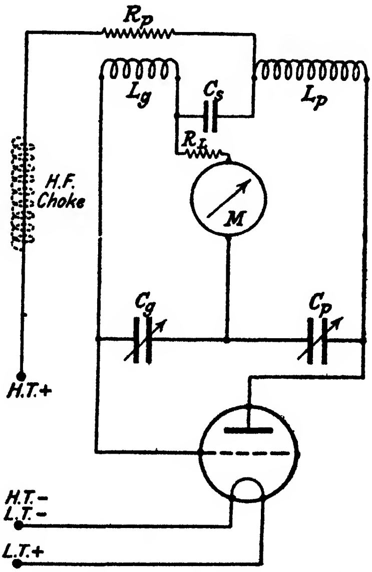


FIG. 36A. DETAILED DIAGRAM OF OSCILLATOR SHOWN IN FIG. 36

VALUES OF COMPONENTS IN FIG. 36A

C_g, C_p = 350 μμF. each;
C_s = 0·01 μF.;
R_L = 2 500 ohms;
R_p = 400 ohms;
M = 0–10 mA.

INDUCTANCES

Turns per Section	Total Turns	Wire Gauge (B. & S.; D.C.C.)	Wavelength Range
2	4	16	12– 32
5	10	16	25– 67
13	26	16	54–150
33	66	22	135–380
74 (2 layers of 37)	148	22	310–800

Test Oscillators. The main purpose of test oscillators is to provide a radio-frequency signal of known wavelength and of controllable output. The signal is usually modulated at about 400–1 000 cyc., the depth of modulation being between 20 and

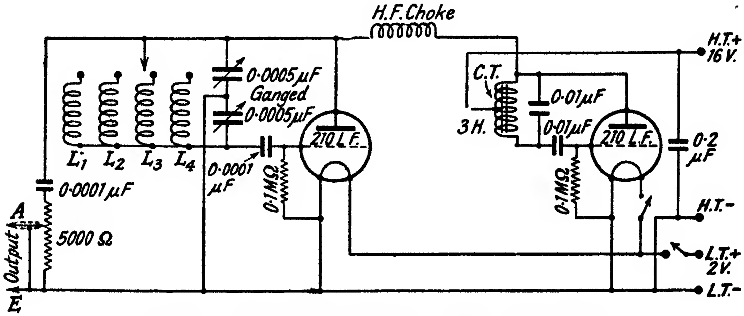


FIG. 37. "WIRELESS WORLD" TEST OSCILLATOR

70 per cent, depending largely upon the tuning of the high-frequency circuit. Although the radio-frequency output is controllable, it is not calibrated in actual values, although an arbitrary calibration in decibels is sometimes provided.

A typical design appeared in the *Wireless World** and the general details are reproduced in Fig. 37.

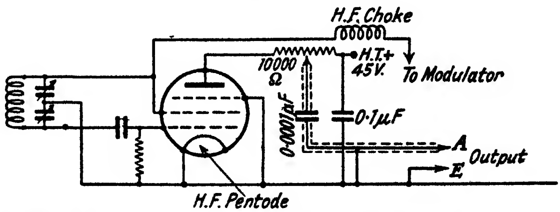


FIG. 38. MODIFIED OUTPUT CIRCUIT FOR "WIRELESS WORLD" DESIGN
(Proposed by Author)

A Colpitts circuit is used for the h.f. oscillator, and a series fed Hartley circuit for the l.f. oscillator which serves the purpose of modulator.

In a subsequent issue of the *Wireless World*, the Author gave a brief description of the circuit of Fig. 38, which has an improved form of output control for Fig. 37. A h.t. voltage of

* 10th–17th May, 1935.

45 volts is desirable on the proper anode, and about 24–36 volts on the screen serving the function of anode. This modification

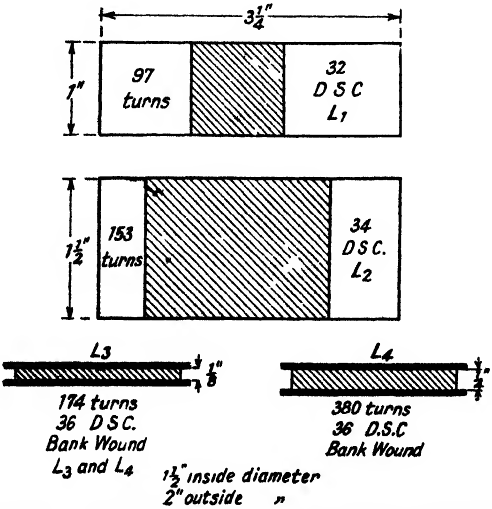


FIG. 39. "WIRELESS WORLD" TEST OSCILLATOR COILS

will, of course, be recognized as the electron coupled circuit. The details of the coils are given in Fig. 39.

A design due to Mr. R. F. Shea, which appeared in the

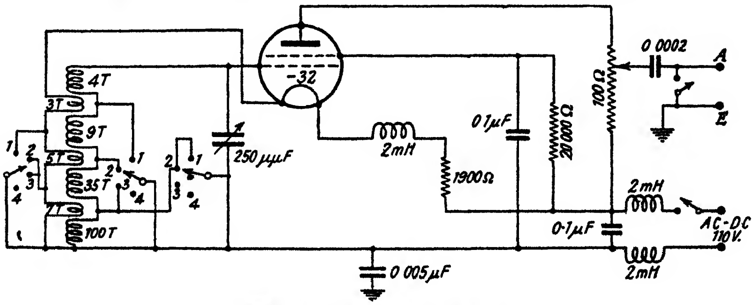
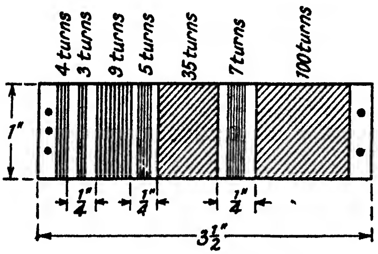


FIG. 40. TEST OSCILLATOR
(Designed in U.S.A. by Mr. R. F. Shea)

American *Q.S.T. Journal* and the American Radio Relay League *Amateur's Handbook*, 1937, is typical of many commercial types of inexpensive test oscillators produced in America.

The circuit diagram is shown in Fig. 40, whilst the coil construction is shown in Fig. 41.

The oscillator may be operated directly from 110 volts a.c. or d.c. supply mains. When used on a.c., a note of mains frequency serves the function of audio modulation. The continuous frequency range is 12 mc. to 545 kc. This type of oscillator is extremely useful in populous areas, where almost every house is fitted with electricity, but its limitations in rural areas will be appreciated. The so called *mains modulation* is a cheap method of securing an audio note, and must be considered far superior to the "squegger" oscillators found in rather large quantities on the British market. To extend the range of Mr. Shea's design to include the European long-wave band, and also intermediate frequencies up to 100 kc., the addition of coils similar to L_3 and L_4 of Fig. 39 will be required. However, these should be tapped at 43 turns and 95 turns respectively to meet the requirements of the electron coupled oscillator. Whether the extra ranges are included or not it is desirable to short-circuit all coils not in use to prevent "dead spots" in turning.



4 turns.	23	En.	B&S.
3 "	32	D.S.C.	"
9 "	28	En.	"
5 "	32	D.S.C.	"
35 "	37	En.	"
7 "	32	"	"
100 "	36	"	"

FIG. 41. COIL DESIGN FOR TEST OSCILLATOR
(Mr. R. F. Shea)

An English design which has enjoyed considerable popularity and received many favourable Press reports is the E.M.I. model, which was designed by the Author whilst with the manufacturers, Messrs. Everett Edgcumbe & Co. Ltd. The oscillator was produced to fulfil the requirements of the service man in as complete a manner as possible. It embodies the following features—

(1) Complete portability.

(2) Robustness and reliability.

(3) Continuous wave range from 175–3 000 metres on fundamental frequencies only.

(4) Liberal overlap on all ranges, and important calibration points coming at about mid-scale.

(5) Calibration accuracy initially better than half of 1 per cent and, with batteries maintained in good order, holding within 1 per cent almost indefinitely.

(6) Wide range of h.f. output control.

The circuit diagram is shown in Fig. 42, whence it will be seen a parallel-fed Hartley circuit is used for the h.f. oscillator and a series-fed Hartley circuit for the modulator. Attention should be directed to the location of the coupling coils as these were the subject of considerable experiment with a view to maintaining reasonably level output on all wave ranges.

The incorporation of a padding condenser for Range 4 obviates the necessity of a very large inductance. It also

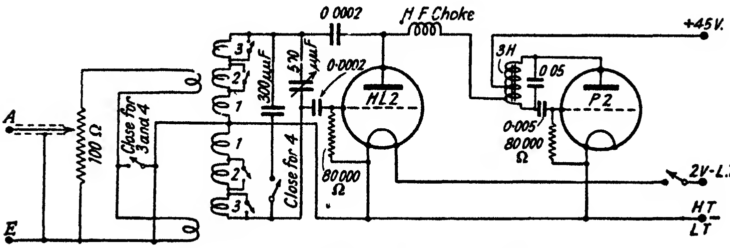


FIG. 42. E.M.I. TEST OSCILLATOR
Skeleton diagram

ensures that Ranges 3 and 4 have a satisfactory degree of overlap. The maximum output is approximately 1 volt and the minimum approximately 10 μV. The modulation depth is reasonably constant for inexpensive apparatus of this type, and varies between 25 and 40 per cent, depending upon the tuning condenser setting.

All test oscillators so far described suffer from relatively inefficient output control circuits. Here the Author would like to record his condemnation of the term "attenuator," so wrongly applied to simple voltage dividers used in most oscillators. Although it is quite impossible to incorporate a precision attenuation network in simple equipment, it is now possible to use improved forms of output circuits due to the introduction of composition type "T" and "L" resistance networks having a similar form to twin volume controls. These controls are made by the Electrad Co. in the U.S.A. The application of a 15-ohm unit to an electron-coupled oscillator is shown in Fig. 43. The maximum output voltage developed across the attenuator is about 0·1 volt, but this is adequate for almost every radio test purpose on modern sets. The use of a 50-ohm potentiometer permits the output to be controlled in extremely fine steps, whilst the minimum output is less than 1 μV.

It is predicted that these units will be used fairly extensively in such apparatus in the future.

Standard Signal Generators. This type of equipment enables the overall performance of radio receivers to be checked in a convenient manner. Tribute must be paid to the published

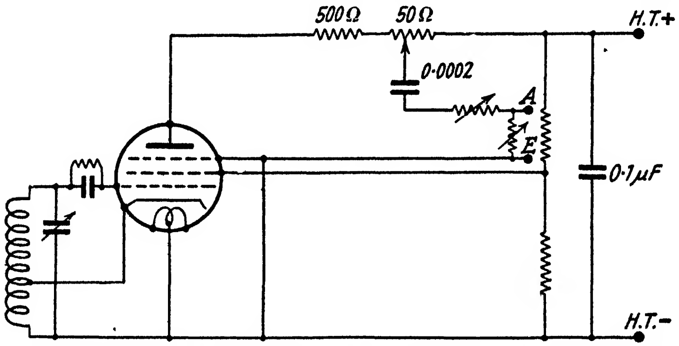


FIG. 43. EXPERIMENTAL OSCILLATOR USING COMPOSITION TYPE ATTENUATOR

works of Mr. H. A. Thomas of the National Physical Laboratory, Mr. H. D. Oakley, whose paper appeared in the *Journal of the A.I.E.E.*,* and Dr. Lewis M. Hull, who described perhaps the first portable pattern of signal generator in his lecture to the Radio Club of America on 13th June, 1928. This original design was commercialized by the General Radio Co., U.S.A., and known as "Type 403." More recent designs are Types 601 and 603, skeleton diagrams of each being given in Figs. 44A and 44B respectively.

The Author is not particularly enamoured with the term "Standard Signal Generator," as an examination of the various diagrams cannot fail to show the reader how various errors can occur in both the modulation and r.f. voltage measurements. For this reason the word "Standard" is not entirely justified.

A design due to Mr. Richard F. Shea was described in *Electronics*, August, 1934. This was specially developed for

* Vol. 46, 1927, pp. 498–503.

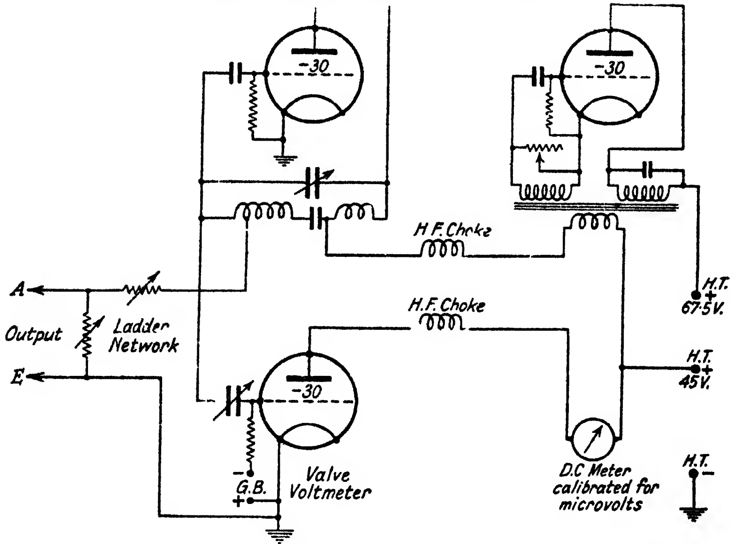


FIG. 44A. G.R. TYPE 601 STANDARD SIGNAL GENERATOR

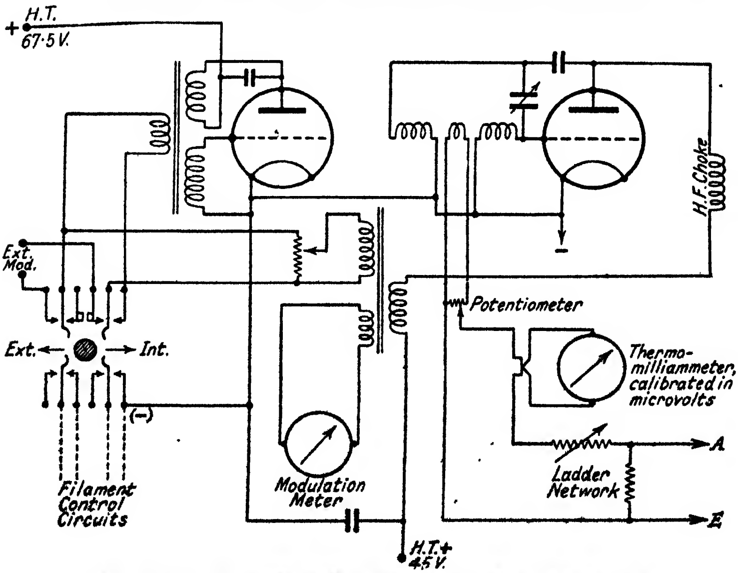


FIG. 44B. G.R. TYPE 603 STANDARD SIGNAL GENERATOR production line tests, and instead of having a continuously variable frequency range, a number of selected fixed frequencies were used and picked out by means of a rotary switch. The range covered by the steps is 16 000 kc. to 115 kc., thus justifying the name "All-wave Signal Generator." Constructional details of the attenuator are given, also mechanical assembly. A skeleton circuit diagram, omitting only the

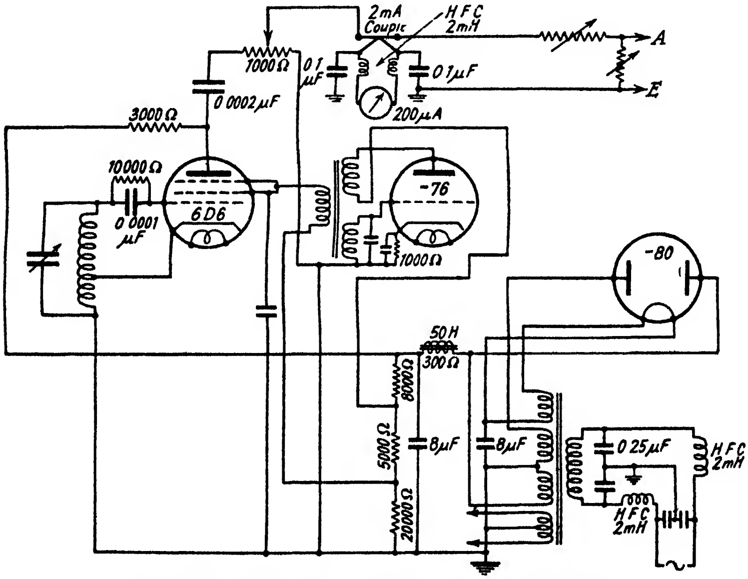


FIG. 45. PRODUCTION LINE SIGNAL GENERATOR
(Designed by R. F. Shea)

cathode resister, "swamp" rheostat, and switching arrangements, is shown in Fig. 45. For the lower frequencies, a 1 000-ohm resister is connected between the cathode and cathode tap on the coil, this serving to reduce harmonic production which is most severe at these frequencies. Each coil in Mr. Shea's design embodies a 1 000-ohm rheostat which serves as a "swamp," and by suitable adjustment of individual rheostats, a level output is obtained on each range. A suggestion the Author desires to make is that better modulation would be obtained by simultaneous modulation of screen and anode. A greater degree of freedom from any tendency to frequency drift due to anode loading may be obtained by grounding the suppressor grid which tends to reduce the output somewhat. However, 0·5 volt is still obtainable, which is adequate for most purposes including a.v.c. checks. After constructing a signal generator of this pattern to Mr. Shea's design and incorporating the above modifications, the Author has found it a very useful tool for mass production tests.

Another signal generator useful for production line tests is

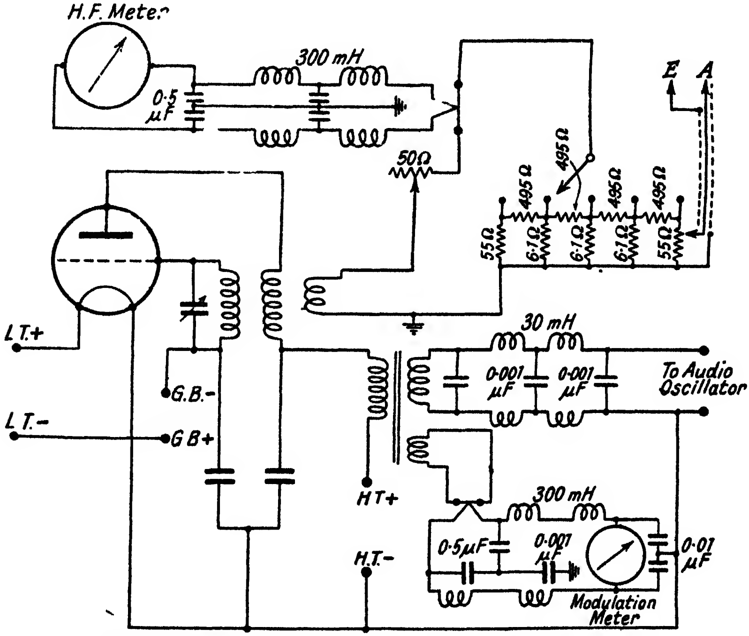


FIG. 46. SIMPLE PRODUCTION LINE SIGNAL GENERATOR (German design)

shown in Fig. 46. This is a very cheap and simple design, but nevertheless quite efficient. No self-contained modulator is employed. A small German firm employs three such units, and modulates them from a single beat-frequency oscillator and amplifier. When receivers are not being tested, the audio oscillator is free for other work. This system is certainly economical and enables the maximum use to be obtained from subsidiary apparatus. The firm in question employs no more than thirty operatives, but yet produces some really good receivers and amplifiers. They are quite well equipped with test gear, mostly constructed by themselves. Such a policy is certain to have its rewards, and the Author wishes that more British small manufacturers would adopt such measures.

The Author has produced a design for an inexpensive signal generator which has some advantages not ordinarily possessed by instruments of this type. The use of suppressor grid modulation enables modulation depths up to 90 per cent to be obtained

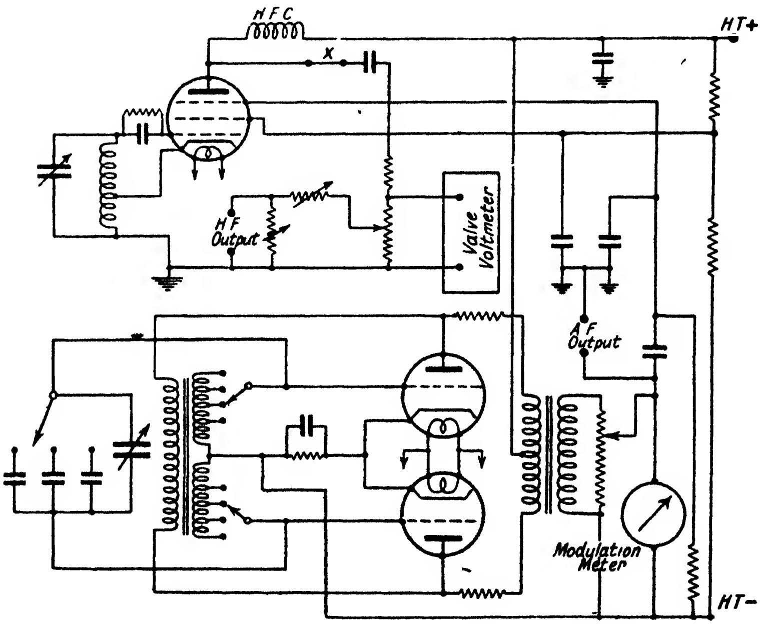


FIG. 47. SIGNAL GENERATOR
(Author's design)

without causing appreciable frequency drift; similarly the distortion produced is of a very low order. The use of a push-pull audio oscillator, variable in discrete steps between 50 and 10 000 cyc., enables fidelity checks to be easily undertaken. A valve voltmeter utilizing an R.C.A. acorn valve, Type 954, is used as the high-frequency voltmeter, as this has negligible frequency errors. The rectifier voltmeter in the audio-frequency oscillator is calibrated against a cathode ray oscillograph to give direct readings of percentage modulation. In the experimental model 8 volts and 15 volts gave 30 per cent and 50 per cent modulation respectively. The rectifier voltmeter utilizes a 1 mA. movement and rectifier which is reasonably free from error in the audio range. The schematic diagram of the valve voltmeter will be found in Section VII which deals with such apparatus. The incorporation of a resistance of 500 to 1 000 ohms between cathode and cathode tap on the h.f. oscillator coil serves the function of reducing harmonics at the lower frequencies. In an improved model a tuned anode circuit, incorporated at the point marked *X*, was fitted in place of the present choke feed; also a coupling coil from which the h.f. output was fed to the attenuator.

SECTION VII

VALVE VOLTMETERS

So many types of valve voltmeters have been developed in the past that it may be confusing to those who are not completely conversant with the limitations of the various types to choose the most appropriate for a specific purpose. Experience has

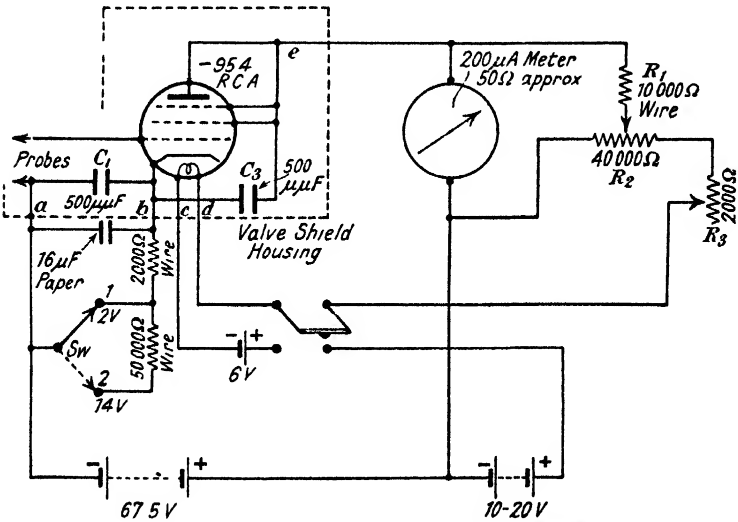


FIG. 48. ACORN PENTODE VALVE VOLTMETER
(R.C.A. design)

Leads *a*, *b*, *c*, *d*, and *e* return inside cable. Lead *a* is connected to valve housing.
The 16 μF. condenser is for calibration at 50 cyc. and measurement at low frequencies.
Resistances R_1, R_2, and R_3 are the "backing off" circuit for the meter.

shown the Author that no better type exists for measurements in the radio spectrum up to 60 Mcyc. than the "Acorn" pentode valve voltmeter developed by R.C.A. engineers. Full details were given in *Radio Engineering** and the *Wireless World*,† to name but two sources of many. A Type 954 Acorn pentode is connected as a triode to increase the mutual conductance, and is housed in a metal casing which also contains the two

* April, 1935. † 20th November, 1936.

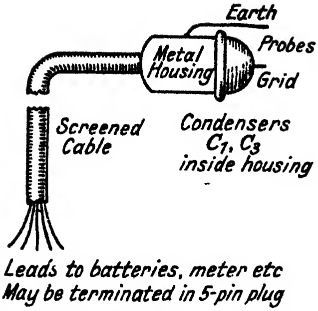
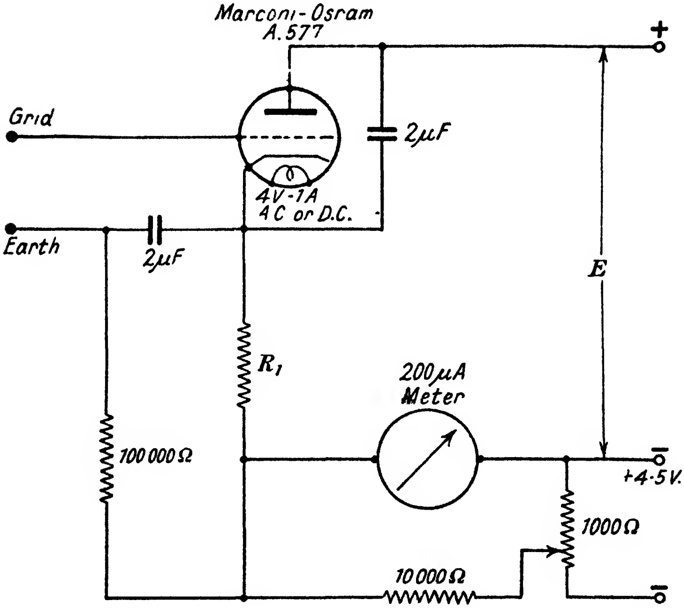


Fig. 48a. Acorn Pentode Valve Voltmeter
General mechanical arrangement



R.M.S. Voltage Range	0-5	0-15	0-50	0-100	0-150
Supply Voltage E (Anode + Cathode Bias)	35	75	270	270	270
Bias Resistance R_1	13 000 ohms	60 000 ohms	0·25 MΩ	0·55 MΩ	0·8 MΩ

Fig. 49. G.E.C. Valve Voltmeter
Multi-range 0·2 volt to 150 volts r.m.s.

by-pass condensers of 500 $\mu\mu$F. each. The leads for potential supplies are cabled, and the valve unit connected at the remote end, thus enabling the grid pin of the valve to be placed at the point of measurement, so obviating long leads.

The voltmeter gives full scale readings of 2 volts r.m.s. and 14 volts r.m.s. on switch positions 1 and 2 respectively. The input impedance is very high, whilst the input capacitance is about 2 $\mu\mu$F. only. The circuit diagram is given in Fig. 48.

For general purpose use at audio-frequencies and in the radio-frequency band up to about 3 Mcyc., the reflex valve voltmeter developed by the General Electric Co. Ltd. is admirable. It features high voltage range, high input impedance (about 5 MΩ. at 1 Mcyc.), and an input capacitance of approximately 10 $\mu\mu$F. The circuit is a development of Medlam and Oschwald's reflex voltmeter, whilst the A.577 valve is a modification of the M.L.4 valve, great care having been taken to reduce internal capacitances. The grid is brought out to a top cap on the glass bulb.

After much experience the Author considers these two designs to be the best available to date for all normal radio- and audio-frequency voltage measurements.

SECTION VIII

VALVES AND VALVE TESTS

THROUGHOUT this book considerable reference has been made to various types of thermionic valves. In general they have been used as oscillators, detectors, and amplifiers. It is therefore desirable to examine the properties of valves operating under such conditions, and to give details of normal types of test equipment. Following conventional practice the equations relative to the static conditions of the valve will be given.

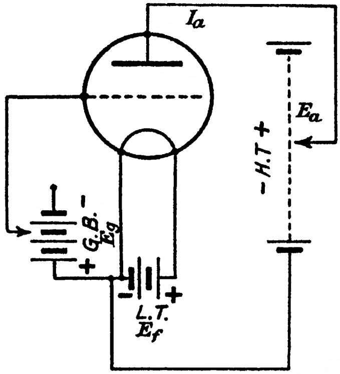


FIG. 50. BASIC CIRCUIT FOR STATIC TESTS ON VALVES

Referring to Fig. 50 for circuit and rotation, we derive for internal resistance (static),

$$\partial E_a / \partial I_a \qquad (95)$$

that is,

$$\frac{\text{Change in anode voltage}}{\text{Change in anode current}}.$$

For mutual conductance (static),

$$\partial I_a / \partial E_g \qquad (96)$$

that is,

$$\frac{\text{Change in anode current}}{\text{Change in grid voltage}}.$$

Such simple tests are applicable to every type of thermionic valve having three or more electrodes, and indeed they form the basis of the majority of commercial valve testers.

The laboratory engineer requires more information than is given by Eq. (95) and Eq. (96), and so an attempt will be made to compress into the small space of one section the most important characteristics of normal valves and also give details of useful valve test methods. Taking the case of a diode operating as a rectifier of alternating current: for a sinusoidal input voltage the deflection of a meter in the anode circuit is sensibly proportional to the average value of the applied voltage and so may be calibrated in terms of such. In the case of a triode operating as an anode bend (grid bias) rectifier, the change in anode current indicated by a suitable meter is then

$$\partial I_p = (G_m{}^2 e_0{}^2/4) \, . \, (\partial^2 I_p/\partial E_g{}^2) \qquad . \quad . \quad . \quad (97)$$

The factor G_m is the mutual conductance of the valve. In general terms it may be said the change in anode current is proportional to the square of the amplitude e_0 of the impressed grid voltage $e_g = e_0 \sin \theta$ and increases the anode current.

In the case of the grid rectifier we have—

$$\partial I_p = G_m R \partial I_g \qquad . \quad . \quad . \quad . \quad . \quad . \quad (98)$$

where G_m = mutual conductance;
∂I_g = change in grid current;
R = resistance of grid leak.

This method produces a decrease in anode current. In the case of triode oscillators of the self-rectifying type this equation is of particular interest, since with a known value of grid excitation, the anode current can be computed. It should be noted that although oscillators of the Dow type and others utilize tetrodes or pentodes, they are basically triode oscillators and Eq. (98) is applicable. This is not the case of the dynatron, however. By the combination of Eq. (97) and Eq. (98), the resultant change in anode current is determined when the valve is operating both as grid bias and grid leak detector. In general, this denotes incorrect circuit conditions. Here

$$\partial I_p = \frac{G_m{}^2 e_0{}^2}{4} \frac{\partial^2 I_p}{\partial E_g{}^2} - G_m R \partial I_g = \partial I_1 - \partial I_2 \qquad . \quad . \quad (99)$$

In the case of oscillatory circuits, it is often desirable to determine the value of the grid current. If the operating grid voltage $\partial I_g R$ is unknown, it can be calculated from the grid rectification effect ∂I_g, since R is known. The part of the grid voltage—grid current characteristic for which positive grid currents exist—for negative potentials can be computed. The kinetic energy of electrons when leaving the hot cathode without an external accelerating field is responsible for the passage of some electrons against a decelerating field of about 1 volt. According to Child, Langmuir, and Schottky, the space current I for electrodes of any form follows the 3/2 power law with respect to the voltage e. The space current passing from the hot cathode to the cold grid structure can be computed from the exponential law

$$I = I_s \varepsilon^{-pE} \quad . \quad . \quad . \quad . \quad . \quad . \quad (100)$$

where $p = 1/(8{\cdot}6 \times 10^{-5}\tau)$;

τ = absolute temperature in ° K.

Then for negative grid potentials E_g, we derive

$$I_g = \beta I_s \varepsilon^{-pE_g} \quad . \quad . \quad . \quad . \quad . \quad (101)$$

by utilizing Moeller's term for distribution factor

$$D = \frac{d(I_g/I)}{d(E_g/E_p)}.$$

These and the subsequent expressions used in connection with this subject are known as the "Moeller series" and are dealt with fully in *Die Elektronenröhren** and *Hilfsbuch für Elektrotechnik.*†

The electrons with a temperature velocity $v\tau$ may then be treated as though they were to leave the hot cathode with zero initial velocity, but so that the cathode has the potential

$$-E_\tau = \frac{I_m}{2q} v\tau^2 .$$

According to Eq. (100) the current contribution dI of such electrons is

$$\frac{dI_g}{dI} = D\,\frac{E_g - E_\tau}{E_p}$$

* F. Vieweg & Sohn, Brunswick, 1929. 1st Edition, p. 223.
† Julius Springer, Berlin, 1928. 1st Edition. p. 335.

and the grid current

$$I_g = \int_{\infty}^{-E_g} \frac{pD\,(E_y - E_r)}{E_p} I_s \varepsilon^{+pE_r}\, dE_r$$

$$= \frac{D}{pE_p} I_s \varepsilon^{-pE_g} \quad . \quad . \quad . \quad . \quad . \quad . \quad (102)$$

which for $D/pE_p = \beta$ checks Eq. (101).

Putting $\beta I_s = a$ for the grid current,

$$I_g = a_\varepsilon^{\ -p(E_c - e_g)}$$

where $e_g = e_0 \sin \theta$.

The average grid current is then

$$I_{av} = \frac{1}{T}\int_0^T I_g dt = a_\varepsilon^{\ -pE_c} \frac{1}{2\pi}\int_0^{2\pi} \varepsilon^{pe_0 \sin \theta}\, d\theta \quad . \quad (103)$$

The steady grid current without the application of an input voltage ($e_g = 0$) is

$$I_g = a\varepsilon^{-pE_c}$$

and the rectification effect in the grid current is computed from

$$\partial I_g = I_{av} - I_g = a\varepsilon^{-pE_c}\left[\frac{1}{2\pi}\int_0^{2\pi} \varepsilon^{pe_0 \sin \theta} - 1\right]$$

$$= a\varepsilon^{-pE_c}\left[\frac{1}{2\pi}\int_0^{2\pi} \sum_{m=1}^{m=\infty} (pe_0)^m \sin^m \theta d\theta\right]$$

$$= a\,\varepsilon^{-pE_c}\left[\frac{p^2 e_0^2}{4} + \frac{p^4 e_0^4}{64} + \ldots \frac{(pe_0/2)^{2m}}{(m\,!)^2}\right]$$

$$= Sa^{-Ec} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (104)$$

where S represents the series in parenthesis.

A much fuller examination of this problem and more elaborate treatise is given by Mr. August Hund in his book, *High-frequency Measurements,** from which the preceding has been summarized in brief terms.

In most valves the interelectrode capacitances are items which must be taken into account (excepting perhaps at low audio frequencies) since they effect the input and output impedances of the valve. When dealing with capacitances in this connection, it is perhaps more convenient to express the action of the valve in terms of electrical networks, that is,

* McGraw-Hill Publishing Co., 1st Edition.

"admittances," and after solution to use the reciprocal to enable the more customary impedance term for input and output to be employed. Thus, for the input admittance we have $\partial I_g/\partial E_g$, $\partial I_p/\partial E_p$ as output admittance, and $\partial I_p/\partial E_g$ as the mutual admittance, which is the effective slope of the work

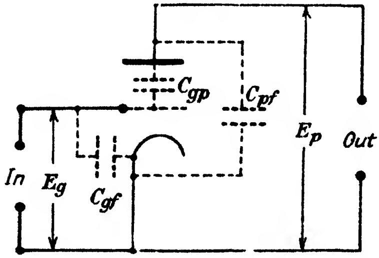


Fig. 51A. Valve Interelectrode Capacitances

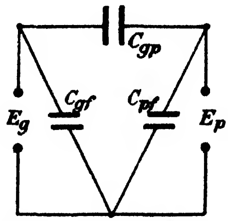


Fig. 51B. Equivalent Diagram of Fig. 51A

region of the valve characteristic. In the delta connection, as in Figs. 51A and 51B for these admittances, we derive

$$\left.\begin{array}{ll}
\dfrac{\partial I_g}{\partial E_g} = \dfrac{\partial I_1}{\partial E_g} & + j\omega(C_{gf} + C_{gp}) \\
\textit{Grid conductance } g_g & \textit{Input capacitance } C_g \\
\dfrac{\partial I_p}{\partial E_p} = \dfrac{\partial I_3}{\partial E_p} & + j\omega(C_{pf} + C_{gp}) \\
\textit{Plate conductance} & \textit{Output capacitance } C_p \\
\dfrac{\partial I_p}{\partial E_g} = \dfrac{\partial I_3}{\partial E_g} & + j\omega C_{gp} \\
\textit{Mutual conductance} & \textit{Mutual capacitance}
\end{array}\right\} \quad . \quad (105)$$

From which—

$$\left.\begin{array}{lll}
\text{Input admittance} & Y_g & = g_g + j\omega C_g \\
\text{Output admittance} & Y_p & = \dfrac{1}{r_p} + j\omega C_p \\
\text{Mutual admittance} & Y_m & = g_m + j\omega C_{gp}
\end{array}\right\} \quad . \quad . \quad (106)$$

The impedances are then

$$\left.\begin{array}{lll}
\text{Input impedance} & Z_g & = 1/Y_g \\
\text{Output impedance} & Z_p & = 1/Y_p \\
\text{Mutual impedance} & Z_m & = 1/Y_m
\end{array}\right\} \quad . \quad . \quad . \quad (107)$$

Having dealt briefly with certain important functions, it is now desirable to examine methods for the measurement of valve characteristics. The first valve tester to be described

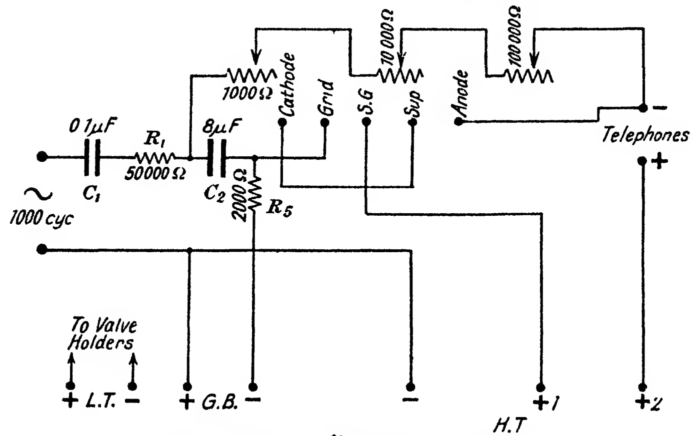


FIG. 52. "WIRELESS WORLD" VALVE TESTER

(Fig. 52) is a design due to the *Wireless World.** This measures to a good degree of accuracy the mutual conductance of a valve, and since this factor is, in effect, the goodness factor

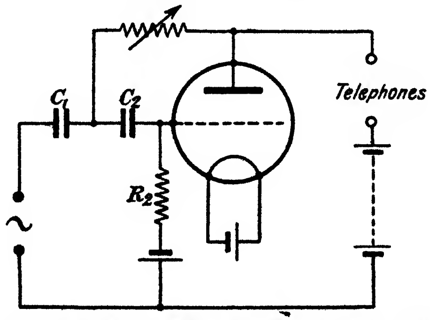


FIG. 52A. FUNDAMENTAL CIRCUIT OF FIG. 52

of the valve, this design can be said to fulfil a definite requirement. The range covered is from 0·001 mA./V. to 10·0 mA./V., and this covers all valves at present available. Instead of complex switching systems, three valve holders are provided

* 21st February, 1936.

covering all types of English bases, and independent connections are made to batteries, etc. This results in great utility and is a guarantee against obsolescence.

A valve tester which measures both mutual conductance and also impedance is a very interesting design due to the Sensitive

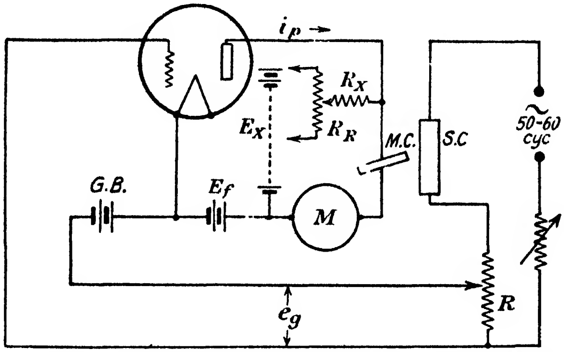


FIG. 53A. SENSITIVE INSTRUMENT CO. VALVE TESTER
Mutual conductance circuit

Instrument Corporation of New York, U.S.A. This is essentially a dynamic (as distinct from static) tester, making use of a dynamometer wattmeter. In this connection, knowing the ordinary wattmeter to be unsuitable for small power, it would seem the sensitive arrangement due to Dr. Drysdale and

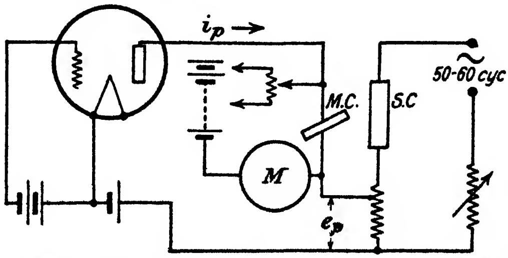


FIG. 53B. SENSITIVE INSTRUMENT CO. VALVE TESTER
Plate impedance circuit $r_p = e_p/i_p$

utilizing a Mumetal core would prove ideal. The Sensitive Instrument Co.'s design was fully described in *Electronics*,* whilst the circuit arrangements are shown in Figs. 53A and 53B. Referring to Fig. 53A, it will be seen that the signal voltage e_g

* December, 1934.

is obtained from a potentiometer in series with the field coil, thus ensuring that the phase of the signal voltage will correspond with the phase of the field voltage. As the valve causes a phase change of 180°, the current through the coil element will thus be in proper relationship.

The calibration on the meter scale (in micromhos, being the reciprocal of mA./V.) is equal to $10^3 i_c/e_g$, where e_g = signal voltage and i_c = meter calibration in milliamperes.

The value of the signal voltage $e_g = 10^3 i_m/S_m$, where i_m = full scale range of meter in milliamperes, and S_m = required range in micromhos.

The "bucking circuit" is used in conjunction with a centre-

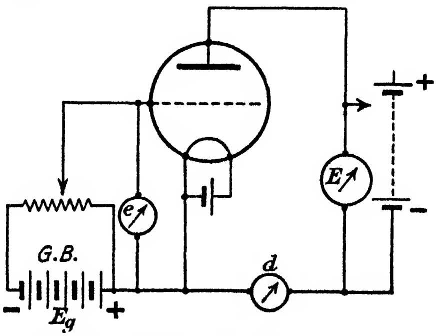


FIG. 54. STATIC AMPLIFICATION TEST

zero ammeter *M* to back off the standing current and so enable the full current change in the anode circuit to be indicated by the wattmeter.

In Fig. 53B the signal voltage e_p is obtained in the same way as the previous case, thus again ensuring proper phase relationships. The calibration in ohms is given by $r_p = e_p/i_c$, where e_p is the applied signal voltage and i_c the indicated alternating current reading. In addition to affording a means of "backing off" standing current, the bucking circuit is very useful in preventing stray d.c. fields which may cause false deflection of the wattmeter. The instrument in its combined form is successfully used in the measurements of r_p and g_m of valves ranging from power triodes to pentagrid mixer valves. A simple scheme for the determination of the static amplification of a valve is given in Fig. 54. A meter indicates the grid bias voltage *e*, and another the anode potential *E*. The ammeter *d* is used only as an equal deflection device, and need not be calibrated. The valve is tested with normal operating voltages, for which let grid bias $e = e_1$ and anode voltage $E = E_1$. These values are noted from the meter readings and the deflection of *d* observed. The anode voltage is then increased about 10 per cent, which value will be called E_2. This will cause increased deflection of *d*, which must be restored to its normal deflection

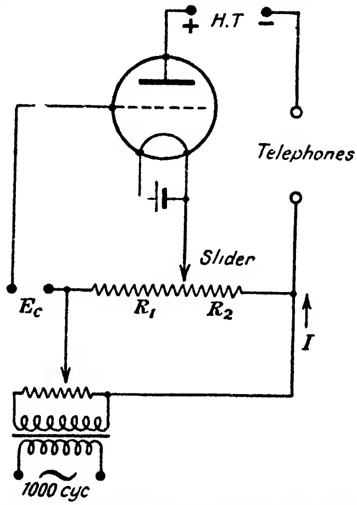


FIG. 55. MILLER BRIDGE FOR DYNAMIC AMPLIFICATION TESTS

by adjusting the grid bias, which will change the reading of *e*. Call the new reading e_2; then amplification factor

$$\mu = (E_2 - E_1)/(e_2 - e_1) \quad . \quad . \quad . \quad . \quad (108)$$

A method for the determination of the dynamic value of amplification factor is shown in Fig. 55. This is the bridge due to Mr. J. M. Miller, and is very widely used in valve factories. A current of 1 000 cyc. having really good waveform is supplied to the potentiometer R_1, R_2. Let e_g be the alternating grid voltage, then the alternating voltage drop across R_2 must be equal to $e_g R_2/R_1$ and is 180° out of phase with respect to the alternating anode voltage $e_p = \mu e_g$. The potentiometer slider is adjusted until silence is observed in the headphones, for which condition $e_g R_2/R_1 = \mu e_g$. The dynamic amplification factor is then

$$\mu = R_2/R_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (109)$$

The dynamic plate resistance of a valve can be determined in a Wheatstone bridge circuit as in Fig. 56, which contains a

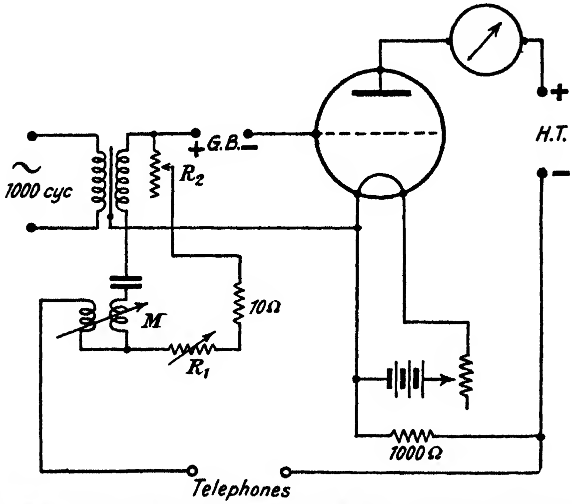


FIG. 56. BRIDGE FOR THE MEASUREMENT OF PLATE RESISTANCE

small mutual inductance m to enable compensation for any slight phase differences which make balancing uncertain. The valve is operated at its normal voltages, and R_1 set at the value used in the previous test (assuming the same valve is under test). R_2 and m are then adjusted until balance is indicated by silence in the phones. At balance

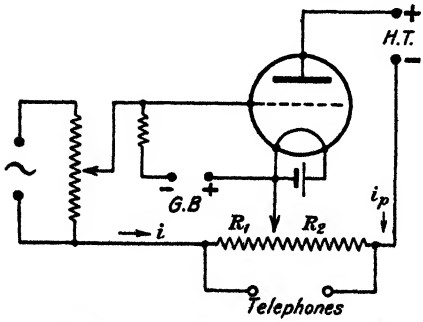


FIG. 57. DETERMINATION OF DYNAMIC MUTUAL CONDUCTANCE OF TRIODES

Plate resistance

$$r_p = 100R_2 \ . \quad (110)$$

Mutual conductance may be computed from the two previous tests; alternatively it may be measured dynamically in the circuit of Fig. 57. As in the previous examples, the potentiometer slider is adjusted until silence is observed in the phones. The alternating current components are i and i_p and such that the voltage drop across $(R_1 + R_2)$ is zero, i.e. $iR_1 = i_pR_2$. But $g_m = \partial i_p/\partial e_g$ and for the linear portion $i_p = g_m e_g = g_m R_3 i$ and

$$\text{Mutual conductance } g_m = R_1/(R_2R_3) \text{ mhos.} \qquad (111)$$

when the resistances are in ohms.

For tetrodes and pentodes the modified circuit of Fig. 58 is used. The conditions are as before, but this time R_1 is varied to obtain balance. $G_m = R_1/(R_2R_3)$. R_2 is usually made

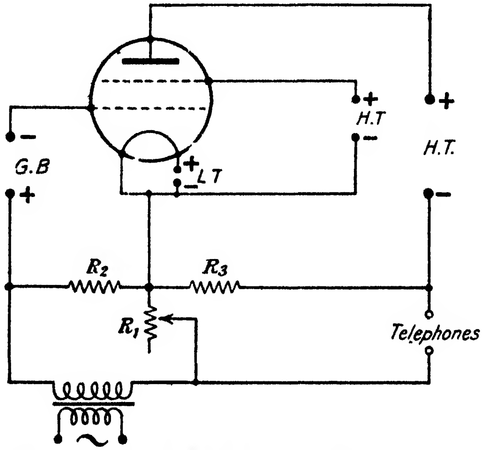


FIG. 58. BRIDGE FOR DYNAMIC MUTUAL CONDUCTANCE OF TETRODES AND PENTODES

1 000 ohms and R_3 100 ohms in these examples; thus in this latter case under such conditions

$$\text{Mutual conductance } G_m = 10R_1,\ R_1 \text{ being in ohms} \qquad (112)$$
(mhos.)

In these test circuits for μ, r_p, and g_m, the reader will have noticed the remarkable similarity of circuit arrangements. This enables the rapid testing of the three principal factors of a valve, and so can be regarded very much as a "universal valve bridge."

Although the simpler valves have been given as examples, it must be understood that the testers are equally suited to the more complex types. These types are usually composite arrangements consisting of two simple types in one envelope. They are best treated as such in tests. The only types not dealt with are diodes, in which class fall rectifier valves. The simplest and most satisfactory method is to test them under working conditions and observe their direct voltage and current output under variable load conditions. This same method can successfully be employed in diodes for h.f. use. The input voltage should be varied between about 3–20 volts alternating, and the corresponding rectified currents noted on a microammeter connected in series with a load resistance which can conveniently be 0·1 MΩ. Tests at 50 cyc. are quite satisfactory.

An attempt has been made in this short section to give a considerable amount of information in a concentrated form, and as far as possible guide the reader in thoroughly sound and practical channels. Most standard textbooks devote a considerable amount of space to the properties of valves, and as in other subjects often give such a large amount of information that it becomes difficult to make any specific choice. The test gear described in this section is thoroughly sound in theory, and in practice reliable, as the Author can confirm from personal experience. The best way to learn about valves is to use them, and above all diligently to read all current publications of merit. New valve types and circuits appear so regularly that even an attempt to describe the multitudinous varieties and their uses in a single book would fail, because the volume would be out of date almost as soon as it was published. Physical functions remain the same, and Chaffee's book* on thermionic tubes is an excellent volume for those readers desiring to make a study of valve design.

* McGraw-Hill Publishing Co., 1st Edition.

SECTION IX

BRIDGES

Bridges of various types are very valuable adjuncts to the well-equipped laboratory. In addition to the ordinary Wheatstone bridge for resistance measurements, the principles of which are well known to even first-year electrical students, other very useful bridge circuits are the Carey-Foster, Hay, Anderson, and Wien, etc.

The energizing source for a.c. bridges requires careful choice, and the Author holds the opinion that a push-pull valve oscillator operating at about 1 000 cyc., and fitted with a screened output transformer, is as good as anything. It must be remembered that harmonics in the source tend to mask the balance point of the bridge; therefore the very low harmonic content of a push-pull triode oscillator, such as is described in the section on "Audio-frequency Sources" is very desirable.

CAPACITANCE BRIDGES

Perhaps the simplest form is shown in Fig. 59, and in this design resistance measurements are also possible. The conditions for balance in this arrangement are satisfied by the simple relations

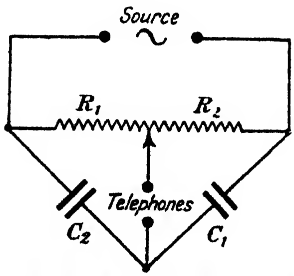


Fig. 59. Gott's Capacitance Comparison Bridge
(Developed in 1881)
The original circuit used a battery and key instead of an a.c. source

$$R_1 : 1/\omega C_2 :: R_2 : 1/\omega C_1.$$

Consequently,

$$R_1/R_2 = C_1/C_2;$$

thus by knowing the exact value of one of the condensers, say C_2 (and as the resistance is accurately known each time balance is attained), the unknown capacitance is easily determined from

$$C_1 = R_1C_2/R_2 \quad . \quad . \quad . \quad . \quad . \quad (113)$$

To avoid this calculation on every occasion, a direct calibration can be made, thus making a decidedly useful instrument.

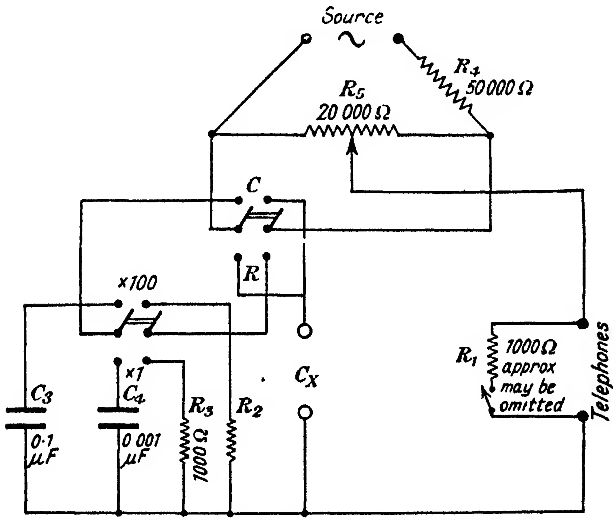


FIG. 60. RESISTANCE AND CAPACITANCE BRIDGE

Fig. 60 shows an elaboration of the bridge whereby capacitance and resistance measurements can be made. Two ranges are provided thus—

	Range I	Range II
R (ohms)	10 — 100 000	1 000 — 10×10^6 (10 MΩ.)
C (μμF.)	10 — 100 000 (0·1 μF.)	1 000 — 10×10^6 (0·001 μF.) (10 μF.)

The various components should be of good quality and preferably accurate to 1 per cent at least. The condensers

C_3 and C_4 should have a low power factor. The resistance R_5 should be a good quality potentiometer of linear characteristics. A suitable type is the British Electrical Resistance Co.'s 25-watt pattern. The scale shape is, of course, not linear, but nevertheless quite "open." This arrangement is admirably suited to rapid tests of batches of resistances and condensers such as are encountered in factories and repair depots. Since no provision is made for power factor balancing, complete silence at the balance point will not be obtained if the condenser under test has a high power factor. Under such conditions, adjust

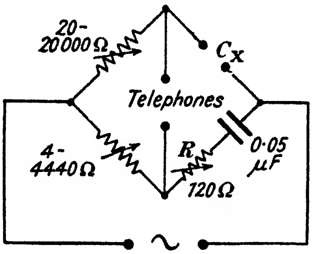


Fig. 61. Capacitance Bridge
(General Radio Co.)

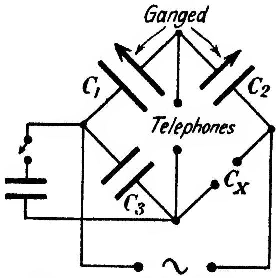


Fig. 62. Simple Capacitance Bridge
(Messrs. H. W. Sullivan Ltd.

for minimum signal strength. The resistance R_1 is included to shunt the phones in case the volume is too great under unbalanced conditions.

A useful capacitance bridge made by the General Radio Co. is shown in Fig. 61. Its range is from 0·001 μF. to 10 μF., the unknown capacitance being read off by means of the dial settings and ratio arms. The variable resistance R gives a rough calibration of the power factor of C_x. It will be seen that the standard reference capacitance is 0·05 μF. Fig. 62 shows a very useful bridge made by Messrs. H. W. Sullivan Ltd., constant scale reading accuracy being a notable feature. Errors do not greatly exceed 2 per cent at any part of the scale. The range is 0·00005 μF.–1·0 μF. in two steps, an additional fixed condenser being connected in parallel with C_3 on the high range. A further design calculated to meet the requirements of manufacturers and repairers is shown in Fig. 63. This arrangement is available as a kit of parts, complete with an etched calibrated scale having five ranges, from the Thordarson

Manufacturing Co., Chicago, U.S.A. The accuracy is of "workshop order," but careful selection of the condensers permits an accuracy of about 5 per cent. The use of the commercial power supply as the source is inexpensive, but unfortunately ordinary headphones do not possess a good response at 50 cyc.-60 cyc. A vibration galvanometer such as the inexpensive 50 cyc. pattern made by the Baldwin Instrument Co., and used instead of the phones will materially improve sensitivity. If one is prepared to go to this additional expense, it would certainly be worth the slight extra cost in obtaining

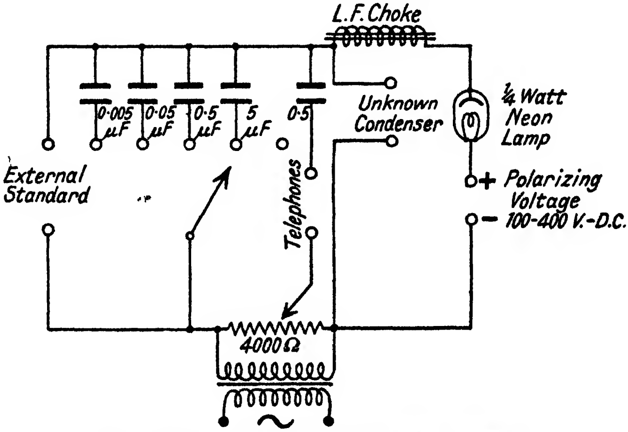


FIG. 63. THORDARSON CONDENSER TESTER

condensers of good quality, accurate to 1 per cent or better. Such improvements will make the bridge accurate to about 2 per cent. Provision for measuring high capacitance electrolytic condensers is made, and this should prove quite useful in many cases. The polarizing voltage should be above 100 volts, but not greater than 400 volts. The neon tube (¼ watt rating) and choke prevent damage to low voltage electrolytic condensers. A normal 1 000 cyc. source may be used instead of the commercial power supply, whence phones are then suitable.

The ranges of this instrument are 0·001 μF.–0·05 μF., 0·01 μF.–0·5 μF., 0·1 μF.–5 μF., and 1 μF.–50 μF. The neon lamp also serves to indicate condenser leakage.

An arrangement due to Mr. D. C. Gall, of Messrs. H. Tinsley & Co., having inherently greater accuracy than the previous types described, is shown in Fig. 64, the whole unit being self-contained and conveniently portable. The instrument would appear to be a development of Maxwell's method of determining capacitance in terms of resistance with the aid of a commutator regularly repeating the process of charge and discharge of a condenser. In the Gall capacitance meter the vibrating contact *B* is driven by a relay magnet *H* on a source *A* of a.c. supply. Alternatively a tuning fork operated from a d.c. source may be used. The charging voltage is provided by the battery *E* and may conveniently be about 24 volts. In operation the instrument is standardized on the standard

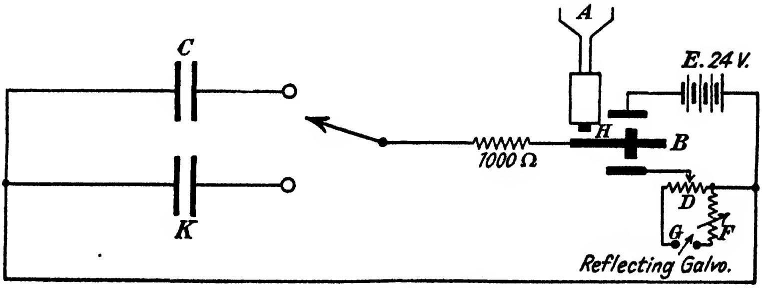


FIG. 64. CAPACITANCE METER
(Gall design: Messrs. H. Tinsley & Co.)

condenser *K* by adjustment of the variable resistance *F* in the loop shunt circuit to give its correct reading on the self-contained reflecting galvanometer scale. Any other desired range is obtained by the adjustment of *D*, after which the unknown condenser is switched in circuit. The long scale of the galvanometer greatly facilitates accurate readings on any of the ranges which are as follows: 0·01 μF., 0·02 μF., 0·05 μF., 0·1 μF., 0·2 μF., 0·5 μF., 1 μF., 2 μF., 5 μF., and 10 μF. This design was described by Mr. Gall in the *Journal of Scientific Instruments*.*

After this brief survey of general purpose instruments, it is now desirable to review types which are essentially of the laboratory class. Reference to the various books on bridge measurements probably leaves the reader in a quandary as to the most suitable type for a given job, in view of the tremendous number of arrangements described. With this in mind, arrangements will be described which, with simple modifications, can be used for a wide range of measurements of *C*, *L*, and *M*.

* Vol. 10, 1933, p. 326.

Also, mention will be made of composite bridges which are sometimes styled *universal bridges*. With these, however, the standard of accuracy is rarely as good as the "single circuit" arrangements.

In the Author's opinion, for the determination of capacitance and power factor with the aid of mutual inductance, the Heydweiller modification of the Carey-Foster bridge is the best. Fig. 65 shows this in its most general form with the incorporation of the Wagner ground (in broken lines); this

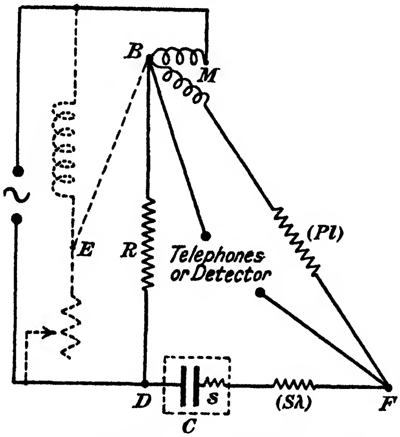


FIG. 65. CAREY-FOSTER BRIDGE
Heydweiller's modification: Wagner ground shown in broken lines

latter item, although not essential, is advantageous for eliminating earth capacitance effects. The components required to assemble the bridge is as follows—

- Campbell mutual inductance (M);
- Carey-Foster auxiliary box containing the resistances R and P_l;
- Inductionless decade resistance box, S 4-dial pattern, graduated in tenths, units, tens, and hundreds, giving a total resistance of 1 111 ohms;
- A.c. source; and
- Detector.

The Campbell mutual inductance has a range from − 0·003 to + 11 mH., the scale being calibrated from − 3 μH. to + 105 μH. over an approximate length of 30 cm., whilst the provision of the two decade dials extends the effective length up to 3 300 cm. Full details of the construction of this unit are given in two papers by Mr. Campbell,* whilst a commercial product is made by the Cambridge Instrument Co. The Carey-Foster auxiliary box† is arranged so that the resistances may be interpreted as microfarads or micromicrofarads directly with the simple use of multipliers. Table XXXV shows the maximum range of capacitance which may be measured using the aforementioned mutual inductance. The decade resistance box should have a very low inductance, and for this purpose Ayrton Perry windings are used in the Cambridge Instrument Co.'s product, a table of residual inductances being given in Table XXXVI. In operation, when an initial balance is obtained we have

$$10^6 M/C = PR - \omega^2 lL + \omega^2(l + \lambda)M$$

and $$M(S + s) = R(L - M) + Pl.$$

From these two equations C and $(S + s)$ can be determined, S being known: s and therefore the power factor of ωC_s can be found

$$C = 10^6 M/PR, \quad . \quad . \quad . \quad . \quad . \quad (114)$$

and when ω is low l and λ can usually be neglected, whence

$$S + s = R(L - M)/M \quad . \quad . \quad . \quad . \quad (115)$$

Taking a practical example—

TABLE XXXV

CAREY-FOSTER AUXILIARY BOX USED WITH CAMPBELL VARIABLE INDUCTANCE (INDUCTOMETER)

(Cambridge Instrument Co. Ltd.)

Values (in ohms) of		Multiplier = $1/PR$	Max. Capacitance Measurable in Microfarads (μF.)
Value of P	Value of R		
100	10	1×10^{-3}	11·1
200	10	5×10^{-4}	5·55
100	50	2×10^{-4}	2·22
200	50	1×10^{-4}	1·11
200	100	5×10^{-5}	0·555
500	50	4×10^{-5}	0·444
500	100	2×10^{-5}	0·222
100	1 000	1×10^{-5}	0·111
200	1 000	5×10^{-6}	0·0555
500	1 000	2×10^{-6}	0·0222
1 000	1 000	1×10^{-6}	0·0111

* *Proc. Phys. Soc.* Vol. 21, 1908, p. 69; and *Phil. Mag.* Vol. 15, 1908, p. 155.

† G. Carey-Foster, *Phil. Mag.*, Vol. 23, 1887, p. 121.

TABLE XXXVI

RESIDUAL INDUCTANCE OF CAMBRIDGE INSTRUMENT CO. DECADE RESISTANCE BOX AT 1 000 CYCLES

Resistance (ohms)	Residual Inductance (μH.)
Zero	+ 0·53
1	+ 0·5
10 × 1	+ 0·79
10	+ 0·8
10 × 10	+ 0·85
100	+ 0·5
10 × 100	− 0·5
1 000	+ 0·4

Let $P = 1\,000$ ohms;
$R = 10$ ohms.

(*a*) To find the capacitance of the connecting leads.
M is found to be 9.

Then lead capacitance

$$= 9/(1\,000 \times 10) = 0{\cdot}0009\ \mu\text{F}.$$

(*b*) To find the capacitance of a condenser under test.
Assume M to be 930 μH.

Then

$$C = (M/PR) - \text{capacitance of leads}$$

$$= \left[\frac{930}{1\,000 \times 10} - 0{\cdot}0009\right] \mu\text{F.} = 0{\cdot}0921\ \mu\text{F.}$$

With this bridge, capacitances from less than a picofarad to several microfarads can be measured. Power factor measurements can be made without reference to any standard condenser.

Two further modifications of this bridge have been developed by Mr. Campbell whereby a somewhat greater degree of accuracy is attained.*

Full details of the Wagner ground will be found in articles by K. W. Wagner and D. W. Dye in *Elektrische Zeitschrift* and *The Electrician*.† In the second part of this section it will be

* A. Campbell, *Proc. Phys. Soc.*, Vol. 43 (1931), p. 564; Vol. 29 (1917), p. 347.

† *Elekt. Zeitschrift*, Vol. 32 (1911), p. 1001; *Electrician*, Vol. 68 (1911), p. 483; and D. W. Dye, *N.P.L. Report*, 1920, p. 62; *Electrician*, Vol. 87 (1921), p. 55.

seen that a modified Carey-Foster bridge is suitable for inductance measurements.

Wien Bridge. This design, due to M. Wien, is considered by some as perhaps the most convenient method for the measurement of capacitance and power factor. In this respect, Mr. Nottage in his book, *Calculation and Measurement of Inductance and Capacity,* expresses such sentiments. The circuit is shown in Fig. 66.

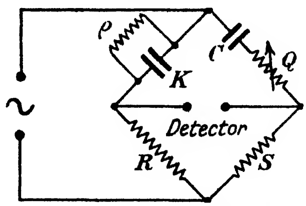


FIG. 66. WIEN BRIDGE

The resistance ρ in parallel with the condenser under test K represents the internal loss of the condenser. C must be a very high-grade condenser having negligible losses. When accurate measurements are desired, the power factor of C must be taken into account.

Balance is obtained by adjusting Q and the decade resistance S. Then,

$$K = \frac{S}{R} \frac{C}{1 + \omega^2 C^2 Q^2} \quad . \quad . \quad . \quad . \quad (116)$$

and

$$\rho = \frac{R}{S} \frac{1 + \omega^2 C^2 Q^2}{\omega^2 C^2 Q} \quad . \quad . \quad . \quad . \quad (117)$$

where $\omega = 2\pi \times$ frequency in cycles per sec.; and

C and K are measured in farads.

Since frequency is involved in these equations, the frequency of the source must be accurately known. Checks against a 1 000 cyc. tuning fork are usually good enough for normal work.

EXAMPLE.

$$R = S = 1\,000\ \Omega.$$
$$C = 0{\cdot}025\ \mu\text{F}.$$
$$g = 150\ \Omega.$$
$$f = 1\,000\ \text{cyc}.$$

From Eq. (116)

$$K = 0{\cdot}025\ \mu\text{F. (neglecting } \omega^2 C^2 Q^2)$$
$$= 0{\cdot}025 \times 10^{-6}\ \text{F}.$$

From Eq. (117)

$$\rho = \frac{1 + [(2\pi \times 1\,000) \times (0{\cdot}025 \times 10^{-6}) \times 150]^2}{[(2\pi \times 1\,000) \times (0{\cdot}025 \times 10^{-6})]^2 \times 150}$$

The Wien bridge is not suitable for inductance measurements.

Schering Bridge. This design, first announced in *Zeitschrift für Instrumententeknik** is widely used in various forms for the measurement of capacitance, power factor, and losses in dielectrics. In this latter item, the Schering bridge is becoming increasingly useful in view of present-day developments in low-loss insulators.

C_1 is the condenser or dielectric thereof under test. C_2 is a calibrated variable condenser of about 0·001 μF. maximum capacitance. R_3 and R_4 are equal value inductionless resistances

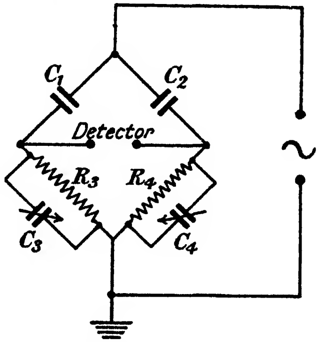


FIG. 67. SCHERING BRIDGE

which can conveniently be two decade resistance boxes, C_3 has a value of about 250 $\mu\mu$F. and serves principally to attain initial balance of the bridge, whilst C_4 has a maximum capacitance of about 0·001 μF. and is calibrated directly in values of power factor at a given frequency (usually 800 cyc. or 1 000 cyc.).

OPERATION. If C_1 is the capacitance of the condenser under test, and S the equivalent series resistance causing the loss,

Then $$C_1 = \frac{C_2R_4}{R_3(\omega^2C_2C_3R_4S + 1)} \quad . \quad . \quad . \quad (118)$$

and $$S = R_3(C_4/C_2 - C_3/C_1) \quad . \quad . \quad . \quad (119)$$

Since the factor $\omega^2C_2C_3R_4S$ is usually small compared with unity, for normal purposes it is justifiable to write

$$C_1 \doteqdot C_2(R_4/R_3),$$

and as $$R_3 = R_4, C_1 \doteqdot C_2 \quad . \quad . \quad . \quad (120)$$

* Vol. 40 (1920), p. 124.

The power factor

$$\cos\phi = \tan\theta = \omega C_1 S \quad . \quad . \quad . \quad (121)$$

(θ normally being very small);

i.e. $\tan\theta \doteq \omega R_4 C_4 - \omega C_3 R_4$

$$= \omega R_4(C_4 - C_3).$$

Power loss in dielectric

$$= VI \times \text{power factor},$$

where V = r.m.s. value of working voltage;
I = r.m.s. value of charging current;

i.e. power loss = $VI \tan\theta$.

But $I = CV\omega$.

Therefore power loss = $V^2\omega C \tan\theta$.

The Cambridge Instrument Co. manufacture a bridge of this type. Another very useful arrangement is made by Messrs. H. W. Sullivan Ltd. Capacitances from 1 $\mu\mu$F. approximately up to 1 000 μF. are directly indicated, whilst power factor is obtained by a simple division of two dial readings. In using a Schering bridge the condenser under test should be subjected to an oscillating potential of about 200 volts, this usually being at telephonic frequency.

To summarize, the Author considers the Carey-Foster arrangement to be most suitable for capacitance measurements, and as will be seen later, the component parts used in its construction are also suitable for inductance measurements. Also the decade resistance boxes are always very useful for other laboratory work. It is truly a "utility" arrangement. For dielectric measurements the Schering bridge is most suitable.

INDUCTANCE BRIDGES

Carey-Foster Bridge. Inductances from about 1 mH. upwards are measured in the *P* arm. Conditions of balance are

$$L' = \frac{(S_2 - S_1)}{R} M \quad . \quad . \quad . \quad . \quad . \quad (122)$$

where L' = unknown inductance;
S_2 = value of resistance to balance bridge with inductance
S_1 = value of resistance to balance bridge without inductance
} in ohms.

EXAMPLE.

$$S_2 = 5\ 640\ \Omega.$$
$$S_1 = 240\ \Omega.$$
$$M = 150\ \mu\text{H}.$$
$$R = 10\ \Omega.$$

Then $$L' = \frac{(5\ 640 - 240) \times 150}{10}\ \mu\text{H}.$$

Campbell-Heaviside Bridge (Equal Ratio). Using the same inductometer as in the Carey-Foster bridge, self-inductances from 0·2 μH. up to 22 000 μH. can be conveniently measured on this bridge. It is therefore very suitable for measurements on coils used in radio receivers. The connections are shown in

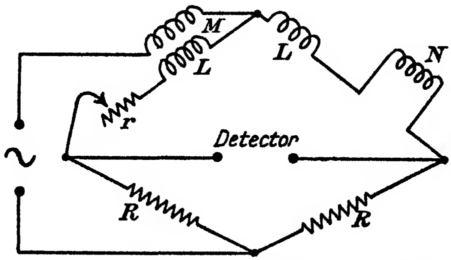


FIG. 68. CAMPBELL-HEAVISIDE BRIDGE

Fig. 68. *R*, *R* are ratio boxes having resistances of the values: 90, 10, 900 : 100, 990, 10, corresponding to multiples of 10, 10, and 100. *r* is a decade resistance box of 11 111 ohms in series with a constant inductance rheostat having a maximum range of about 0·5 ohm. This operates as a fine adjustment.

OPERATION. Short circuit *N* and obtain initial balance by adjustment of *r* and *M*. Call the readings r_1 and M_1. Bring *N* (the unknown) into circuit and readjust *r* and *M* for balance. Call these readings r_2 and M_2; then

self-inductance $N = 2(M_2 - M_1)$

effective resistance of $N = r_2 - r_1$

Errors due to lead capacitances are cancelled by differential readings.

Anderson Bridge. Mr. W. H. Nottage considers this to be perhaps the most useful bridge for inductance measurements. Herein inductance - is compared against capacitance with independence of frequency. The circuit is shown in Fig. 69. The bridge is of simple character requiring only a standard condenser of 1 μF. in addition to ratio boxes for P and R, a decade box for r (11 111 ohms), and a fixed resistance Q which can conveniently be 1 000 ohms. LS is the inductance and effective resistance of the coil under test. Adjust R and r to balance, then

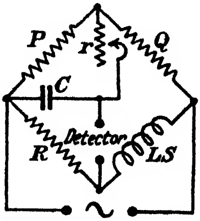


FIG. 69. ANDERSON BRIDGE

$$L = 10^{-6}C(Rr + Sr + PS)$$

$$SP = QR,$$

or if $P = Q$, then

$$L = 10^{-6}CR(2r + P) \quad . \quad . \quad . \quad . \quad (123)$$

$$S = R \quad . \quad . \quad . \quad . \quad . \quad . \quad (124)$$

where L is in henries;
C is in microfarads.

EXAMPLE.

$P = 1\ 000\ \Omega.$;
$Q = 1\ 000\ \Omega.$;
$C = 1\ \mu$F.;
$R = 10\ \Omega.$;
$r = 300\ \Omega.$

From Eq. (123), $L = (10^{-6} \times 1 \times 10)\,(2 \times 300 + 1\ 000)$.
From Eq. (124), $S = R = 10\ \Omega.$

This bridge is not particularly suitable for low inductance measurements.

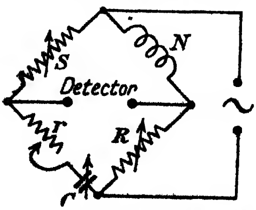


FIG. 70. HAY'S BRIDGE

Hay's Bridge. This arrangement is very well suited to the measurement of large inductances such as iron-core chokes, transformers, etc. It is shown in its original form in Fig. 70, whilst Fig. 71 shows a very worthwhile improvement which permits the measurement of iron-core coils carrying direct current. This latter improvement was suggested to the Author by Mr. A. Wilkinson who, the Author believes, is responsible for the modified arrangement.

Data for Components.

$R = 1\ 111$ ohms (decade box);
$r = 11\ 111$ ohms;
$S = 11\ 111$ ohms;
$C = 0{\cdot}999\ \mu$F. (decade condenser).

Adjust R and r for balance; then

$$L = RSC/(1 + \omega^2C^2r^2)$$
$$= RSC \text{ approximately} \quad . \quad . \quad . \quad (125)$$

Effective resistance of N

$$= RS\omega^2C^2r(1 + \omega^2C^2r^2)$$
$$= LC\omega^2r \text{ approximately} \quad . \quad . \quad . \quad (126)$$

Hay-Wilkinson Bridge. The conditions for balance and calculations are precisely the same as for the normal Hay's bridge. A few notes on Mr. Wilkinson's modifications are perhaps justified. The condensers C_1, C_2, C_3, and C_4 (Fig. 71) are d.c. stoppers and conveniently of 1–4 μF. capacitance. They must be chosen so as to present a negligible impedance at operating frequency. This may be 50 cyc., whence a vibration galvanometer is the best form of detector, or at telephonic frequency, whence headphones are suitable. The arm R is a decade box when no d.c. is being carried by the coil under test, but when d.c. flows, this is replaced by a normal tubular slide rheostat. This is preferable to passing d.c. through a decade box. The resistance of the rheostat should be known in effective resistance measurements. The rheostat R_1 serves to control the direct current flow through the coil under test. The two chokes L_1, each side of the battery, serve to prevent a.c. taking this path. Values of about 30 henries are indicated, although this is not a critical value. A valve voltmeter of the grid rectifier type, having a

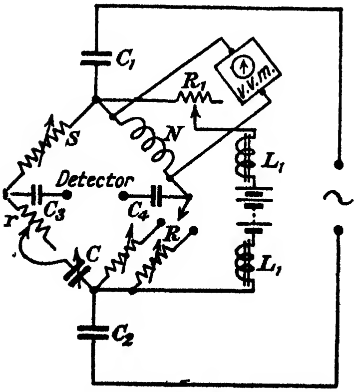


FIG. 71. HAY-WILKINSON BRIDGE

range up to about 5 volts, indicates the alternating volts across the choke. Variation of this voltage, of course, causes a change in inductance.

It should be particularly observed that the condenser *C* in both arrangements must be of high quality and low losses.

The Author considers this arrangement the best for measurements on large coils with or without magnetizing current.

COMPOSITE BRIDGES

Bridges under this heading are often termed *impedance bridges*, and are so arranged that with a minimum amount of apparatus, inductance, capacitance, and resistance may be measured on one instrument. This, of course, is often a great convenience, but as one seldom gets something for nothing in this world, the price to be paid for the added convenience is a somewhat lower order of accuracy than with the specialized bridge circuits.

Perhaps the best example of a composite bridge is the General Radio Co.'s Type 625A impedance bridge, whilst the G.R. Type 293A universal bridge forms another excellent example. The Type 625A bridge is in effect a skeleton bridge circuit, one arm containing a logarithmic rheostat whilst the other three arms are brought out to pairs of terminals, to which external standards as may be required and the unknown quantity are connected. Useful values for these external standards are 10 000, 4 000 and 100 ohms, 1 μF. and 0·01 μF. With these standards the range of the instrument is—

R. 1 ohm–1 MΩ.;
L. 100 μH.–100 henries;
C. 100 μμF.–100 μF.

Additional external standards will of course increase the range of the instrument. The logarithmic rheostat has a maximum range of 10 000 ohms. Table XXXVII gives the various combinations corresponding to the circuits employed and the range of *L*, *C*, and *R* measurable with each given combination. It will be seen that both Hay's and Maxwell's bridges are used for inductance measurements. The use of the Maxwell bridge permits easier determination of the factor *Q* ($Q = \omega L/R$), since the term involving frequency in Hay's bridge is obviated. The circuits used are given in Figs. 72 (*a*) to (*d*). The General Radio

Co.'s Type 650A bridge is developed from the basic arrangements of Type 625A, and has the advantage of not requiring external standards. The makers state the accuracy to be 1 per cent for capacitance and resistance, and 2 per cent for induc-

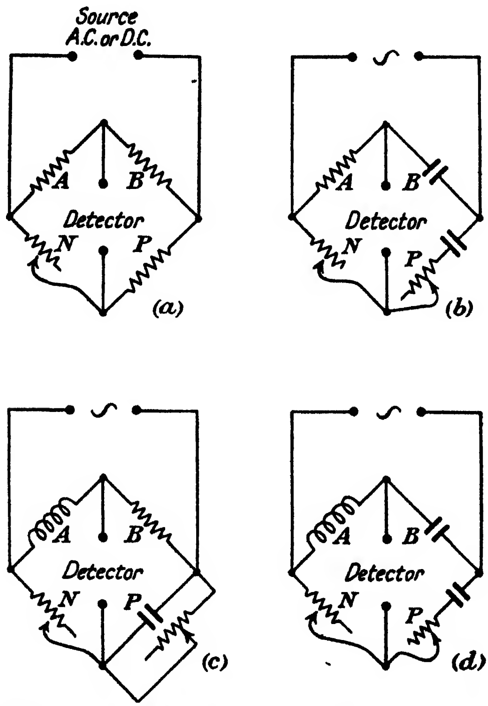


FIG. 72. POSSIBLE COMBINATIONS WITH GENERAL RADIO CO. TYPE 625A IMPEDANCE BRIDGE

(a) Wheatstone bridge; resistance measurements
(b) Wien bridge; capacitance measurements
(c) Maxwell bridge; inductance measurements (Q measurements, any value)
(d) Hay's bridge; inductance measurements (Q values, 10)

tance, although at the lower values of each type these figures are 5 per cent and 10 per cent respectively.

The Type 293A Universal Bridge is in effect a fundamental circuit consisting of three resistance arms and suitable terminal arrangements to facilitate the setting up of various types of bridges. *L*, *C*, and *R* can be measured up to about 50 kc.

The resistance arms consist of two decades of 11 111 ohms (4-dials), and the third being a step resistance as multiplier

TABLE XXXVII

POSSIBLE RANGE OF MEASUREMENTS WITH GENERAL RADIO CO. TYPE 625A IMPEDANCE BRIDGE

(Unknown Ranges in Bolder Type)

A Arm	*B* Arm	*P* Arm
100 ohms	10 000 ohms	**1 000 ohms to 1 000 000 ohms**
1 000 "	10 000 "	**100 " " 100 000 "**
10 000 "	10 000 "	**10 " " 10 000 "**
10 000 "	1 000 "	**1 " " 1 000 "**
10 000 "	100 "	**0·1 " " 100 "**
10 000 "	10 "	**0·01 " " 10 "**
10 000 "	1 "	**0·001 " " 1 "**

A Arm	*B* Arm	*P* Arm
100 ohms	**0·1 μF. to 100 μF.**	1 μF.
1 000 "	**0·01 " " 10 "**	1 "
10 000 "	**0·001 " " 1 "**	1 "
1 000 "	**100 μμF. " 0·1 "**	0·01 "
10 000 "	**10 " " 0·01 "**	0·01 "
10 000 "	**1 " " 1 000 μμF.**	1 000 μμF.

A Arm	*B* Arm	*P* Arm
100 mH. to 100 henries	10 000 ohms	1 μF.
10 " " 10 "	1 000 "	1 "
1 " " 1 "	10 000 "	0·01 "
100 μH. " 100 mH.	1 000 "	0·01 "
10 " " 10 "	100 "	0·01 "
1 " " 1 "	10 "	0·01 "

arm, having resistance values of 1, 10, 100, 1 000, and 10 000 ohms. Table XXXVIII gives the various arrangements used for a wide variety of measurements.

TABLE XXXVIII

POSSIBLE ARRANGEMENTS WITH GENERAL RADIO CO. TYPE 293A UNIVERSAL BRIDGE

L = self-inductance; *M* = mutual-inductance; *R* = Resistance; *C* = capacitance; *f* = frequency

Bridge	Unknown to be Measured	Known Bridge Elements
Impedance . . .	*R*	*R*
	L	*L* and *R*
	C	*C* and *R*
Grover	*C*	*C* and *R*
Schering	*C*	*C* and *R*
Maxwell	*L*	*C* and *R*
Owen	*L*	*C* and *R*
Hay	*L*	*C*, *R*, and *f*
Resonance	*L*	*C* and *f*
Wien	*C*	*R* and *f*
	L	*R* and *L*
	f	*C* and *R*
Anderson	*L*	*C* and *R*
Campbell	*L*	*M* and *R*
Carey-Foster . . .	*C*	*M* and *R*
	M	*C* and *R*

RADIO-FREQUENCY BRIDGES

The various bridge arrangements so far described have been for operation on direct current or alternating current at power or telephonic frequency. With the development of the radio art, the limitations imposed by testing components, destined for radio-frequency operation, at telephonic frequencies have manifested themselves in several ways. Amongst these a notable instance is the newly developed sprayed mica ceramic fixed condenser. At telephonic frequencies, samples tested by the Author have shown high power factors, but at frequencies of the order of 1 Mc. the power factor has been quite low, and there are reasons to suppose a further improvement takes place at still higher frequencies. Nevertheless, there is a limit to this improvement with frequency, and at 500 Mc. certain types are particularly inefficient. Contrary to the beliefs of many people, the inductance of a good tuning coil is but slightly different at 1 000 cyc. as compared with 1 000 000 cyc. Tests conducted by the Author with a few coils, first at 1 000 cyc. and then at 1 Mc. showed the difference to be but a few parts in 10 000. However, the effective resistances of the coils at the two frequencies was very different. Likewise, capacitance measurements at 1 000 cyc. were in very close agreement with those at 1 Mc., the error observed again being a few parts in 10 000, though as in the previous case, the power factor was different.

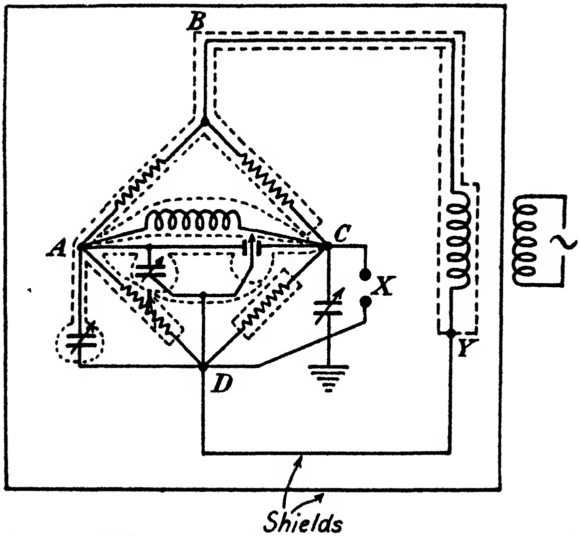


Fig. 73. Campbell-Colpitts Radio-frequency Bridge

Each element is shielded as indicated by the broken lines. The shields are connected at common points and to the main box shield at Y

In describing radio-frequency bridges, the Author finds from experience that up to the present time no r.f. bridge yet commercially produced is comparable in versatility, accuracy, or robustness and reliability with the various telephonic frequency bridges. In due fairness, it must be said that really intensive work on r.f. bridges dates back about ten years from its commencement, whereas arrangements like Gott's bridge go back over half a century, and this was by no means the first design. An early arrangement was the Campbell-Colpitts capacitance and conductance bridge described in the *Bell System Technical Journal*,* the circuit diagram of which is given in Fig. 79. The authors of the above paper specify a frequency

* Vol. 7, 1928, p. 70.

range of from 30 cyc. to 100 kc. A later paper in this same journal* states that a properly screened bridge should be effective up to 2 Mc. The resistance arms are the most uncertain elements in a r.f. bridge at frequencies of 1 Mc. or higher, since their effective resistances and reactances may differ by a considerable comparative amount. When capacitances below 100 $\mu\mu$F. are to be measured, the uncertainties of the bridge

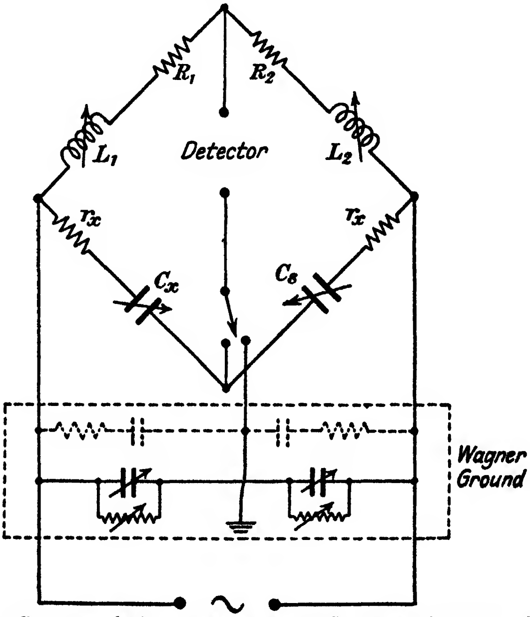


FIG. 74. STRATTON'S RADIO-FREQUENCY BRIDGE (ORIGINAL DESIGN)
Shielding not shown

become more serious as the capacitance involved may be comparable with the total stray capacitances of the bridge. Formidable as these items may appear, it will be seen they have been materially overcome in recent designs. Fig. 74 shows Stratton's R.F. Bridge for the measurement of the resistance and capacitance of condensers. Full details appeared in the *Journal of the Optical Society of America*.† The bridge will be recognized as an impedance bridge, but instead of balancing the bridge by means of a resistance R in series with C_s to correct

* *B.S.T.J.*, Vol. 8, 1929, p. 560.
† Vol. 13, 1930, p. 471.

phase differences, Stratton uses variometers of some 4 mH. maximum inductance. He shows that when the bridge is balanced by the variometers, the power factor of the condenser C_x (which, incidentally, is also the tangent of the angle of phase difference) is

$$\omega L_1/R_1 - \omega L_2/R_2.$$

With this arrangement the frequency limit is about 500 kc. and is principally set by the variometers, the self-capacitance of which reduces the apparent inductances as the frequency

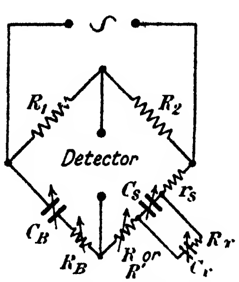


Fig. 75. Stratton's Modified Radio-frequency Bridge

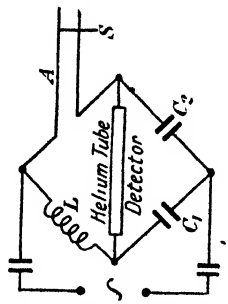


Fig. 76. Watson's Method (1909)

increases. Very comprehensive shielding of all elements, and complete box screening of each arm, is featured. In a modified design (Fig. 75) Stratton dispensed with the variometers. With the standard condenser C_s, and R set to a suitable value, balance is obtained by adjusting R_B and C_B. Their values need not be known. The test condenser C_x with its equivalent loss resistance R_x is added in parallel with C_s, and this is reduced to a value C_s', R being changed to R' until a second balance is attained. Then

$$R_x = (R - R')\,(C_s/C_x)^2 \quad . \quad . \quad . \quad (127)$$

and $$C_s - C_s' = C_x \text{ (very nearly)} \quad . \quad . \quad . \quad (128)$$

From a point of historical interest the Watson bridge for the measurement of inductance at radio frequency is given. The detector is a helium tube (Fig. 76). By observing the distance S has to move along the parallel wires A (which form the standard inductance) to restore balance when an unknown inductance L is introduced into the adjacent arm, L is calculated from the known value of the equivalent portion of A. The scheme was described in the *Electrician.**

Turning now to present-day practice, the General Radio Co.'s Type 516C R.F. Bridge is an excellent example of modern design. The circuit diagram is given in Fig. 77. To appreciate its salient features it is desirable to examine an earlier design

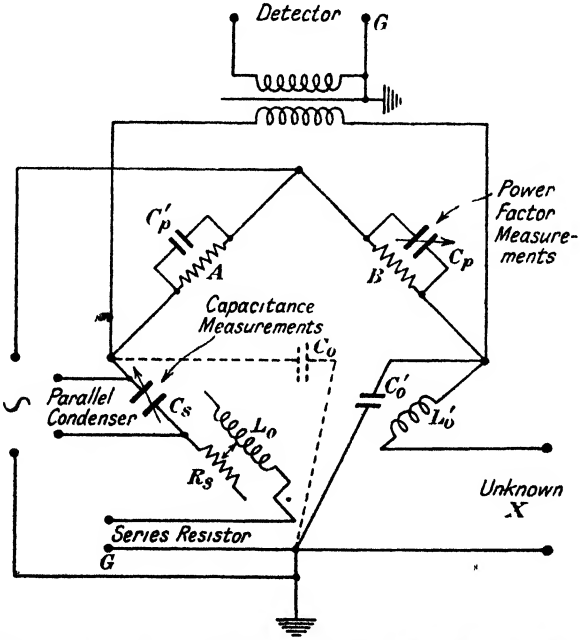


FIG. 77. GENERAL RADIO CO. 516C RADIO-FREQUENCY BRIDGE

of the same Company and compare the two models. The earlier model is styled 516A. Herein we find an air core shielded output transformer being reasonably efficient over the range 10 kc.–5 Mc. By placing it in the output of the bridge, it carries no current when the bridge is balanced, and thereby no external field. The ratio arms are 100 ohms each, and are wound on thin mica cards to reduce their residual inductance to small limits. The balance condenser has a maximum value of 0·001 μF. The variable resistance for power factor balance is a three-dial decade box having steps of 10, 1, and 0·1 ohms

* March 5th, 1909, p. 809.

per dial. The maximum inductance of the box is given as 1·1 μH., and minimum 0·2 μH.

Superficially one may get the impression upon examining Type 516A that almost everything possible had been done to render it suitable for radio-frequency work. Yet upon comparison with Type 516C, it becomes apparent that radical changes have been made, and these effect a considerable improvement.

The decade resistance box in the earlier model is the cause of considerable errors, primarily because the resistance is not pure, and secondly because the inductive component of this impure resistance is a quantity varying from 0·2 μH. to 1·1 μH., dependent upon the setting of the decade. Small though this inductive component may seem, it has an appreciable reactance at radio frequencies, and since this latter quantity changes by dial settings an error in capacitance measurements results. Since the effective capacitance of a condenser in series with an inductance is given by the expression

$$C' = C/(1 - \omega^2 LC) \quad . \quad . \quad . \quad . \quad (129)$$

where C' = effective capacitance in farads;

C = capacitance of condenser in farads;

L = inductance in henries;

ω = $2\pi f$ radians per second;

it will be seen that the effective capacitance is greater than the nominal capacitance by the factor

$$1/(1 - \omega^2 LC) \quad . \quad . \quad . \quad . \quad . \quad (130)$$

Thus in Type 516A the presence of 1 μH. in series with the condenser at half-scale results in an effective capacitance of 510 $\mu\mu$F. instead of the normal 500 $\mu\mu$F. at 1 Mc. At 2 Mc. the error is 8 per cent approximately, since it is a function of the square of the frequency. In Type 516C efforts have been made to maintain the inductance of the decade constant at all dial settings. This is done by a scheme of differential compensation such that an amount of inductance equal to the inductance of the added resistance cards is removed from the circuit. Similarly, when the amount of resistance, and hence inductance, is reduced, a corresponding amount is added. The arrangement will be seen as R_s and L_0 in the current diagram. To avoid an error in capacitance readings, a small amount of inductance L_0' is placed in series with the unknown quantity. L_0' is so proportioned as to make the total inductance in the X arm equal to L_0 plus stray lead inductances in the L_0 arm.

It will be seen that the standard condenser C_s is ungrounded, thus resulting in a definite capacitance to earth on both sides. However, since the capacitance of the stator to earth C_0 is effectively in parallel with the entire arm of the bridge, it can

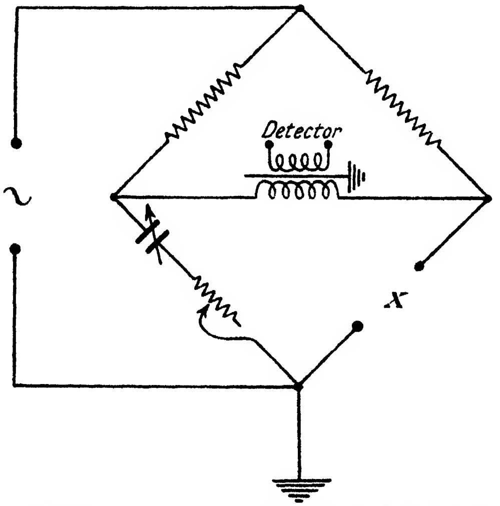


FIG. 78. GENERAL RADIO CO. 516A RADIO-FREQUENCY BRIDGE
Forerunner of Type 516C (Fig. 77). Improved design

cause serious error in the setting of the decade resistance. This will be obvious from the following expression.

$$R' = \frac{R}{(1 - C_0/C)(1 - \omega^2 LC)^2} \quad . \quad . \quad . \quad (131)$$

where R' = effective resistance in ohms;
R = decade dial setting in ohms;
C_0 = earth capacitance (given as 30 $\mu\mu$F. approx.);
C = capacitance of C_s in farads.

To eliminate this error, an equal capacitance is connected across the X arm (indicated as C_s'). By such precautions the bridge can be made to read radio-frequency resistance directly. The circuit of Type 516A is given in Fig. 78. The reader will make a pardonable error if he thinks the Author has been unduly pessimistic or even sceptical about r.f. bridges. This is not really the case, but from the Author's experience, considerable reserve must be practised in dealing with these new radio-frequency designs lest the unwarrantable opinion should be formed that r.f. bridges are just like d.c. or l.f. bridges in all ways except they are energized by radio-frequency sources. Nothing could be farther from the truth. It is interesting, in view of the last two examples, to consider the basic requirements of a radio-frequency bridge. The ideal is Utopian, viz. condensers must be purely capacitative, loss-free, and inductionless; inductances must be pure, that is, contain no resistive or capacitative components. Resistors must be purely resistive—capacitance and inductance free. Then, all these items achieved, there must be no stray couplings in the built-up bridge. Readers have all justification in saying "impossible" and the Author is in complete agreement. However, some constructive ideas are wanted, and in view of the requirements a few suggestions will be put forward.

In the first case, it is considered that r.f. bridges should be designed for operation at a fixed frequency of, say, 1 or 2 Mc. This will assist computation of the various "impure" components and making fixed compensation. The standard condenser is the least troublesome item, as condensers are now available which have losses so small that they can be ignored. Reference should be made in this respect to the first part of this book (Section III, "Condensers"), for Griffiths's expressions for the "goodness factor" of condensers. It then becomes obvious that the main force of attack must be directed towards resistance elements.

It is contrary to physical laws to hope to wind, by any devices, an inductionless, capacitance-free resistance. However, by judicious manipulation of the effective C and L of a resistance, it is possible to get it very nearly pure.* By balancing the reactive components of L and C at a given frequency, zero phase angle is achieved, and hence pure resistance. Because of this, the Author qualifies his belief that a fixed frequency is desirable. Stray capacitances are still present, and one can only use equalizing methods, so far as is yet known. The Author has speculated on the idea of using a thermionic valve for the resistance arm, as large impedance changes are possible

* See *A.C. Bridge Methods*, by Hague. Pitman.

with certain types, such as variable-mu h.f. pentodes. Other difficulties are brought in turn but the idea may be worth consideration. R.f. bridge technique is much like the radio art ten or twenty years ago. We know the fundamental requirements to attain a given performance, but as yet the component parts are not sufficiently developed to enable us to reach our goal. The subject is full of opportunities for research and design.

SECTION X

SUBSIDIARY FACTORY TEST GEAR

THE design of Fig. 79 (*a*) to (*c*) for the rapid testing of r.f. coils, etc., is due to the Author. The standards of reference are quartz crystals having frequencies of 1·5 Mc., 600 kc., 300 kc., and 150 kc. respectively, corresponding to the top and bottom ends

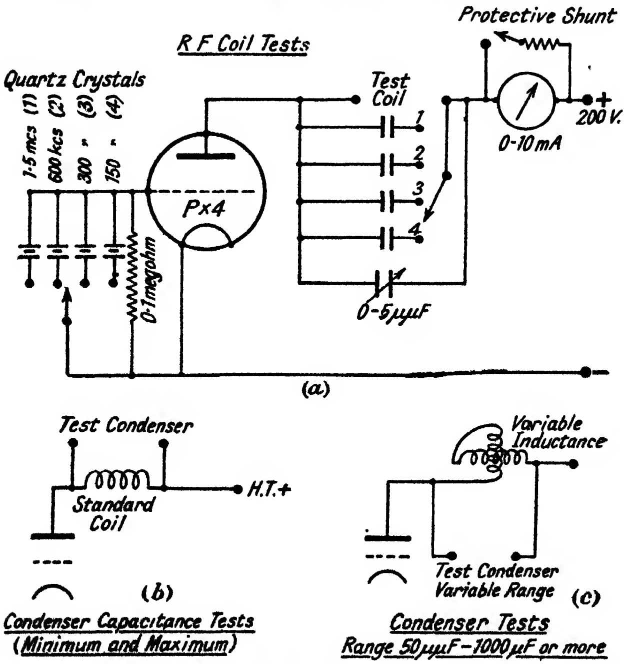


FIG. 79. RADIO-FREQUENCY COIL AND CONDENSER TEST UNIT
(Author's design)

of the m.w. and l.w. bands respectively. The capacitances corresponding to each frequency should closely simulate those in working conditions. Suggested figures are 50 μμF. and 500 μμF. for top and bottom of each range. These should be fixed condensers of good quality, and a trimmer condenser having a range of about 5 μμF. is useful as a device arbitrarily calibrated in "add turns—remove turns" giving numbers. Since the anode current of the valve changes rapidly with the slightest percentage "off tune," the anode milliammeter serves as an excellent indicator. Using a P.X.4 valve with 200 volts on the anode, the optimum anode current is 5 mA., whilst 2·5 mA. and 7·5 mA. correspond to frequencies less than 0·005 per cent off exact resonance, in the case of the 1·5 Mc. crystal. A safety shunt should be connected across the meter to prevent serious overload when the tuned circuit is appreciably off resonance. A 2-gang, 4-contact switch is admirable for quickly changing from one wavelength to another. This same circuit is easily adapted to the measurement of tuning condensers. In this case, either the m.w. or l.w. pair of crystals suffices. A standard coil is connected as per Fig. 79 (*b*), and the condenser is rotated so that an index mark corresponds to the top and bottom of either waveband. Accurate condensers will cause resonance with the crystal frequencies, and hence excitation of the crystal. The milliammeter indicates as in the first example. For the measurement of capacitances between 50 $\mu\mu$F. and 1 000 $\mu\mu$F. a variable inductance is used instead of the fixed standard inductance. This can be calibrated in terms of capacitance corresponding to its inductance at a fixed frequency. In spite of the use of four quartz crystals, the overall cost of the complete equipment should not exceed about £6 to £8. The accuracy of measurements is of a high order.

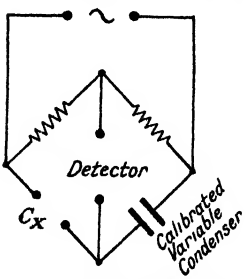


FIG. 80. FUNDAMENTAL CIRCUITS OF CAPACITANCE "LIMIT" BRIDGE (FIG. 81)

In production tests, it is not so much a question of the absolute value of a given resistor or condenser as the requirement of its limit error corresponding to the nominal value. Hence we now have bridges designed to accept components within given limits. Two such designs will be briefly described, and the reader is referred to the full description by Mr. C. G. Seright in *Electronics*.* Fig. 80 shows the basic arrangement of a capacitance "limit" bridge which will handle capacitances from 0·0015 μF. to 0·15 μF., the tolerance being 0–15 per cent. Fig. 81 gives full constructional details and component values.

* Sept., 1934.

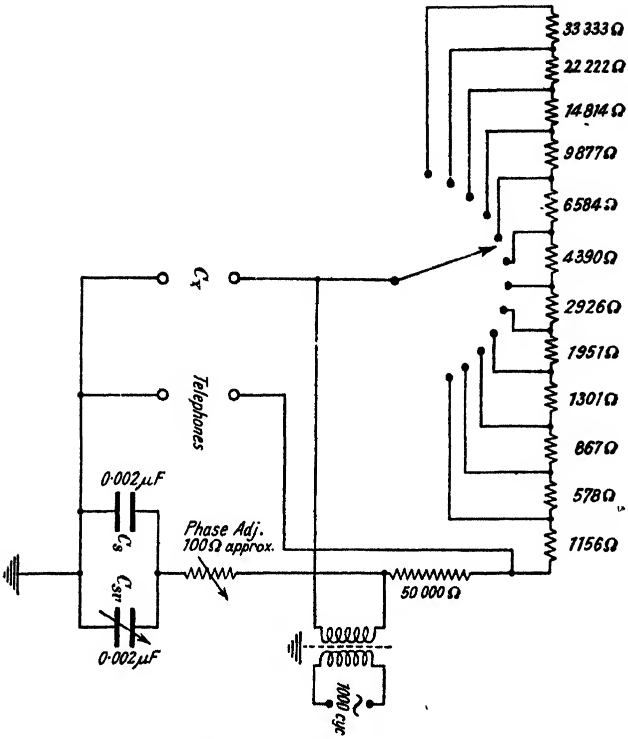


FIG. 81. CAPACITANCE "LIMIT" BRIDGE
Arbitrary marks are made on the C_{sv} scale to indicate tolerance limits

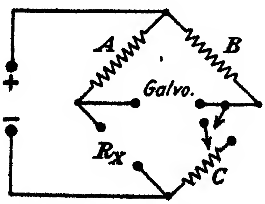


FIG. 82. FUNDAMENTAL CIRCUIT OF FIG. 83

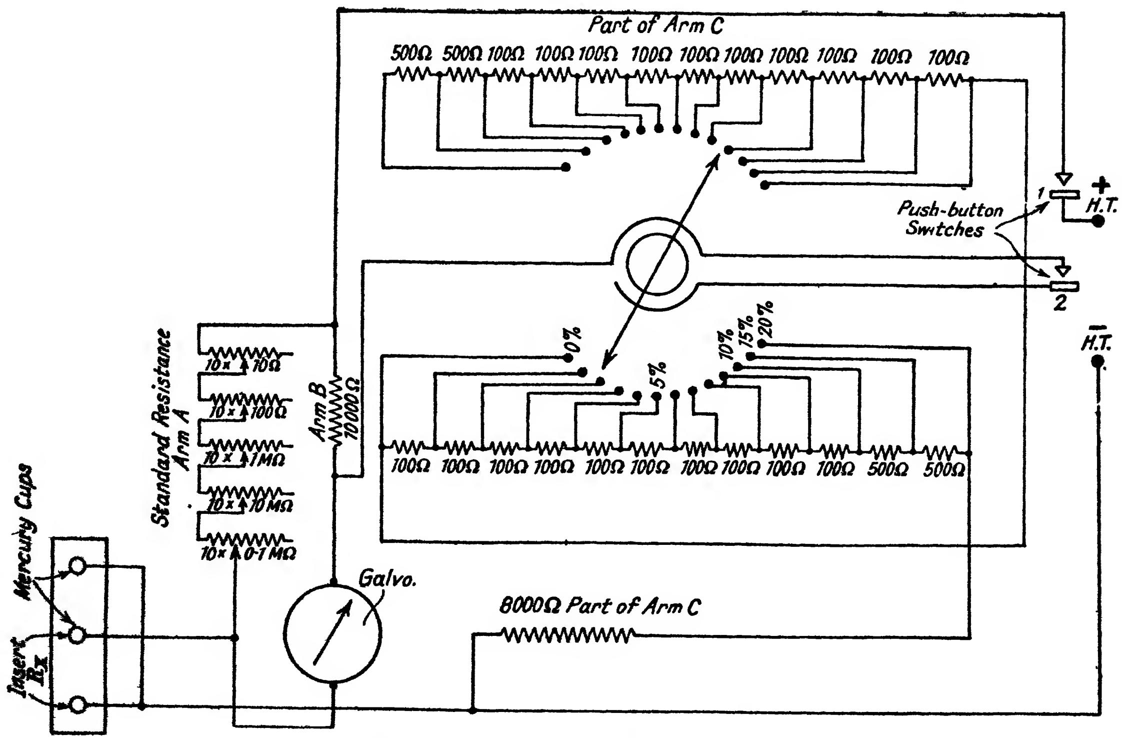


FIG. 83. RESISTANCE "LIMIT" BRIDGE

The bridge should preferably be calibrated after assembly rather than depend upon calculated capacitance readings. Separate scale plates are advantageous for each range, or some method of covering the ranges not in use to prevent errors due to an operator reading the wrong scale.

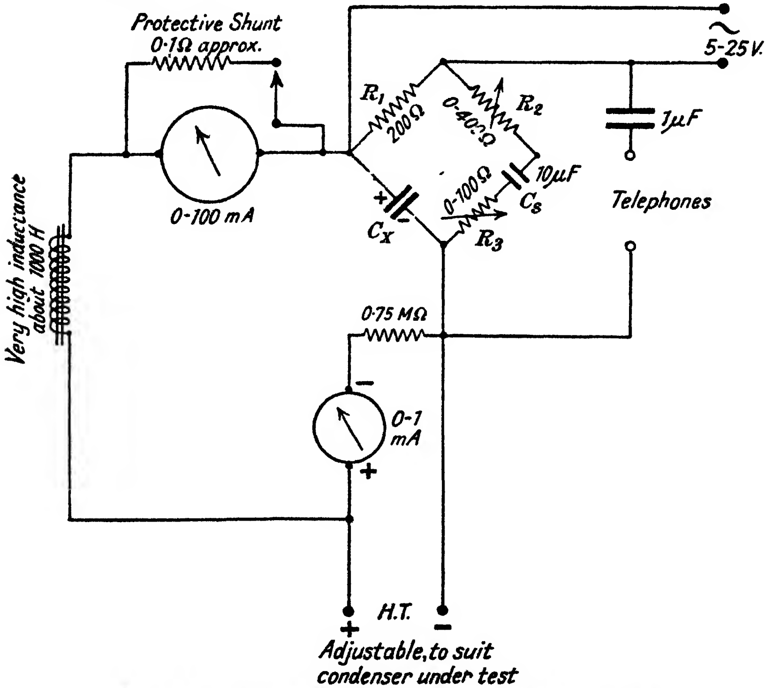


FIG. 84. DEELEY'S ELECTROLYTIC CONDENSER BRIDGE

$$C_x = 1C/(2fEC - 1)$$

where C_x = capacitance of condenser in farads;
C = capacitance of standard condenser in farads;
I = alternating current in amperes; and
E = alternating voltage across C_x and C_s in volts

Fig. 82 shows a Wheatstone bridge arranged as a resistance "limit" bridge, whilst Fig. 83 gives constructional details and component values. This bridge will handle values from a few ohms up to a megohm with a tolerance of 0–20 per cent. In both designs, a 1 000 cyc. oscillator is used as the source, and under factory conditions it may be advantageous to fit an audio-frequency amplifier between the "detector" and phones.

For the production testing of electrolytic condensers, Mr. Paul McKnight Deeley of the Cornell-Dubilier Corporation, U.S.A., has designed a number of methods. Of those described in his article in *Electronics** the bridge method is perhaps the best.

This is shown in full in Fig. 84 and conditions of balance are

$$R_1/R_2 = C_x/C_s; \; C_x = (R_1/R_2)C_s \qquad . \qquad . \quad (132)$$

where C_x = test condenser;

C_s = standard condenser.

Both R_2 and R_3 must be adjusted to balance, and when balance is attained, the resistance of R_3 is the equivalent series resistance of the test condenser.

* July, 1935.