temperature rise. A transformer is never so small that it heats up more in the first operating interval than at the end of many intervals.

From equation 30 can be found a relation between weight, losses, and final temperature rise. For, since heat is dissipated at 0.008 watt per sq in./°C rise, and the area A_S of the equivalent sphere is $4\pi r_e^2$,

$$\theta_0 = \frac{\text{Total watts loss}}{0.008A_S} = \frac{\text{Total watts loss}}{0.1\left(\dfrac{\text{Total weight in pounds}}{1.073}\right)^{2/3}} \tag{31}$$

where θ_0 is the final temperature rise in centigrade degrees. This equation is subject to the same approximations as equation 28; test results show that it is most reliable for transformers weighing 20 lb or more, with 55°C temperature rise at 40°C ambient.

3. RECTIFIER TRANSFORMERS AND REACTORS

Rectifiers are used to convert alternating into direct current. The tubes generally have two electrodes, the cathode and the anode. Both high vacuum and gas-filled tubes are used. Sometimes for control purposes the gas-filled tubes have grids, which are discussed in Chapter 8.

A high-vacuum rectifier tube characteristic voltage-current curve is shown in Fig. 48. Current flows only when the anode is positive with respect to the cathode. The voltage on this curve is the internal potential drop in the tube when current is drawn through it. This voltage divided by the current gives effective tube resistance at any point. Tube resistance decreases as current increases, up to the emission limit, where all the electrons available from the cathode are used. Filament voltage governs the emission limit and must be closely controlled. If the filament voltage is too high, the tube life is shortened; if too low, the tube will not deliver rated current at the proper voltage.

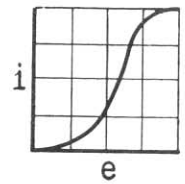


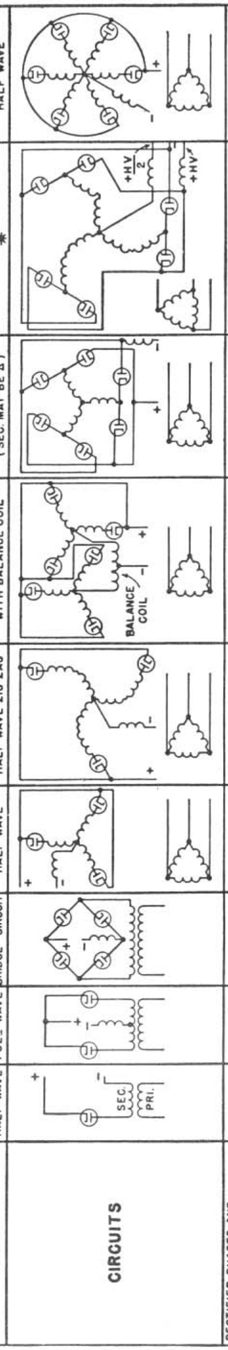
FIG. 48. High-vacuum rectifier voltage-current curve.

Gas-filled rectifier tubes have internal voltage drop which is virtually constant and independent of current. Usually this voltage drop is much lower than that of high vacuum tubes. Consequently, gas-filled tubes are used in high power rectifiers, where high efficiency and low regulation are important. In some rectifiers, silicon or germanium crystals or selenium disks are used as the rectifying elements.

In this chapter, the rectifier circuits are summarized and then rectifier transformers and reactors are discussed.

24. Rectifiers with Reactor-Input Filters. Table VII gives commonly used rectifier circuits, together with current and voltage relations in the associated transformers. This table is based on the use of a reactor-input filter to reduce ripple. The inductance of the choke is assumed to be great enough to keep the output direct current constant. With any finite inductance there is always some superposed

TABLE VII - TABLE OF RECTIFIER CIRCUITS

TYPE	SINGLE PHASE HALF WAVE	SINGLE PHASE FULL WAVE	SINGLE PHASE BRIDGE CIRCUIT	3 PHASE HALF WAVE	3 PHASE HALF WAVE ZIG ZAG	DOUBLE 3 PHASE WITH BALANCE COIL ‡	3 PHASE FULL WAVE (SEC. MAY BE Δ)	3 PHASE FULL WAVE ZIG ZAG *	SIX PHASE HALF WAVE
CIRCUITS									
RECTIFIER PHASES AND NUMBER OF TUBES	1	2	2-4	3	3	6	6	6	6
PHASES OF A-C SUPPLY	1	1	1	3	3	3	3	3	3
SECONDARY VOLT PER LEG	2.22	1.11 (HALF SECT.)	1.11 (WHOLE)	0.855	0.985 (HALF LEG.493)	0.855	0.428	0.493 HALF LEG .247	0.74
PRIMARY VOLTAGE	2.22	1.11	1.11	0.855	0.855	0.855	0.428	0.428	0.74
SECONDARY CURRENT PER LEG	1.57	0.707	1.00	0.577	0.577	0.289	0.816	0.816	0.408
PRIMARY CURRENT PER LEG	1.21	1.000	1.00	0.471	0.408	0.408	0.816	0.816	0.577
SECONDARY K V A	3.48	1.57	1.11	1.48	1.71	1.48	1.05	1.21	1.81
PRIMARY K V A	2.68	1.11	1.11	1.21	1.05	1.05	1.05	1.05	1.28
AVERAGE OF PRIMARY AND SECONDARY K V A ∅	3.08	1.34	1.11	1.35	1.38	1.26	1.05	1.13	1.55
INVERSE PEAK VOLTAGE	3.14	3.14	1.57	2.09	2.09	2.09	1.05	1.05	2.09
RMS CURRENT PER TUBE	1.57	0.707	0.707	0.577	0.577	0.289	0.577	0.577	0.408
PEAK CURRENT PER TUBE	3.14	1.00	1.00	1.00	1.00	0.500	1.00	1.00	1.00
AVERAGE CURRENT PER TUBE	1.00	0.50	0.50	0.33	0.33	0.167	0.33	0.33	0.167
RIPPLE FREQUENCY	f	2 f	2 f	3 f	3 f	6 f	6 f	6 f	6 f
RMS RIPPLE VOLTAGE	1.11	0.472	0.472	0.177	0.177	0.04	0.04	0.04	0.042
RIPPLE PEAKS	+2.14 −1.00	+0.57 −1.00	+0.57 −1.00	+0.209 −0.291	+0.209 −0.291	+0.057 −0.077	+0.057 −0.077	+0.057 −0.077	+0.087 −0.077
LINE POWER FACTOR	0.373	0.90	0.90	0.826	0.955	0.955	0.955	0.955	0.955

NOTE: THE VALUES OF VOLTAGE AND CURRENT ARE EFFECTIVE OR R.M.S., UNLESS OTHERWISE STATED; THEY ARE GIVEN IN TERMS OF THE AVERAGE DC VALUES, AND THE KILOVOLT-AMPERES IN TERMS OF DC KILOWATT OUTPUT. PERFECT TRANSFORMERS, RECTIFIERS AND DC CHOKE ARE ASSUMED, AND $N_1/N_2 = 1$, EXCEPT IN ZIG-ZAG CIRCUITS. SECONDARY AND PRIMARY REFER TO ANODE TRANSFORMER.
‡ FREQUENCY OF BALANCE COIL VOLTAGE 3f. BALANCE COIL VOLTAGE 0.356. PEAK BALANCE COIL VOLTAGE 0.605. BALANCE COIL K V A 0.173.
* USE COLUMN 5 VALUES FOR HALF VOLTAGE TAP. ∅ MAGNETIZING CURRENT ASSUMED NEGLIGIBLE IN ALL CASES.

This table is based mostly on "Polyphase Rectification Special Connections," by R. W. Armstrong, *Proc. I.R.E.*, Vol. 19, Jan. 1931.

ripple current which is neglected in the table, and which is considered further in Chapter 4.

The single-phase half-wave rectifier ordinarily has discontinuous output current, and its output voltage is therefore highly dependent upon the inductance of the input filter choke. For this reason, the currents and voltages are given for this rectifier without a filter.

The difference between primary and secondary v-a ratings in several of these rectifiers does not mean that instantaneous v-a values are different; it means that because of differences in current wave form the rms values of current may be different for primary and secondary.

Unbalanced direct current in the half-wave rectifiers requires larger transformers than in the full-wave rectifiers. This is partly overcome in three-phase transformers by the use of zigzag connections. The three-phase full-wave rectifier can be delta-connected on both primary and secondary if desired; the secondary current is multiplied by 0.577 and the secondary voltage by 1.732. Anode windings have more turns of smaller wire in the delta connection. Single-phase bridge and three-phase full-wave rectifiers require notably low a-c voltage for a given d-c output, low inverse peak voltage on the tubes, and small transformers.

25. Rectifiers with Capacitor-Input Filters. When the filter has no reactor intervening between rectifier and first capacitor, rectifier current is not continuous throughout each cycle and the rectified wave form changes. During the voltage peaks of each cycle, the capacitor charges and draws current from the rectifier. During the rest of the time, no current is drawn from the rectifier, and the capacitor discharges into the load.

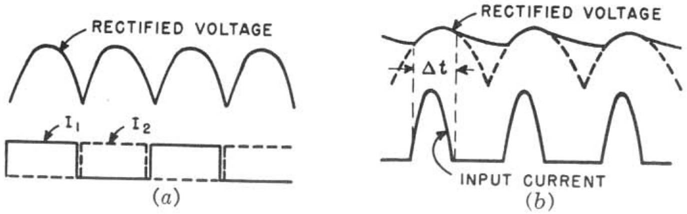


FIG. 49. Voltage and current comparisons in reactor-input and capacitor-input circuits.

Comparison between the rectified voltage of reactor-input and capacitor-input filters in a single-phase full-wave rectifier may be seen in Figs. 49(a) and (b), respectively. The two tube currents I_1 and I_2 in (a) add to a constant d-c output, whereas in (b) the high-peaked tube currents flow only while the rectified voltage is higher than the average d-c voltage. *Average* current per tube in both cases is half the rectifier output. With large values of capacitance, the rectified voltage

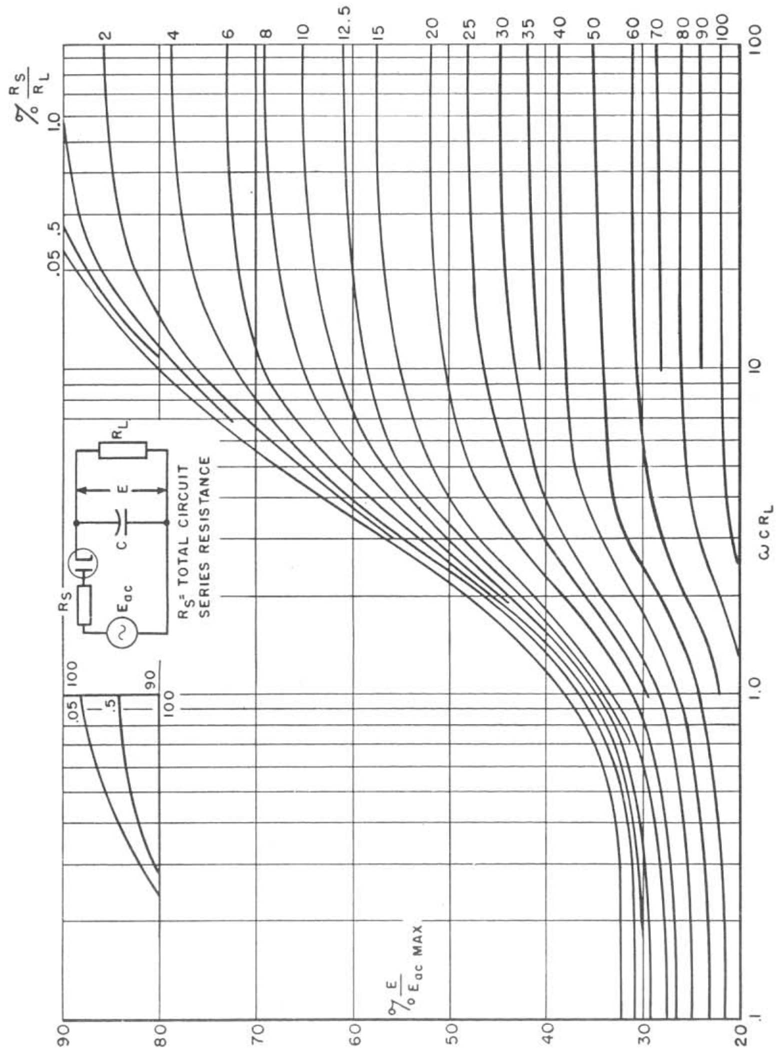


FIG. 50. Relation of peak sine voltage to d-c voltage in half-wave capacitor-input circuits.

in Fig. 49(*b*) increases to within a few per cent of the peak voltage. Ripple, average rectified voltage output, and rectifier current are dependent on the capacitance, the supply line frequency, and the load resistance. They are dependent also on rectifier internal resistance because it affects the peak value of current which the filter capacitor can draw during the charging interval Δt.

Analysis of this charge-discharge action involves complicated Fourier series which require a long time to calculate.[1] Satisfactory

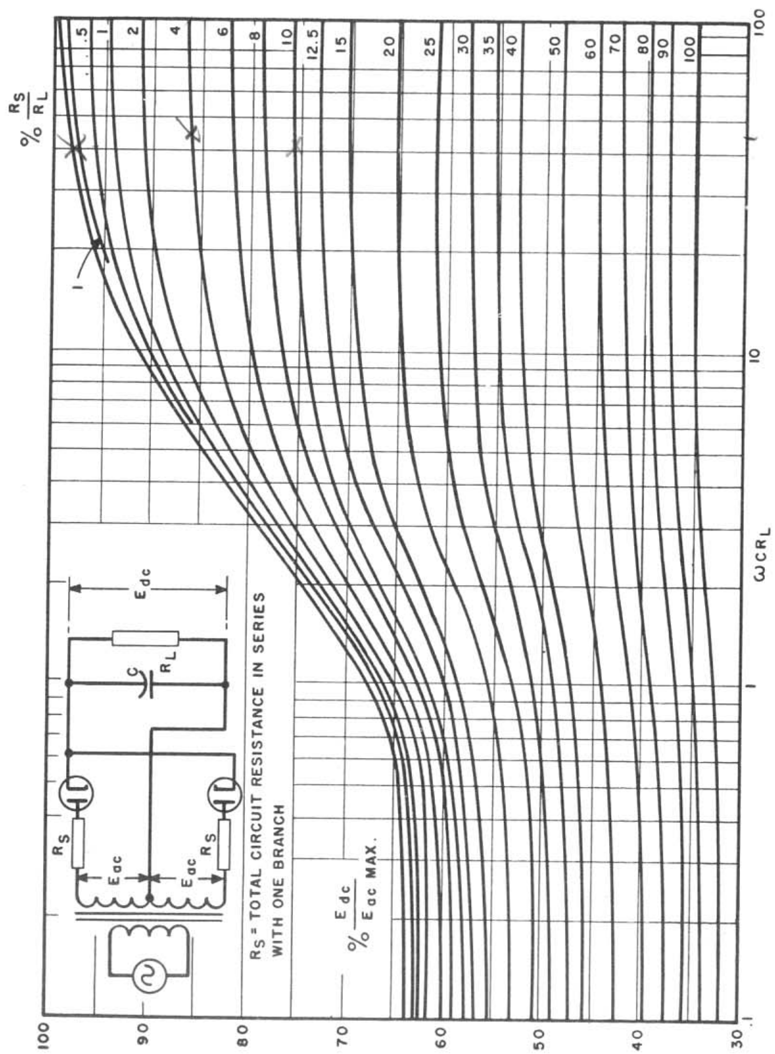


FIG. 51. Relation of peak sine voltage to d-c voltage in full-wave capacitor-input circuits.

voltage and current values have been obtained from experimental measurements by Schade [2] and are shown in Figs. 50, 51, and 52 for

[1] See "Diode Rectifying Circuits with Capacitance Filters," by D. L. Waidelich, *Trans. AIEE, 61,* 1161 (December, 1941).

[2] "Analysis of Rectifier Operation," by O. H. Schade, *Proc. I.R.E., 31,* 341 (July, 1943).

single-phase half-wave and full-wave rectifiers. In these figures R_S is the rectifier series resistance, including the transformer resistance.

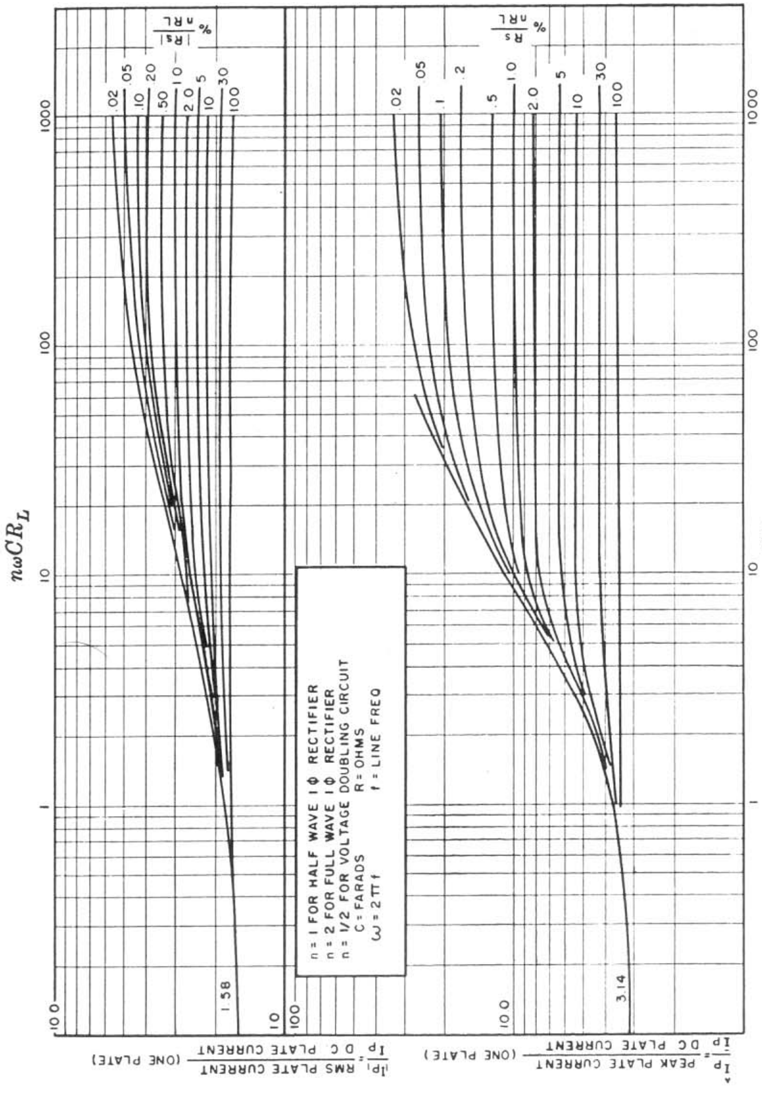


FIG. 52. Relation of peak, average, and rms diode current in capacitor-input circuits.

Results accurate to within 5 per cent are obtained if the rectifier resistance corresponding to peak current $\hat{I}_P$ is used in finding R_S. The process is cut-and-try, because $\hat{I}_P$ depends on R_S, and vice versa, but two trials usually suffice. Resistance is in ohms, capacitance is in

farads, and ω is 2π times the supply frequency. Three-phase rectifiers are rarely capacitor-input because of their larger power.

In Fig. 52 the peak current indicates whether the peak current of a given tube is exceeded, and the rms current determines the transformer secondary heating. The v-a ratings are greater, but *ratios* of primary to secondary v-a ratings given in Table VII hold for capacitor-input transformers also.

26. Voltage Doublers. To obtain more d-c output voltage from a rectifier tube, the circuit of Fig. 53 is often used. With proper values of circuit elements the output is nearly double the a-c peak voltage. Tube inverse peak voltage is little more than the d-c output voltage, and no d-c unbalance exists in the anode transformer. Current output available from this circuit is less than from the single-phase full-wave circuit for a given rectifier tube. Current relations are given in Fig. 52.

Voltage tripling and quadrupling circuits also are used, either to increase the d-c voltage or to avoid the use of a transformer.[1]

27. Filament Transformers. Low-voltage filament transformers are used for heating tube filaments at or near ground potential. Often the filament windings of several tubes are combined into one transformer. Sometimes this requires several secondary windings. In terms of a single secondary transformer a 5 or 6 secondary unit requires about 50 per cent greater size and weight. But these multiwinding transformers are smaller than five or six separate units; this warrants designing them specially in many instances.

Rectifier tube filaments often operate at high d-c voltages and require windings with high voltage insulation. It is usually not feasible to combine high-voltage windings with low-voltage windings when the high voltage is more than 3,000 volts direct current because of insulation difficulties, particularly in the leads. Large rectifier filaments are usually heated by separate transformers; in polyphase rectifiers, all tube filaments are at high voltage, and some secondary windings may be combined. See the three-phase full-wave rectifier in Table VII, where the $+HV$ lead connects to a winding which heats the filaments of three tubes.

Low capacitance filament windings are sometimes required for high-frequency circuits. The problem is not particularly difficult in small v-a ratings and at moderate voltages. Here air occupies most of the space between windings. In larger ratings the problem is more difficult, because the capacitance increases directly as the coil mean turn

[1] See "Analyses of Voltage Tripling and Quadrupling Circuits," by D. L. Waidelich and H. A. Taskin, *Proc. I.R.E.*, *33*, 449 (July, 1945).

length for a given spacing between windings. As voltage to ground increases, there comes a point beyond which creepage effects necessitate

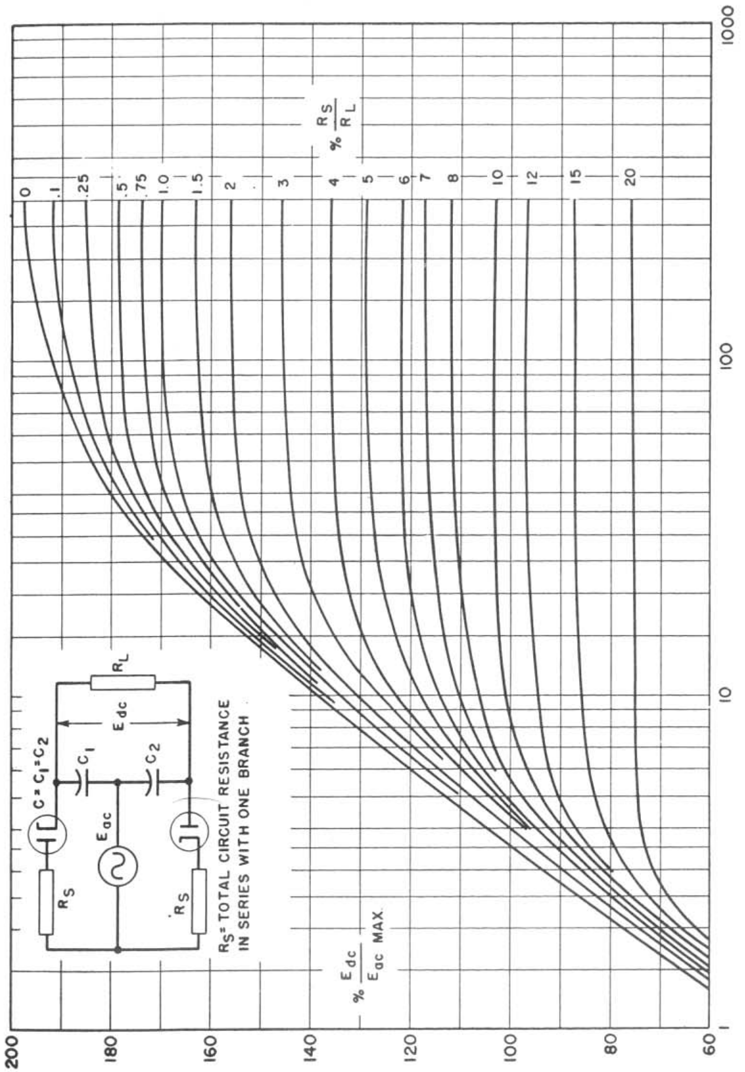


FIG. 53. Relation of peak sine voltage to d-c voltage in voltage-doubling circuit.

oil-insulated windings, whereupon the capacitance jumps 2 to 1 for a given size and spacing. There is a value of capacitance below which it is impossible to go because of space limitations in the transformer. What this value is in any given case may be estimated from the fact

that the capacitance in $\mu\mu$f of a body in free space is roughly equal to one-half its largest dimension in centimeters.

Except for the differences just mentioned, the design of filament transformers does not differ much from that of small 60-cycle power

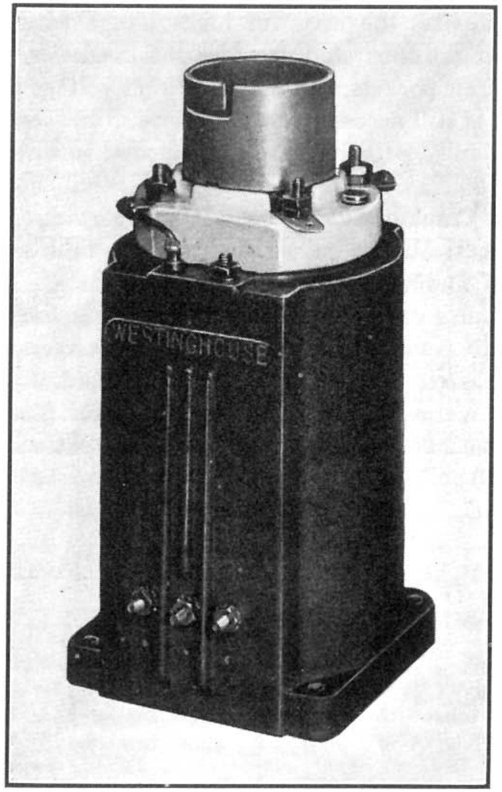


FIG. 54. 15 kv filament transformer enclosed in insulating case.

transformers. The load is constant and of unity power factor. Leakage reactance plays practically no part, because of its quadrature relationship to the load. Output voltage may therefore be figured as in Fig. 3(*c*) (p. 8). It should be *accurately* calculated, however, to maintain the proper filament emission and life.

When a tube filament is cold, the filament resistance is a small fraction of its operating value. In large tubes it is often necessary to protect the tube filaments against the high initial current they would draw at rated filament voltage. This is done by automatically reducing the starting voltage through the use of a current-limiting transformer having magnetic shunts between primary and secondary windings. The design of these transformers is somewhat special, and is included in Chapter 8.

High-voltage filament transformers are sometimes mounted in an insulating case, as in Fig. 54, with the tube socket on top. This arrangement eliminates the need for high-voltage wiring between the transformer and the tube, and provides the insulation for the socket. The problem of air pockets at the base of high-voltage bushings is also eliminated. It is still necessary to insulate well between windings and to fill the case fully with insulating compound in order to eliminate corona.

28. Filament Transformer Design. It is important that design work be done systematically to save the designer's time and to afford a ready means of finding calculations at a later date. To attain these ends a calculation form, such as that in Fig. 55, is used. The form is usually made to cover several kinds of transformers, and only the spaces applicable to a filament transformer are used.

Suppose that a transformer is required to supply filament power for four single-phase full-wave rectifiers having output voltages of 2,000, 500, 250, and 250 volts, respectively, with choke-input filters, as follows:

Primary voltage 100

Frequency 60 cycles

Four secondaries for the following tube filaments:

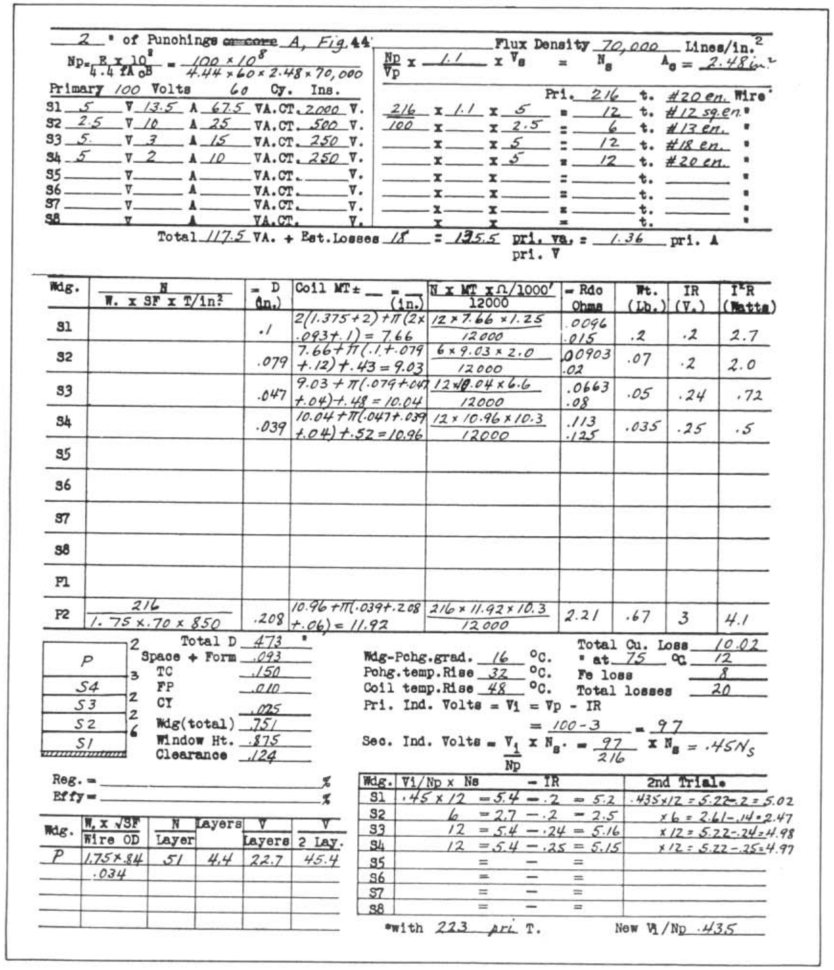
2—872 tubes:	5 volts	13.5 amp	Insulated for +2000 v d-c
2—866 tubes:	2.5 volts	10 amp	Insulated for + 500 v d-c
1—5U4G tube:	5 volts	3 amp	Insulated for + 250 v d-c
1—5Y3GT tube:	5 volts	2 amp	Insulated for + 250 v d-c

Ambient temperature: 40°C

First comes the choice of a core. Data such as those in Fig. 43 are helpful in this, and so is design experience in the modification of such data by the specified requirements. The core used here is a 2-in. stack of laminations *A*, Fig. 44, which is described more fully in Fig. 56, and has enough heat dissipation surface for this rating. For silicon steel, an induction of 70,000 lines per square inch is practical. The primary turns can be figured from equation 4 by making the substitution $\phi = BA_c$ and transposing to

$$N_1 = \frac{E \times 10^8}{4.44 f A_c B} \tag{32}$$

where A_c is the core cross-sectional area, or product of the core tongue width and stack dimension, and B is the core induction. In this transformer, with 90 per cent stacking factor, $A_c = 2 \times 0.9 \times 1.375 = 2.48$ sq in., and the primary turns are found to be 216.



2 " of Punchings ~~or core~~ A, Fig. 44 Flux Density 70,000 Lines/in.²

$Np = \frac{E \times 10^8}{4.44 f A_c B} = \frac{100 \times 10^8}{4.44 \times 60 \times 2.48 \times 70{,}000}$ $\frac{Np}{Vp} \times 1.1 \times V_s = N_s$ $A_c = 2.48$ in.²

Primary 100 Volts 60 Cy. Ins. Pri. 216 t. #20 en. Wire

S1 5 V	13.5 A	67.5 VA.	CT. 2000 V.	$\frac{216}{100}$ × 1.1 × 5 = 12 t.	#12 sq. en. "
S2 2.5 V	10 A	25 VA.	CT. 500 V.	× × 2.5 = 6 t.	#13 en. "
S3 5 V	3 A	15 VA.	CT. 250 V.	× × 5 = 12 t.	#18 en. "
S4 5 V	2 A	10 VA.	CT. 250 V.	× × 5 = 12 t.	#20 en. "
S5 V	A	VA.	CT. V.	× × = t.	"
S6 V	A	VA.	CT. V.	× × = t.	"
S7 V	A	VA.	CT. V.	× × = t.	"
S8 V	A	VA.	CT. V.	× × = t.	"

Total 117.5 VA. + Est. Losses 18 = 135.5 $\frac{\text{pri. va.}}{\text{pri. V}}$ = 1.36 pri. A

Wdg.	N / W. x SF x T/in.²	= D (in.)	Coil MT ± ___ = ___ (in.)	N x MT x Ω/1000′ / 12000	= Rdc Ohms	Wt. (Lb.)	IR (V.)	I²R (Watts)
S1		.1	2(1.375+2)+π(2×.093+.1) = 7.66	12×7.66×1.25 / 12000	.0096 .015	.2	.2	2.7
S2		.079	7.66+π(.1+.079+.12)+.43 = 9.03	6×9.03×2.0 / 12000	.00903 .02	.07	.2	2.0
S3		.047	9.03+π(.079+.047+.04)+.48 = 10.04	12×10.04×6.6 / 12000	.0663 .08	.05	.24	.72
S4		.039	10.04+π(.047+.039+.04)+.52 = 10.96	12×10.96×10.3 / 12000	.113 .125	.035	.25	.5
S5								
S6								
S7								
S8								
P1								
P2	216 / 1.75×.70×850	.208	10.96+π(.039+.208+.06) = 11.92	216×11.92×10.3 / 12000	2.21	.67	3	4.1

P 2, S4 3, S3 2, S2 2, S1 6

Total D .473 "
Space + Form .093
TC .150
FP .010
CY .025
Wdg(total) .751
Window Ht. .875
Clearance .124

Wdg-Pchg.grad. 16 °C.
Pchg.temp.Rise 32 °C.
Coil temp.Rise 48 °C.

Total Cu. Loss 10.02
" at 75 °C 12
Fe loss 8
Total losses 20

Pri. Ind. Volts = V_1 = Vp - IR = 100 - 3 = 97

Sec. Ind. Volts = $\frac{V_1}{Np} \times N_s = \frac{97}{216} \times N_s = .45N_s$

Reg. = %
Effy = %

Wdg.	W. x √SF / Wire OD	N / Layer	Layers	V / Layers	V / 2 Lay.
P	1.75×.84 / .034	51	4.4	22.7	45.4

Wdg.	V1/Np x Ns = – IR =	2nd Trial*
S1	.45 × 12 = 5.4 – .2 = 5.2	.435×12 = 5.22 – .2 = 5.02
S2	6 = 2.7 – .2 = 2.5	×6 = 2.61 – .14 = 2.47
S3	12 = 5.4 – .24 = 5.16	×12 = 5.22 – .24 = 4.98
S4	12 = 5.4 – .25 = 5.15	×12 = 5.22 – .25 = 4.97
S5	= – =	
S6	= – =	
S7	= – =	
S8	= – =	

*with 223 pri. T. New V_1/Np .435

Fig. 55. Filament transformer design calculations.

Below this calculation are set down the primary voltage and frequency, and the voltage, current, volt-amperes, and insulation voltage for all secondary windings. These are designated S_1 to S_4 for identification. From the sum of the individual v-a figures, the transformer rating is found. To it is added an estimate of losses to obtain the input volt-amperes, and the primary current.

Next an estimate of the regulation is made (10 per cent) and added to unity to obtain the multiplier 1.1 in the estimate of secondary turns near the top of the calculation form. From the currents listed, the wire size for each winding is chosen. Round enameled wire is used for

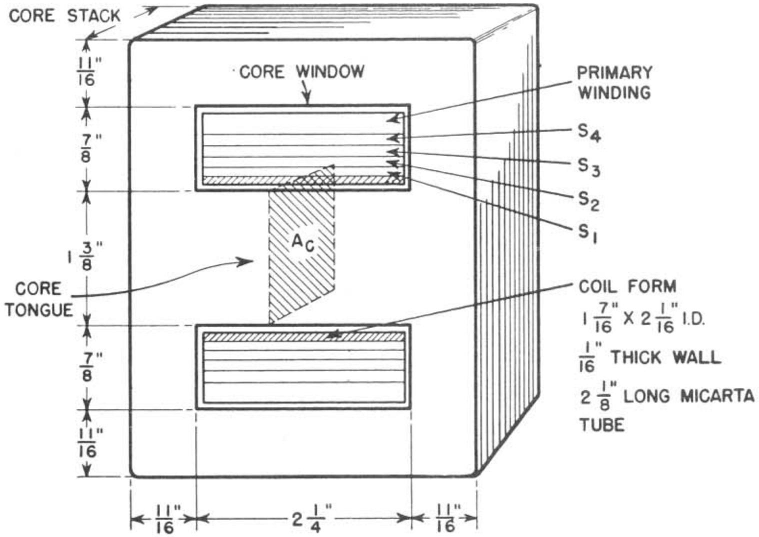


FIG. 56. Dimensions and coil section of filament transformer.

each winding except S_1, and for it No. 12 square wire is used to save space. The largest wire is placed next to the coil form to prevent damage in winding to the smaller wires.

The next task is to find out whether the wire chosen will fit in the core window space. Winding height D is entered for each winding. For each secondary this is the wire diameter, because the wire is wound in a single layer. D for S_1 is slightly larger than the wire dimension to allow for the bulge that occurs when square wire is wound. The twelve turns of S_1 occupy about $1\frac{1}{4}$ in. of horizontal winding space. The core window is $2\frac{1}{4}$ in. wide. From this is subtracted $\frac{1}{8}$ in. for clearance, leaving $2\frac{1}{8}$ in. total coil width. Margins on each side of S_1 are therefore $\frac{1}{2}(2\frac{1}{8} - 1\frac{1}{4}) = \frac{7}{16}$ in. According to Fig. 40 (p. 48) this provides over 8 kv breakdown strength, which is well above the 5-kv test voltage for S_1. Other secondary windings have lower test voltages and wider margins, and hence have more than adequate creepage distances.

The $\frac{1}{16}$-in.-thick Micarta rectangular tube used for the coil form has a corona voltage of 2,700 rms, which affords about 23 per cent safety factor over the normal operating voltage at the tube filaments. Over S_1 are wound six wraps of 0.010-in.-thick treated cloth, which has 2,600-volt corona limit. Winding S_2 supplies a filament at 500 volts of the same polarity as S_1. Hence only 1,500 volts direct current or 1,660 volts alternating current occur across this insulation. At the right of the small sketch in Fig. 55 are listed the number of wraps of 0.010-in.-thick treated cloth over each section of winding. These are added together to give the columnar figure of 0.150 for TC.

The primary winding is wound without layer insulation and with an area space factor of 70 per cent. Cotton is wound in with the wire to form walls $\frac{3}{16}$ in. thick on either side of the primary; this accounts for the low space factor and for the $1\frac{3}{4}$-in. winding traverse. The coil is finished with two layers of treated cloth, a layer of 0.010-in. fishpaper for mechanical protection, and a 0.025-in. serving of untreated cotton yarn or tape to hold it together. The total winding adds up to 0.751 in., leaving 0.124 in. clearance, about the right allowance for winding slack for four secondaries.

Mean turns are figured from equations 26 and 27, with 5 per cent incremental increase in S_2, S_3, and S_4 for leads. With the mean turn values the winding resistances, weights of copper, and IR and I^2R for each winding can be found. To S_1, S_2, and S_3 winding resistance is added lead resistance, and the lower figure is the sum of the two in each case. Total copper loss is multiplied by 1.3 to correct for 75°C operating temperature. The core weight is 6.8 lb, and the grade of steel used has 1.17 watts per pound at 70,000 lines per square inch. This gives a core loss of 8 watts, and a total of copper and core loss of 20 watts. After these losses are divided by the appropriate ordinates from Figs. 44 and 45 (pp. 55 and 56) the coil temperature rise is figured at 48 centigrade degrees, which is safe for class A insulation.

We know by now that the design is safe, but secondary voltages still must be checked. The method of equation 13 is used. Output voltages on first trial range from 0 to 4 per cent high. S_2 voltage is correct but out of line with the rest. Changing S_2 leads to a larger size makes the per cent voltage drops more nearly alike, and increasing the primary turns to 223 brings all output voltages to correct value within 1.2 per cent. Filament voltage should be kept within 2 per cent for these tubes, to allow for meter error. Primary voltage per layer is checked at the lower left; this is equivalent to 22.7 volts per mil of wire enamel, which is safe practice.

If the design were deficient in any respect, even down to the last things figured, some change would have to be made which would require recalculation of all or part of the transformer; hence the importance of good estimating all the way along.

The filament transformer outlined above had a center tap (C.T.) in each filament winding. Such taps are used with directly heated cathodes, especially when plate current is large, to prevent uneven distribution of filament emission. In windings for supplying filaments of small tubes, center taps are sometimes omitted. Ripple in the rectified output then increases, and transformer core flux density becomes asymmetrical. Whether these effects are permissible depends on operating conditions. Usually plate current is much smaller than filament current, so that center-tap leads may be smaller in copper section than start and finish leads. A certain amount of space is required for these leads; rectifier wiring is also more time-consuming when there are center taps. Nevertheless, the extra work and size may be justified by improved performance.

An even number of turns, such as were used in the transformer windings described in this section, results in center-tap placement on the same coil end as the start and finish leads; if there were an odd number of turns, the tap lead would be at the opposite end. In a single-core, single-coil design, an odd number of turns cannot be center-tapped exactly. Usually the unbalance caused by the tap being a half-turn off center is not serious, but it should not be disregarded without calculation.

29. Anode Transformers. Anode transformers differ from filament transformers in several respects.

(*a*) *Currents* are non-sinusoidal. In a single-phase full-wave rectifier, for instance, current flows through one half of the secondary during each positive voltage excursion and through the other half during each negative excursion. For half of the time each half-secondary winding is idle.

(*b*) *Leakage inductance* not only determines output voltage but also affects rectifier regulation in an entirely different manner than with a straight a-c load. This is discussed in Chapter 4.

(*c*) *Half-wave rectifiers* carry unbalanced direct current; this may necessitate less a-c flux density, hence larger transformers, than full-wave rectifiers. Unbalance in the three-phase half-wave type can be avoided by the use of zigzag connections, but an increase in size over full-wave results because of the out-of-phase voltages. These connections are desirable in full-wave rectifiers when half voltage is obtained from a center tap. See Table VII.

(*d*) *Single-phase full-wave rectifiers with two anodes* have higher secondary volt-amperes for a given primary v-a rating than a filament transformer. Bridge-type (four-anode) rectifiers have equal primary and secondary volt-amperes, as well as balanced direct current, and plate transformers for these rectifiers are smaller than for other types. Three-phase rectifier transformers are smaller in total size but require more coils. The three-phase full-wave type has equal primary and secondary v-a ratings.

(*e*) *Induced secondary voltage* is much higher. Filament transformers are insulated for this voltage but have a few secondary turns of large wire, whereas anode transformers have many turns of small wire. For this reason the volts per layer are higher in anode transformers, and core windows having proportionately greater height and less width than those in Fig. 56 are often preferable. This trend runs counter to the conditions for low leakage inductance and makes it necessary to interleave the windings. Figure 57 shows the windings of a single-phase full-wave rectifier transformer with the primary interleaved between halves of the secondary. This arrangement is especially adaptable to transformers with grounded center tap. The primary-secondary insulation can be reduced to the amount suitable for primary to ground. This is called *graded* insulation.

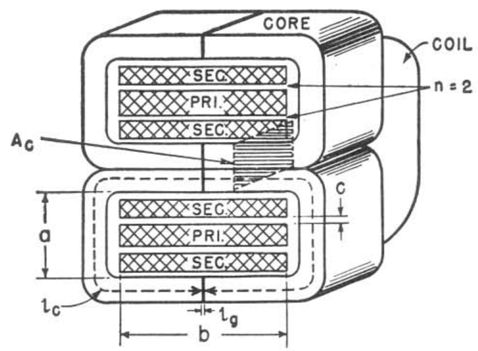


FIG. 57. Dimensions and coil section of anode transformer. Construction shown is for shell-type transformer with 2 Hipersil cores.

In large power rectifiers of the gas-filled or pool types, anode current under short-circuit conditions may be very great, and anode transformer windings must be braced to prevent damage. If the conductors are small, solventless varnish is useful for solidly embedding the conductors.

30. Leakage Inductance. Flux set up by the primary winding which does not link the secondary, or vice versa, gives rise to leakage or self-inductance in each winding without contributing to the mutual flux. The greater this leakage flux, the greater the leakage inductance, because the inductance of a winding equals the flux linkages with unit current in the winding. In Fig. 57, all flux which follows the core path l_c is mutual flux. Leakage flux is the relatively small flux which threads the secondary winding sections, enters the core, and returns to the other side of the secondaries, without linking the primary. The same is true of flux linking only the primary winding. But it is almost impossible for flux to leave the primary winding, enter the core, and re-enter the primary without linking part of the secondary also. The more the primary and secondary windings are interleaved, the less leakage flux there is, up to the limit imposed by flux in the spaces c between sections. These spaces contain leakage flux also; indeed, if there is much interleaving or if the spaces c are large, most of the leakage flux flows in them. Large coil mean turn length, short winding traverse b, and tall window height a all increase leakage flux.

Several formulas have been derived for the calculation of leakage inductance. That originated by Fortescue[1] is generally accurate, and errs, if at all, on the conservative side:

$$L_S = \frac{10.6N^2 MT(2nc + a)}{10^9 n^2 b} \tag{33}$$

where L_S = leakage inductance of both windings in henrys, referred to the winding having N turns
MT = mean length of turn for whole coil in inches
n = number of dielectrics between windings ($n = 2$ in Fig. 57)
c = thickness of dielectric between windings in inches
a = winding height in inches
b = winding traverse in inches.

The greatest gain from interleaving comes when the dielectric thickness c is small compared to the window height; when nc is comparable to the window height, the leakage inductance does not decrease much as n is increased. It is often difficult to reduce the leakage inductance which occurs in high-voltage transformers because of leakage flux in

[1] See *Standard Handbook for Electrical Engineers,* McGraw-Hill Book Co., New York, 1922, 5th ed., p. 413.

spaces c. A small number of turns, short mean turn, and low, wide core windows all contribute to a low value of leakage inductance.

31. Anode Transformer Design. Let the requirements of a rectifier be

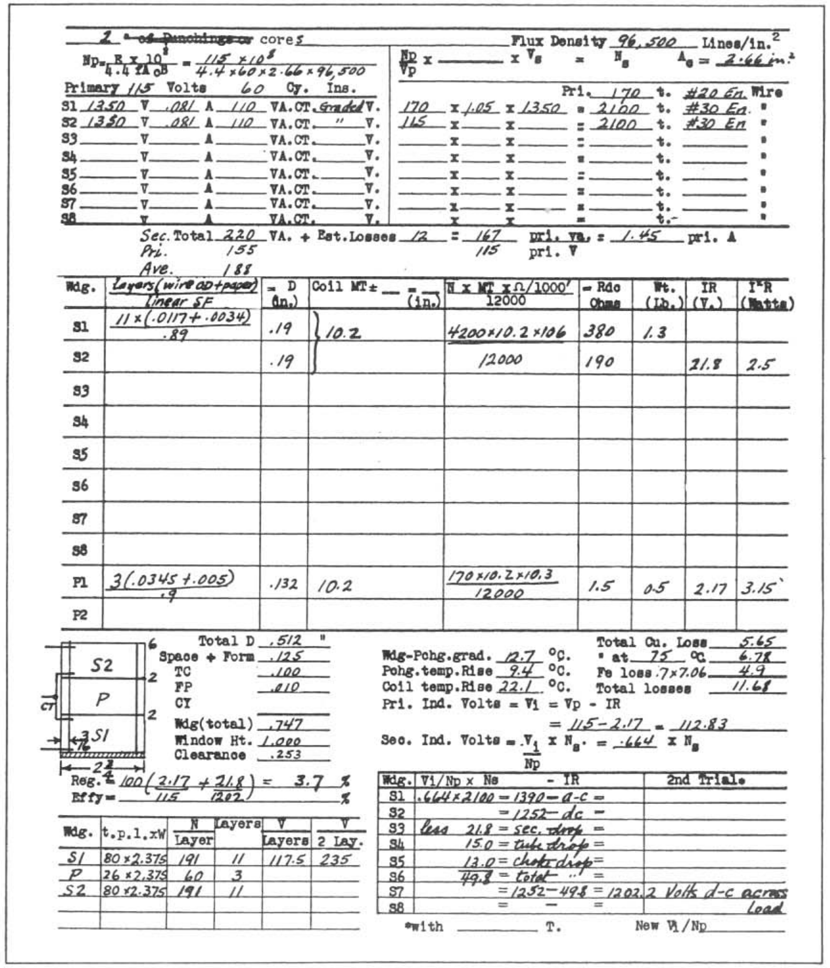
1,200 volts 115 ma rectifier d-c output
Single-phase full-wave circuit with 866 tubes
Primary 115 volts 60 cycles
Rectifier regulation 5 per cent maximum
Ambient 55°C

To fulfill these requirements, a reactor-input filter must be used. If 1 per cent is allowed for reactor IR drop, a maximum of 4 per cent regulation is left in the anode transformer. The approximate secondary output voltage is $1{,}200 \times 2.22 = 2{,}660$, say 2,700 volts. The center tap may be grounded. Suppose that a transformer like the one in Fig. 57 is used. The calculations are given in Fig. 58. The various steps are performed in the same order as in filament transformers. The grain-oriented type C core is worked at 38 per cent higher induction, with but 60 per cent of the core loss of Fig. 55; its strip width is 2¼ in., build-up ⅝ in., and window 1 in. by 3 in. for each core loop. Note the difference in primary and secondary volt-amperes and winding heights. Since the primary and secondary are symmetrical about the primary horizontal center line, they have the same mean turn length. Losses and temperature rise are low. Regulation governs size. Secondary layer voltage is high enough to require unusually thick layer paper. This coil is wound on a multiple-coil machine. Winding height is figured on the basis of layer paper adequate for the voltage instead of from Table VI (p. 39), but turns per layer are taken from this table. Since adjacent layers are wound with opposite directions of traverse, the highest voltage across the layer insulation is twice the volts per layer. Layer insulation is used at 46 volts per mil in the secondary; this counts the 1.7 mils of double enamel, which must withstand impregnation without damage. Anode leads and margins withstand 5 kv rms test voltage. Since the secondary center tap is grounded, two thicknesses of 0.010-in. insulation between windings are sufficient. Clearance of 0.253 in. allows room for in-and-out coil taping.

Secondary leakage inductance, from equation 33, is

$$\frac{10.6 \times 4{,}200^2 \times 10.2(4 \times 0.020 + 0.747)}{4 \times 2.375 \times 10^9} = 0.166 \text{ henry}$$

At 60 cycles this is 6.28 × 60 × 0.166 = 63 ohms, which would be 240 ohms if the secondary were a single section, and which would increase regulation as set forth in Chapter 4. The regulation calcu-



2 ~~No. of Punchings or~~ cores S — Flux Density 96,500 Lines/in.^{2}

$N_p = \frac{E \times 10^8}{4.44 f A_c B} = \frac{115 \times 10^8}{4.4 \times 60 \times 2.66 \times 96,500}$ — $\frac{N_p}{V_p} \times$ ____ $\times V_s = N_s$ — $A_c = 2.66$ in.^{2}

Primary 115 Volts 60 Cy. Ins. — Pri. 170 t. #20 En. Wire

S1	1350 V	.081 A	110 VA. CT.	Graded V.	170 x	1.05 x	1350 =	2100 t.	#30 En. "
S2	1350 V	.081 A	110 VA. CT.	" V.	115 x	x	=	2100 t.	#30 En "
S3	V	A	VA. CT.	V.	x	x	=	t.	"
S4	V	A	VA. CT.	V.	x	x	=	t.	"
S5	V	A	VA. CT.	V.	x	x	=	t.	"
S6	V	A	VA. CT.	V.	x	x	=	t.	"
S7	V	A	VA. CT.	V.	x	x	=	t.	"
S8	V	A	VA. CT.	V.	x	x	=	t.	"

Sec. Total 220 VA. + Est. Losses 12 = 167 pri. va. = 1.45 pri. A
Pri. 155 — 115 pri. V
Ave. 188

Wdg.	Layers(wire OD+paper) / Linear SF	= D (in.)	Coil MT ± __ = __ (in.)	N x MT x Ω/1000′ / 12000	= Rdc Ohms	Wt. (Lb.)	IR (V.)	I²R (Watts)
S1	11 × (.0117 + .0034) / .89	.19	10.2	4200 × 10.2 × 106 / 12000	380	1.3		
S2		.19			190		21.8	2.5
S3								
S4								
S5								
S6								
S7								
S8								
P1	3(.0345 + .005) / .9	.132	10.2	170 × 10.2 × 10.3 / 12000	1.5	0.5	2.17	3.15
P2								

S2 6 — P 2 — S1 2 — CT — 3/16 — 2 3/4

Total D .512 "
Space + Form .125
TC .100
FP .010
CY
Wdg(total) .747
Window Ht. 1.000
Clearance .253

Wdg-Pchg. grad. 12.7 °C. — Total Cu. Loss 5.65
Pchg. temp. Rise 9.4 °C. — " at 75 °C 6.78
Coil temp. Rise 22.1 °C. — Fe loss .7 × 7.06 4.9
Total losses 11.68

Pri. Ind. Volts = V_1 = Vp - IR = 115 - 2.17 = 112.83

Sec. Ind. Volts = $\frac{V_1}{N_p} \times N_s$ = .664 x N_s

Reg. = $100\left(\frac{2.17}{115} + \frac{21.8}{1202}\right) = 3.7$ %
Effy = %

Wdg.	t.p.l. x W	N / Layer	Layers	V / Layers	V / 2 Lay.
S1	80 × 2.375	191	11	117.5	235
P	26 × 2.375	60	3		
S2	80 × 2.375	191	11		

Wdg.	V1/Np x Ns - IR	2nd Trial*
S1	.664 × 2100 = 1390 — a-c =	
S2	= 1252 — dc =	
S3	less 21.8 = sec. drop =	
S4	15.0 = tube drop =	
S5	13.0 = choke drop =	
S6	49.8 = total " =	
S7	= 1252 — 49.8 = 1202.2 Volts d-c across load	
S8	= — =	

*with ____ T. New V_1/Np ____

Fig. 58. Anode transformer design calculations.

lated in Fig. 58 is that due to primary *IR* calculated in the normal manner, plus I_{dc} times one-half the secondary winding resistance.

When high voltage is induced in a winding, the layer insulation and coil size may often be reduced by using the scheme shown in Fig. 59. This is applicable to a plate transformer of the single-phase full-wave

type with center tap grounded. It then becomes practical to make the secondary in two separately wound vertical halves or part coils. One of the part coils is assembled with the turns in the same direction as those of the primary, and the other part coil is reversed so that the turns are in the opposite direction. The two start leads are connected together and to ground as in Fig. 59. It is necessary then to provide only sufficient insulation between windings to withstand the primary test voltage. Channels may be used to insulate the secondaries from the core. With higher voltages, it may be necessary to provide pressboard spacers between the secondary part coils, or to tape the secondary coils separately, but margins must be provided sufficient to prevent creepage across the edges of the spacers.

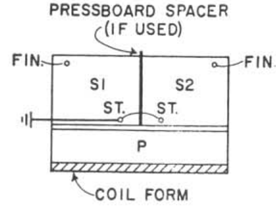


Fig. 59. Anode transformer with C.T. grounded.

32. Combined Anode and Filament Transformers. Anode and filament windings are combined into a single transformer mainly in low-power ratings such as those in receivers and grid bias power supplies.

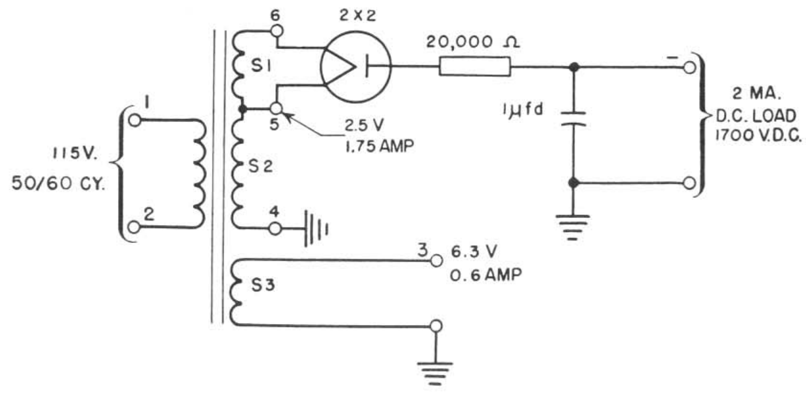


Fig. 60. Power supply transformer.

One widely used combination includes the anode and filament windings for a rectifier and a filament winding for the amplifier tubes. Figures 60 and 61 show how winding insulation sometimes may be graded to require a minimum of insulation and space. The high-voltage filament winding S_1 is placed over the coil form to take advantage of its thick insulation. Layer insulation is sufficient between S_1 and S_2, and between S_2 and S_3. Over and under the primary winding is 115-volt insulation. Thus Fig. 61 is a high-voltage transformer with no high-voltage insulation in it except what is incidental to the coil form.

Combined anode and filament transformers are difficult to test for regulation or output voltage aside from operation in the rectifier circuit itself, because a-c loads do not duplicate rectifier action. Most transformers of this kind are used in rectifiers with capacitor-input filters or with fixed loads in which regulation is not important.

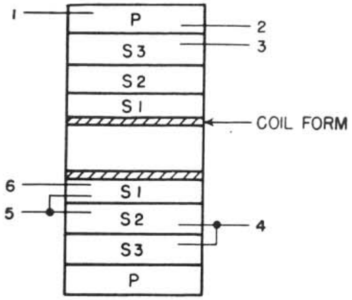


FIG. 61. Winding arrangement to save insulation.

Ratings are easier to predict. Anode secondary v-a rating is the product of rms voltage and current, but the corresponding portion of primary v-a rating depends on the rectifier and is found as mentioned in Sections 24 and 25. To this is added the sum of filament winding v-a ratings, and the primary current can then be calculated from the total volt-amperes.

33. Power Supply Frequency. Foregoing examples were based on a 60-cycle supply. Twenty-five-cycle transformer losses are lower for a given induction. It follows that induction can be increased somewhat over the 60-cycle value, but saturation currents prevent a decided increase. Larger size results, nearly 2:1 in volume. Otherwise 25-cycle transformers are not appreciably different from 60-cycle transformers.

Power supply frequencies of 400 and 800 cycles are used mainly in aircraft and portable equipment to save weight and space. Silicon-steel core materials 0.005 in. thick are principally used at these frequencies to reduce eddy currents. Losses at 400 and 800 cycles for three core materials are shown in Fig. 62. These losses can be the controlling factors in determining transformer size, because a given material saturates at nearly the same induction whether the frequency is 60 cycles or 800 cycles, but the core loss is so high at 800 cycles that the core material cannot be used near the saturation density. The higher the induction the higher the core heating. For this reason, class B insulation can be used in many 400- and 800-cycle designs to reduce size still further. If advantage is taken of both the core material and insulation, 800-cycle transformers can be reduced to 10 per cent of the size of 60-cycle transformers of the same rating. Typical

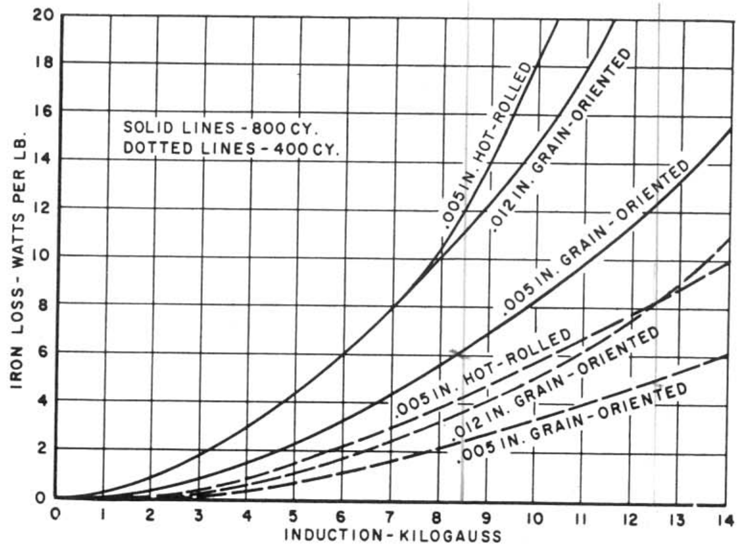


FIG. 62. Silicon-steel core loss at 400 and 800 cycles.

combinations of grain-oriented core material and insulation are as follows:

Frequency	Strip Thickness	B-Gauss	Class of Insulation	Operating Temperature
60	0.014	15,000	A	95°C
400	0.005	12,500	B	140°C
800	0.005	8,500	B	140°C

In very small units, these flux densities may be used at lower temperatures and with class A insulation because of regulation. The special 4-mil steel developed for 400 cycles makes possible size reduction comparable to that for 800 cycles. The necessity for small dimensions, especially in aircraft apparatus, continually increases the tendency to use materials at their fullest capabilities.

Many small 60-cycle transformers have core loss which is small compared to winding or copper loss. This condition occurs because inductance is limited by exciting current rather than by core loss. As size or frequency increases, this limitation disappears, and core loss is limited only by design considerations. Under such circumstances, the ratio of core to copper loss for maximum rating in a given size may be found as follows. Let

W_e = core loss
W_s = copper loss
K_1, K_2, etc. = constants
E = secondary voltage
I = secondary current

For a transformer with a given core, winding, volt-ampere rating, and frequency, $W_e \approx K_1E^2$. For a given winding, $W_s = K_2I^2$. Also, for a given size, $W_e + W_s = K_3$, a quantity determined by the permissible temperature rise. Hence the transformer volt-ampere rating is approximately

$$EI = \sqrt{\frac{W_eW_s}{K_1K_2}}$$
$$= K_4\sqrt{W_e(K_3 - W_e)}$$

For a maximum, the rating may be differentiated with respect to W_e, and the derivative equated to zero:

$$0 = K_3 - 2W_e$$

whence

$$W_e = K_3/2$$

so that $W_s = K_3/2$, or copper and core losses are equal for maximum rating.

Although this equality is not critical, and is subject to many limitations such as core shape, voltage rating, and method of cooling, it does serve as a guidepost to the designer. If a transformer design is such that a large disparity exists between core and copper losses, size or temperature rise often may be reduced by a redesign in the direction of equal losses.

34. An 800-Cycle Transformer Design.

Primary 120 volts 800 cycles
Rectifier to deliver 0.2 amp at +450 volts using 5U4G in single-phase full-wave circuit with 0.5-μfd capacitor input filter.

Figures 51 and 52 tell whether the product ωCR_L will produce the necessary d-c output without exceeding the rectifier tube peak inverse voltage rating and peak current rating.

$$\omega CR_L = 6.28 \times 800 \times 0.5 \times 10^{-6} \times (450/0.2) = 5.65$$

For R_S assume a peak current of 0.5 amp. Average anode characteristics show 97 volts tube drop, or 97 ÷ 0.5 = 194 ohms at peak current. R_S/R_L = 194/2,250 = 0.086. Add 5 per cent for transformer windings; estimated R_S/R_L = 13.6 per cent.

Check on Peak Current from Fig. 52.

$$n\omega CR_L = 11.3$$

$$\hat{I}_p = 5I_p = 5 \times 0.1 = 0.5 \text{ amp}$$

the peak value assumed. Rms current in tube plates and secondary windings is 2 × 0.1 = 0.2 amp. Output voltage, from Fig. 51, is 0.69 peak a-c voltage per side. Hence secondary rms voltage per side is 450 × 0.707 ÷ 0.69 = 460 volts, and secondary volt-amperes = 2 × 460 × 0.2 = 184. The anode transformer must deliver 2 × 460 = 920 volts at 0.2 amp rms. Primary volt-amperes = 0.707 × 184 = 130.

Inverse peak voltage is the peak value of this voltage plus the d-c output, because the tube filament is at d-c value, plus a small amount of ripple, while one anode has a maximum of peak negative voltage, during the non-conducting interval. Thus peak inverse voltage is 460 × 1.41 + 450 = 1,100 volts, which is within the tube rating.

Choice of core for this transformer is governed by size and cost considerations. Assume that the core works at 8,500 gauss. The loss per pound for 0.005-in. silicon steel and grain-oriented steel is 12.2 and 6.6, respectively. (See Fig. 62.) But punchings have 80 per cent stacking factor, whereas the type C core has 90 per cent. In this thickness 0.005-in. grain-oriented steel compares still better with ordinary silicon steel than Fig. 62 would indicate and so will be used for the core.

Let two type C cores be used with the following dimensions:

Strip width	¾ in.	Window height	⅝ in.
Build	⅜ in.	Window width	1½ in.
Total net core area	0.506 sq in.	Core weight	0.75 lb

Turns could be figured from equation 32, except that the induction is in gauss. Since many core data are given in gauss, equation 32 is changed for convenience to

$$N_1 = \frac{3.49E \times 10^6}{fA_cB} \tag{34}$$

where dimensions are in inches and B is in gauss. Primary turns are then

$$\frac{3.49 \times 120 \times 10^6}{800 \times 0.506 \times 8{,}500} = 122$$

Final design figures are:

Primary 122 turns	No. 26 glass-covered wire	d-c resistance 1.8 ohms
Secondary 900 turns	No. 29 glass-covered wire	d-c resistance 38 ohms

Primary copper loss at 100°C	=	3.35 watts
Secondary copper loss at 100°C	=	2.04 watts
Core loss 6.6 × 0.75	=	4.95 watts
Total losses		10.34 watts

With an open-type mounting and mica insulation this transformer has a temperature rise of 75 centigrade degrees.

35. Polyphase Transformers. In large power rectifiers three-phase supplies are generally used. Accurate phase voltages must be maintained to avoid supply frequency ripple in the output. Delta-connected primaries are shown in Table VII for the various rectifiers; these are preferable to open-delta because phase balance is better, and to Y-connections because of possibly high third harmonics. Open-delta connections require only two single-phase transformers instead of three, but a similar saving may be had by using a single core-type three-phase unit which retains the phase-balance advantage. The main drawback to a three-phase core is its special dimensions. Often, to use standard parts, three single-phase units are employed in the smaller power ratings. But if the power is hundreds or thousands of kilowatts, the cores are built to order, and the weight saving in a three-phase core is significant.

Two- and three-phase filament transformers are used with output tubes for large broadcast stations to heat filaments uniformly and reduce hum in the r-f output.

36. Design Chart. In preceding sections, it has been stated that special conditions require tailored designs. Windings for simple low-voltage 60-cycle transformers may be chosen from the chart of Fig. 63. This chart is based upon the following conditions:

(*a*) Two untapped concentric windings; primary wound first.
(*b*) Operating voltage in both windings less than 1,000 volts.
(*c*) Power supply frequency 60 cycles.
(*d*) Maximum temperature rise 40°C in 65°C ambient.
(*e*) Resistive loads.

(*f*) Equal I^2R losses in primary and secondary.
(*g*) Solventless resin impregnated coils.
(*h*) Open-type assemblies like those of Fig. 15.
(*i*) Grain-oriented silicon-steel cores.

It was found that 40°C rise in the four smallest sizes resulted in excessive voltage regulation. For example, a small filament transformer would deliver correct filament voltage at room ambient temperature of 25°C, but at 105°C this voltage dropped to less than the published tube limit. Hence the winding regulation in the two smallest transformers was limited to 15 per cent, and in the next two larger sizes to 10 per cent. In still larger sizes, the 40°C temperature limit held the regulation to less than 10 per cent.

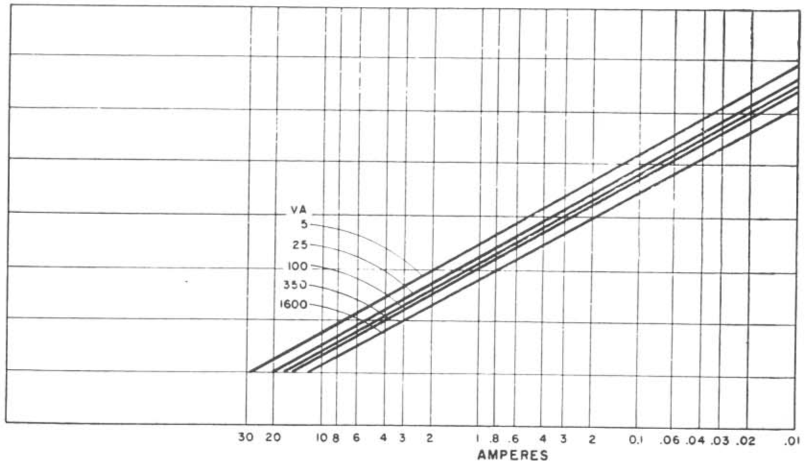
In using the chart, ratings rarely fall exactly on the v-a values assigned to each core. Hence a core is generally chosen with somewhat greater than required rating. Lower regulation and temperature rise than maximum then result. Wire size in quadrant I also increases in discrete sizes, and if the chart indication falls between two sizes the smaller size should be used.

Instructions for Using Fig. 63.

1. Choose a core from Table VIII which has a v-a rating equal to or greater than that required.
2. From rated primary and secondary voltages, find number of turns for both windings in quadrant IV.
3. From rated primary and secondary currents, find wire size for both windings in quadrant I.
4. Project turns across to quadrant III to obtain winding resistances.

Table VIII. Transformer Size, Rating, and Regulation

Core	Maximum V-A Rating	% Regulation	Total Weight (lb)	Overall Dimensions (inches)
1	5	15	0.38	1¾ × 1¾ × 1¾
2	10	15	0.68	1⅞ × 2⅜ × 1¾
3	25	10	1.2	2¼ × 2⅞ × 2¼
4	50	10	2.2	2½ × 3⅛ × 2½
5	100	8	3.8	3⅛ × 3¾ × 3
6	200	6	6.4	3⅞ × 4¾ × 3⅝
7	350	4	11.0	4⅜ × 5⅜ × 4
8	500	3	15	5⅛ × 6⅛ × 5
9	1,000	2.2	24	5⅞ × 6¾ × 6⅛
10	1,600	1.8	36	7¼ × 8¼ × 7½
11	3,200	1.2	75	9¾ × 12¾ × 8



II

III

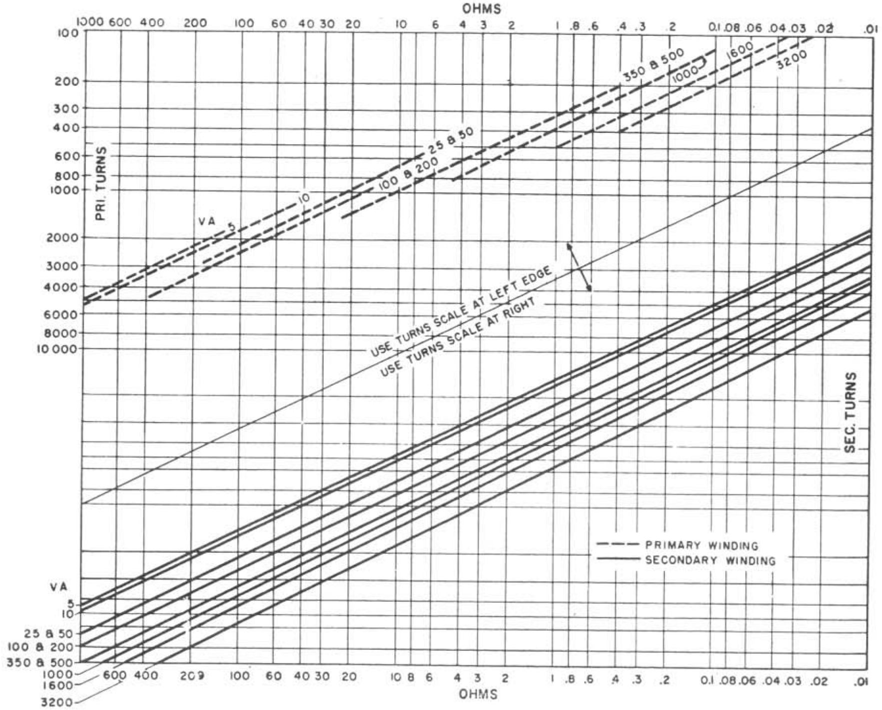
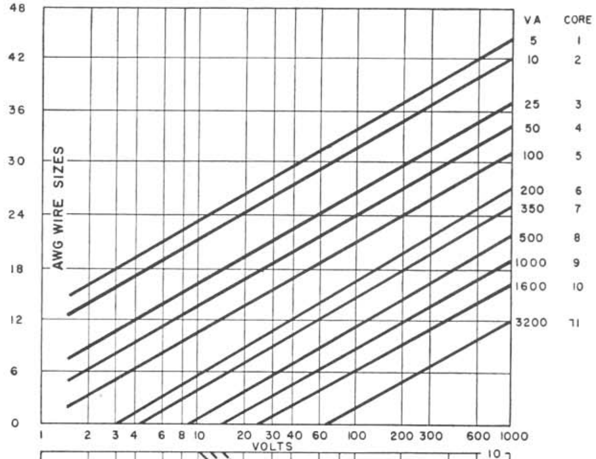


FIG. 63. Low-voltage 60-cycle transformer design chart.



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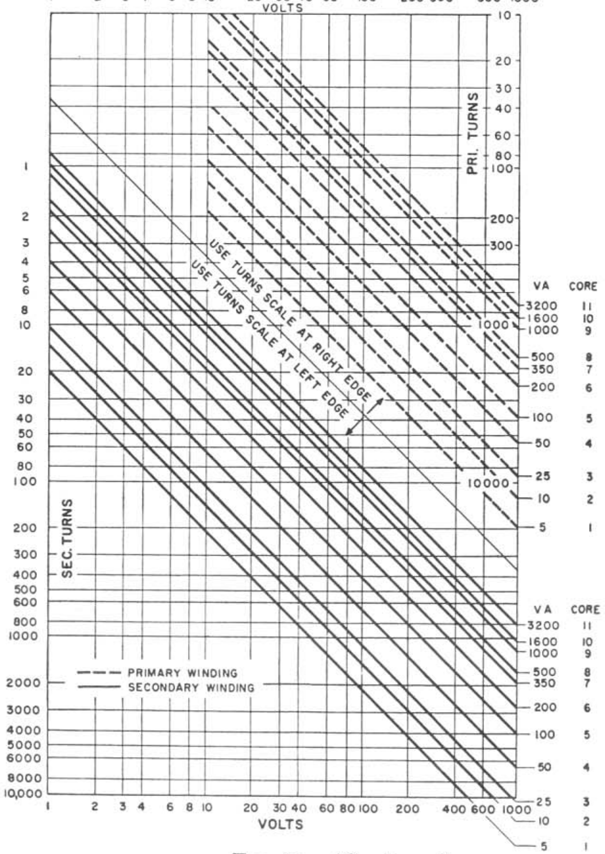

IV

FIG. 63. *(Continued)*

Departures from the assumed conditions preclude direct application of Fig. 63, but the chart is still useful as a starting point in design. For some common modifications, the following notes apply:

1. For each additional secondary winding reduce core maximum rated volt-amperes by 10 per cent. Choose wire size from quadrant II.
2. For 50-cycle transformers, reduce core maximum rated v-a 10 per cent.
3. When permissible temperature rise is higher than 40°C, core maximum volt-amperes equal (v-a in table) $\times \sqrt{\text{temperature rise}/40°\text{C}}$.

Example. A transformer is required for 115/390 volts, 60 cycles, to deliver 77 volt-amperes. This rating falls between the maxima for cores 4 and 5. Using core 5 at 115 volts, we read, from Fig. 63, for the primary, 440 turns of No. 22 wire and 3 ohms d-c resistance; for the secondary, 1,700 turns of No. 27 wire and 40 ohms d-c resistance.

37. Reactors. Reactors are used in electronic power equipment to smooth out ripple voltage in d-c supplies, so they carry direct current in the coils. It is common practice to build such reactors with air gaps in the core to prevent d-c saturation. The air gap, size of the core, and number of turns depend upon three interrelated factors: inductance desired; direct current in the winding; and a-c volts across the winding.

The number of turns, the direct current, and the air gap determine the d-c flux density, whereas the number of turns, the volts, and the core size determine the a-c flux density. If the sum of these two flux densities exceeds saturation value, noise, low inductance, and non-linearity result. Therefore a reactor must be designed with knowledge of all three of the conditions above.

Magnetic flux through the coil has two component lengths of path: the air gap l_g, and the length of the core l_c. The core length l_c is much greater geometrically than the air gap l_g, as indicated in Fig. 57, but the two components do not add directly because their permeabilities are different. In the air gap, the permeability is unity, whereas in the core its value depends on the degree of saturation of the iron. The effective length of the magnetic path is $l_g + (l_c/\mu)$, where μ is the permeability for the steady or d-c component of flux.

Reactor design is, to a large extent, the proportioning of values of air gap and magnetic path length divided by permeability. If the air gap is relatively large, the reactor inductance is not much affected by

changes in μ; it is then called a *linear* reactor. If the air gap is small, changes in μ due to current or voltage variations cause inductance to vary; then the reactor is non-linear.

When direct current flows in an iron-core reactor, a fixed magnetizing force H_{dc} is maintained in the core. This is shown in Fig. 64 as the vertical line H_{dc} to the right of zero H in a typical a-c hysteresis loop, the upper half DB_mD' of which corresponds to that in Fig. 21. Increment ΔH of a-c magnetization, superposed on H_{dc}, causes flux density increment ΔB, with permeability μ_Δ equal to the slope of dotted line AB_m. ΔB is twice the peak a-c induction B_{ac}. It will be recalled from Fig. 19 that the normal induction curve OB_m is the locus of the end points of a series of successively smaller major hysteresis loops. Since the top of the minor loop always follows the left side of a major loop, as H_{dc} is reduced in successive steps the upper ends of corresponding minor loops terminate on the normal induction curve.

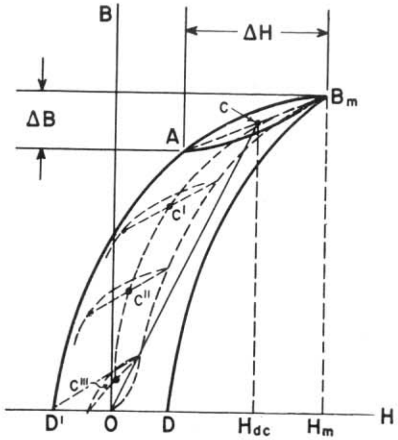

FIG. 64. Incremental permeability with different amounts of d-c magnetization.

Dotted-line slopes of a series of minor loops are shown in Fig. 64, the midpoints of which are C, C', C'', and C'''. Increment of induction ΔB is the same for each minor loop. It will be seen that the width of the loop ΔH is smaller, and hence μ_Δ is greater, as H_{dc} is made smaller.

Midpoints C, C', etc., form the locus of d-c induction. The slope of straight line OC is the d-c permeability for core magnetization H_{dc}. It is much greater than the slope of AB_m. Hence incremental permeability is much smaller than d-c permeability. This is true in varying degree for all the minor loops. The smaller ΔB is, the less the slope of a minor loop becomes, and consequently the smaller the value of incremental permeability μ_Δ. The curve in Fig. 65 marked μ is the normal permeability of 4% silicon steel for steady values of flux, in other words, for the d-c flux in the core. It is 4 to 20 times as great as the incremental permeability μ_Δ for a small alternating flux superposed upon the d-c flux. The ratio of μ to μ_Δ gradually increases as d-c flux density increases.

Because of the low value of μ_Δ for minute alternating voltages, the effective length of magnetic path $l_g + (l_c/\mu_\Delta)$ is considerably greater for alternating than for steady flux. But the inductance varies inversely as the length of a-c flux path. If, therefore, the incremental permeability is small enough to make l_c/μ_Δ large compared to l_g, it follows that small

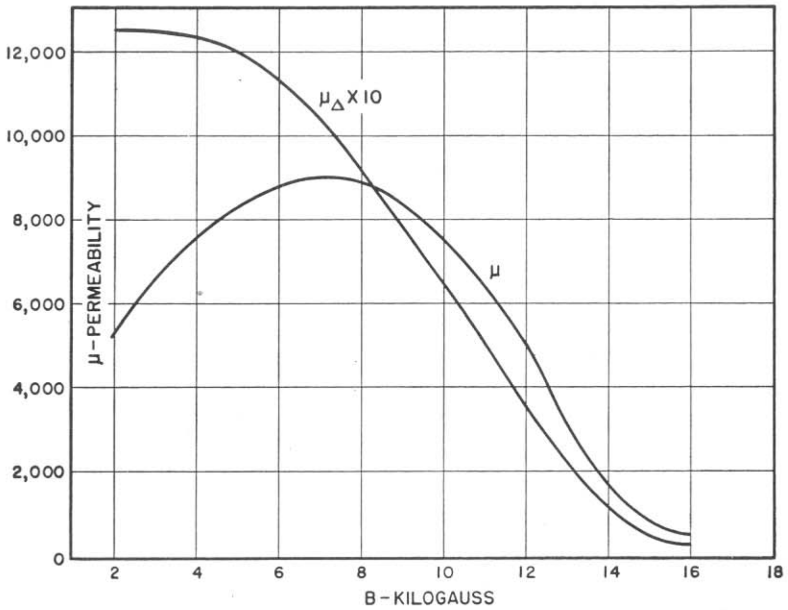

Fig. 65. Normal and incremental permeability of 4% silicon steel.

variations in l_g do not affect the inductance much. For this reason the exact value of the air gap is not important with small alternating voltages.

Reactor size, with a given voltage and ratio of inductance to resistance, is proportional to the stored energy LI^2. For the design of reactors carrying direct current, that is, the selection of the right number of turns, air gap, and so on, a simple method was originated by C. R. Hanna.[1] By this method, magnetic data are reduced to curves such as Fig. 66, plotted between LI^2/V and NI/l_c from which reactors can be designed directly. The various symbols in the coordinates are:

[1] "Design of Reactances and Transformers Which Carry Direct Current," by C. R. Hanna, *J. AIEE, 46,* 128 (February, 1927).

L = a-c inductance in henrys
I = direct current in amperes
V = volume of iron core in cubic inches
$= A_c l_c'$ (see Fig. 57 for core dimensions)
A_c = cross section of core in square inches
l_c = length of core in inches
N = number of turns in winding
l_g = air gap in inches

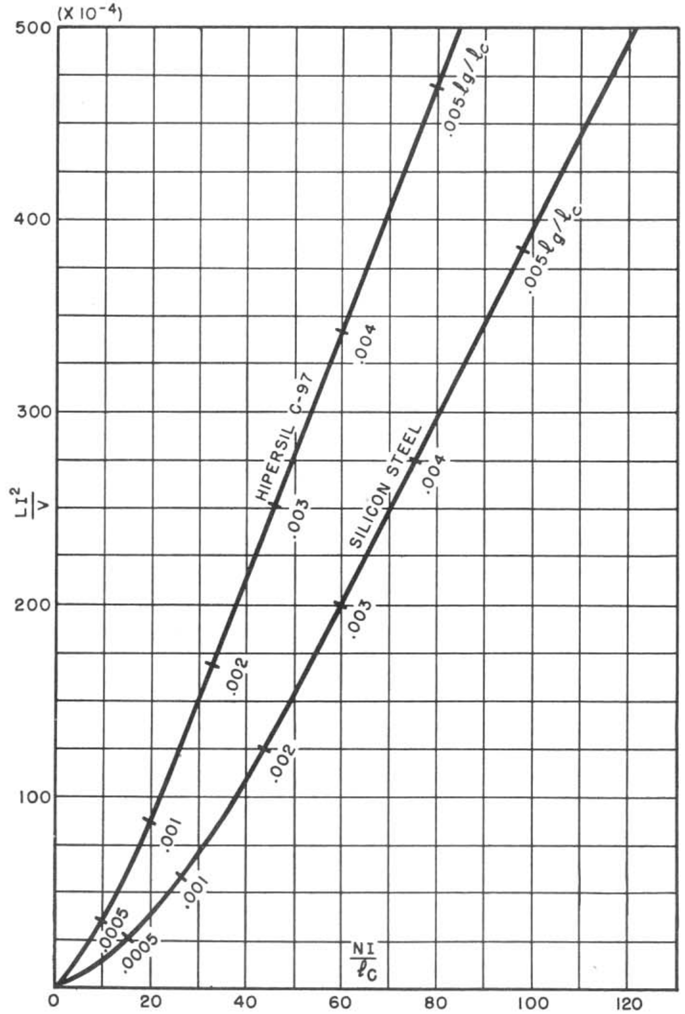

Fig. 66. Reactor energy per unit volume versus ampere-turns per inch of core.

Each curve of Fig. 66 is the envelope of a family of fixed air-gap curves such as those shown in Fig. 67. These curves are plots of data based upon a constant small a-c flux (10 gauss) in the core but a large and variable d-c flux. Each curve has a region of optimum usefulness, beyond which saturation sets in and its place is taken by a succeeding curve having a larger air gap. A curve tangent to the series of fixed air-gap curves is plotted as in Fig. 66, and the regions of optimum use-

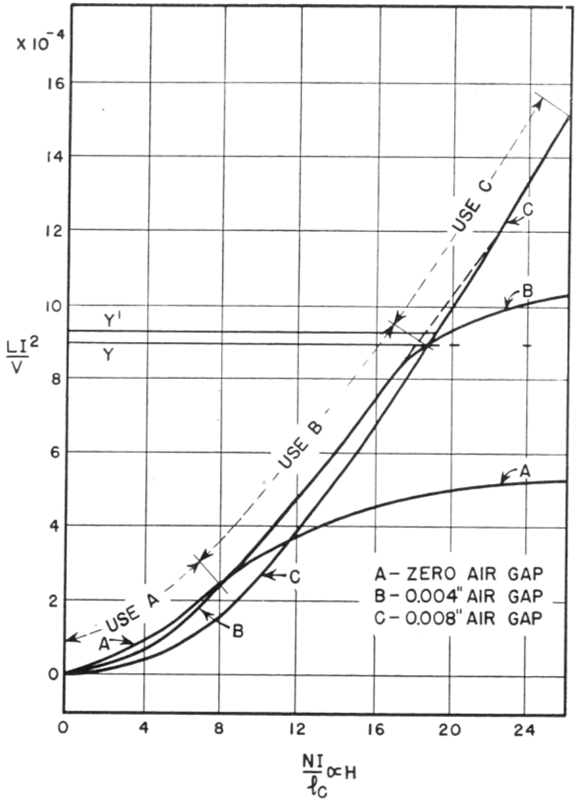

FIG. 67. Fixed air-gap curves. For $B_{dc} \gg B_{ac}$, air gap is not critical.

fulness are indicated by the scale l_g/l_c. Hence Fig. 66 is determined mainly by the d-c flux conditions in the core and represents the most LI^2 for a given amount of material.

Figure 67 illustrates how the exact value of air gap is of little consequence in the final result. The dotted curve connecting B and C is for a 6-mil gap. Point Y' represents the maximum inductance that could be obtained from a given core for $NI/l_c = 19$. Point Y is the inductance obtained if a gap of either 4 or 8 mils is used. The differ-

ence in inductance between Y and Y' is 4 per cent, for a difference in air gap of 33 per cent.

An example will show how easy it is to make a reactor according to this method.

Example. Assume a stack of silicon-steel laminations having a cross section ⅞ in. by ⅞ in., and with iron filling 92 per cent of the space. The length of the flux path l_c in this core is 7½ in. It is desired to know how many turns of wire and what air gap are necessary to produce 70 henrys when 20 ma direct current are flowing in the winding.

This problem is solved as follows:

$$A_c = (0.875)^2 \times 0.92 = 0.71 \text{ sq in.}$$

$$V = 0.71 \times 7.5 = 5.3$$

$$\frac{LI^2}{V} = \frac{70 \times 4 \times 10^{-4}}{5.3} = 53 \times 10^{-4}$$

In Fig. 66 the abscissa corresponding to $LI^2/V = 53 \times 10^{-4}$ is $NI/l_c = 25$ for silicon steel. The ratio of air gap to core length l_g/l_c is between 0.0005 and 0.001.

$$NI/l_c = 25$$

$$N = (25 \times 7.5)/0.020 = 9{,}350 \text{ turns}$$

The total air gap is nearly $0.001 \times 7\frac{1}{2}$ or 7.5 mils; the gap at each joint is half of this value, or 3.75 mils.

The conditions underlying Hanna's method of design are met in most applications. In receivers and amplifiers working at low audio levels, the alternating voltage is small and hence the alternating flux is small compared to the steady flux. Even if the alternating voltage is of the same order as the direct voltage, the alternating flux may be small, especially if a large number of turns is necessary to produce the required inductance; for a given core the alternating flux is inversely proportional to the number of turns. D-c resistance of the coil is usually fixed by the regulation or size requirements. Heating seldom affects size.

38. Reactors with Large A-C Flux. With the increasing use of higher voltages, it often happens that the a-c flux is no longer small compared to the d-c flux. This occurs in high-impedance circuits where the direct current has a low value and the alternating voltage has a high value. The inductance increases by an amount depending on the values of a-c and d-c fluxes. Typical increase of inductance is shown in Fig. 68 for a reactor working near the saturation point. Increasing a-c flux soon adds to the saturation, which prevents further inductance

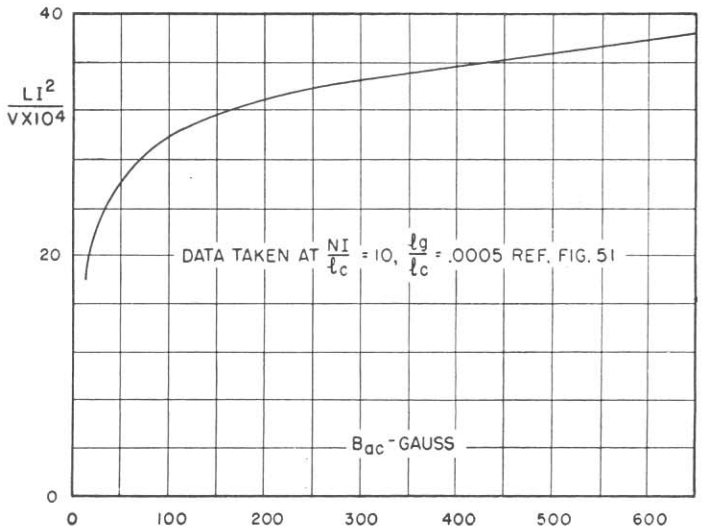

FIG. 68. Increase of inductance with a-c induction.

increase and accounts for the flattening off in Fig. 68. Saturation of this sort may be avoided by limiting the value of the d-c flux.

To illustrate the effect of these latter conditions, suppose that a reactor has already been designed for negligibly small alternating flux and operates as shown by the minor loop with center at G, Fig. 69. Without changing anything else, suppose that the alternating voltage across the reactor is greatly increased, so that the total a-c flux change is from zero to B_m. (Assume that the reactor still operates about point G.) The hysteresis loop, however, becomes the unsymmetrical figure $OB_mD'O$. The average permeability during the positive flux swing is represented by the line GB_m, and during the negative flux swing by OG. The slope of GB_m is greater than that of the minor loop; hence, the first effect exhibited by the reactor is an increase of inductance.

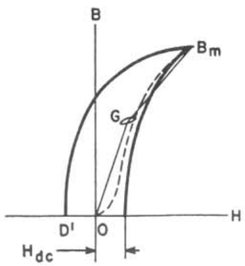

FIG. 69. Change of permeability with a-c induction.

The increase of inductance is non-linear, and this has a decided

effect upon the performance of the apparatus. An inductance bridge measuring such a reactor at the higher a-c voltage would show an inductance corresponding to the average slope of lines OG and GB_m. That is, the average permeability during a whole cycle is the average of the permeabilities which obtain during the positive and negative increments of induction, and it is represented by the average of the slopes of lines OG and GB_m. But if the reactor were put in the filter of a rectifier, the measured ripple would be higher than a calculated value based upon the bridge value of inductance. This occurs because the positive peaks of ripple have less impedance presented to them than do the negative peaks, and hence they create a greater ripple at the load. Suppose, for example, that the ripple output of the rectifier is 500 volts and that this would be attenuated to 10 volts across the load by a linear reactor having a value of inductance corresponding to the average slope of lines OG and GB_m. With the reactor working between zero and B_m, suppose that the slope of OG is 5 times that of GB_m. The expected average ripple attenuation of 50:1 becomes 16.7:1 for positive flux swings, and 83.3:1 for negative, and the load ripple is

$$\frac{1}{2}\left(\frac{500}{16.7} + \frac{500}{83.3}\right) = 18 \text{ volts}$$

or an increase of nearly 2:1 over what would be anticipated from the measured value of inductance.

This non-linearity could be reduced by increasing the air gap somewhat, thereby reducing H_{dc}. Moreover, the average permeability increases, and so does the inductance. It will be apparent that decreasing H_{dc} further means approaching in value the normal permeability. This can be done only if the maximum flux density is kept low enough to avoid saturation. Conversely, it follows that, if saturation is present in a reactor, it is manifested by a decrease in inductance as the direct current through the winding is increased from zero to full-load value.

In a reactor having high a-c permeability the equivalent length of core l_c/μ is likely to be small compared to the air gap l_g. Hence, it is vitally important to keep the air gap close to its proper value. This is, of course, in marked contrast to reactors not subject to high a-c induction.

If a choke is to be checked to see that no saturation effects are present, access must be had to an inductance bridge. With the proper values of alternating voltage across the reactor, measurements of inductance can be made with various values of direct current through it.

If the inductance remains nearly constant up to normal direct current, no saturation is present, and the reactor is suitable for the purpose. If, on the other hand, the inductance drops considerably from zero direct current to normal direct current, the reactor very probably is non-linear. Increasing the air gap may improve it; otherwise, it should be discarded in favor of a reactor which has been correctly designed for the purpose.

Filter reactors subject to the most alternating voltage for a given direct voltage are those used in choke-input filters of single-phase rectifiers. The inductance of this type of reactor influences the following:

Value of ripple in rectified output.
No-load to full-load regulation.
Transient voltage dip when load is suddenly applied, as in keyed loads.
Peak current through tubes during each cycle.
Transient current through rectifier tubes when voltage is first applied to rectifier.

It is important that the inductance be the right value. Several of these effects can be improved by the use of swinging or tuned reactors. In a swinging reactor, saturation is present at full load; therefore the inductance is lower at full load than at no load. The higher inductance at no load is available for the purpose of decreasing voltage regulation. The same result is obtained by shunt-tuning the reactor, but here the inductance should be constant from no load to full load to preserve the tuned condition.

In swinging reactors, all or part of the core is purposely allowed to saturate at the higher values of direct current to obtain high inductance at low values of direct current. They are characterized by smaller gaps, more turns, and larger size than reactors with constant inductance ratings. Sometimes two parallel gaps are used, the smaller of which saturates at full direct current. When the function of the reactor is to control current by means of large inductance changes, no air gap is used. Design of such reactors is discussed in Chapter 9.

The insulation of a reactor depends on the type of rectifier and how it is used in the circuit. Three-phase rectifiers, with their low ripple voltage, do not require the turn and layer insulation that single-phase rectifiers do. If the reactor is placed in the ground side of the circuit one terminal requires little or no insulation to ground, but the other terminal may operate at a high voltage to ground. In single-phase rectifiers the peak voltage across the reactor is E_{dc}, so the equivalent rms voltage on the insulation is $0.707E_{dc}$. But for figuring $B_{\max}$ the rms voltage is $0.707 \times 0.67E_{dc}$. Reactor voltages are discussed in Chapter 4.

39. Linear Reactor Design. A method of design for linear reactors is based on three assumptions which are justified in the foregoing:

(*a*) The air gap is large compared to l_c/μ, μ being the d-c permeability.
(*b*) A-c flux density depends on alternating voltage and frequency.
(*c*) A-c and d-c fluxes can be added or subtracted arithmetically.

From (*a*) the relation $B = \mu H$ becomes $B = H$. Because of fringing of flux around the gap, an average of $0.85B$ crosses over the gap. Hence $B_{dc} = 0.4\pi NI_{dc}/0.85l_g$. With l_g in inches this becomes

$$B_{dc} = 0.6NI_{dc}/l_g \quad \text{gauss} \tag{35}$$

Transposing equation 34

$$B_{ac} = (3.49E \times 10^6)/fA_cN \quad \text{gauss} \tag{36}$$

The sum of B_{ac} and B_{dc} is $B_{\max}$, which should not exceed 11,000 gauss for 4% silicon steel, 16,000 gauss for grain-oriented steel, or 10,000 gauss for a 50% nickel alloy. Curves are obtainable from steel manufacturers which give incremental permeability μ_Δ for various combinations of these two fluxes. Figure 70 shows values for 4% silicon steel.

By definition, inductance is the flux linkages per ampere or, in cgs units,

$$L = \frac{\phi N}{10^8 I_M} = \frac{B_{ac}A_cN}{10^8 I_M} \tag{37}$$

But

$$B_{ac} = \frac{0.4\pi NI_M}{l_g + (l_c/\mu_\Delta)}$$

If this is substituted in equation 37

$$L = \frac{3.19N^2A_c \times 10^{-8}}{l_g + (l_c/\mu_\Delta)} \quad \text{henrys} \tag{38}$$

provided that dimensions are in inches. The term A_c in equation 38 is greater than in equation 36 because of the space factor of the laminations; if the gap is large A_c is greater still because the flux across it fringes. With large gaps, inductance is nearly independent of μ_Δ, at least with moderate values of $B_{\max}$. With small gaps, permeability

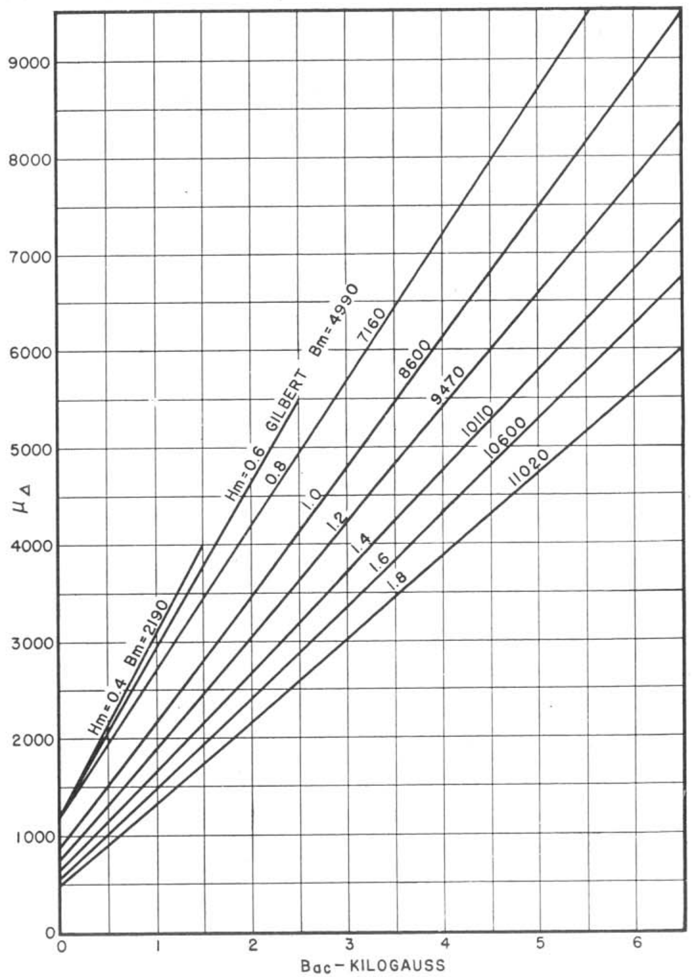

FIG. 70. Incremental permeability for 4% silicon steel with high a-c induction.

largely controls. There is always a certain amount of gap even with punchings stacked alternately in groups of 1. Table IX gives the approximate gap equivalent of various degrees of interleaving laminations for magnetic path l_c of 5.5 in.

TABLE IX. EQUIVALENT GAPS WITH INTERLEAVED LAMINATIONS

0.014-in. Laminations Alternately Stacked	Equivalent Air Gap in Inches (Total) with Careful Stacking
In groups of 1	0.0005
In groups of 4	0.001
In groups of 8	0.002
In groups of 12	0.003
In groups of 16	0.004
Butt stacking with zero gap	0.005

Example. An input reactor is required for the filter of a 1,300-volt, ¼-amp, single-phase, full-wave, 60-cycle rectifier. Let N = 2,800 turns, net A_c = 2.48 sq in., gross A_c = 2.76 sq in., l_c = 9 in., l_g = 0.050 in. The 120-cycle voltage for figuring B_{ac} is 0.707 × 0.67 × 1,300 = 605 volts.

$$B_{dc} = \frac{0.6 \times 2{,}800 \times 0.25}{0.050} = 8{,}400$$

$$B_{ac} = \frac{3.49 \times 605 \times 10^6}{120 \times 2.48 \times 2{,}800} = \underline{2{,}540}$$

$$B_{\max} = \qquad 10{,}940 \text{ gauss}$$

Figure 70 shows

$$\mu_\Delta = 2{,}650$$

$$L = \frac{3.19 \times (2{,}800)^2 \times 2.76 \times 10^{-8}}{0.050 + \frac{9}{2650}} = 13.0 \text{ henrys}$$

40. Linear Reactor Chart. In the preceding section, it was assumed that the core air gap is large compared to l_c/μ, where μ is the d-c permeability. In grain-oriented steel cores the air gap may be large compared to l_c/μ_Δ, because of the high *incremental* permeability of these cores. When this is true, variations in μ do not affect the total effective magnetic path length or the inductance to substantial degree. Reactor properties may then be taken from Fig. 71. In order to keep the reactor linear, it is necessary to limit the flux density. For grain-oriented silicon-steel cores, inductance is usually linear within 10 per cent if the d-c component of flux B_{dc} is limited to 12,000 gauss and the a-c component B_{ac} to 3,000 gauss.

Dotted lines in quadrant I are plots of turns vs. core area for a given wire size and for low-voltage coils, where insulation and margins are governed largely by mechanical considerations. Core numbers in Fig. 71 have the same dimensions and weight as in Table VIII.

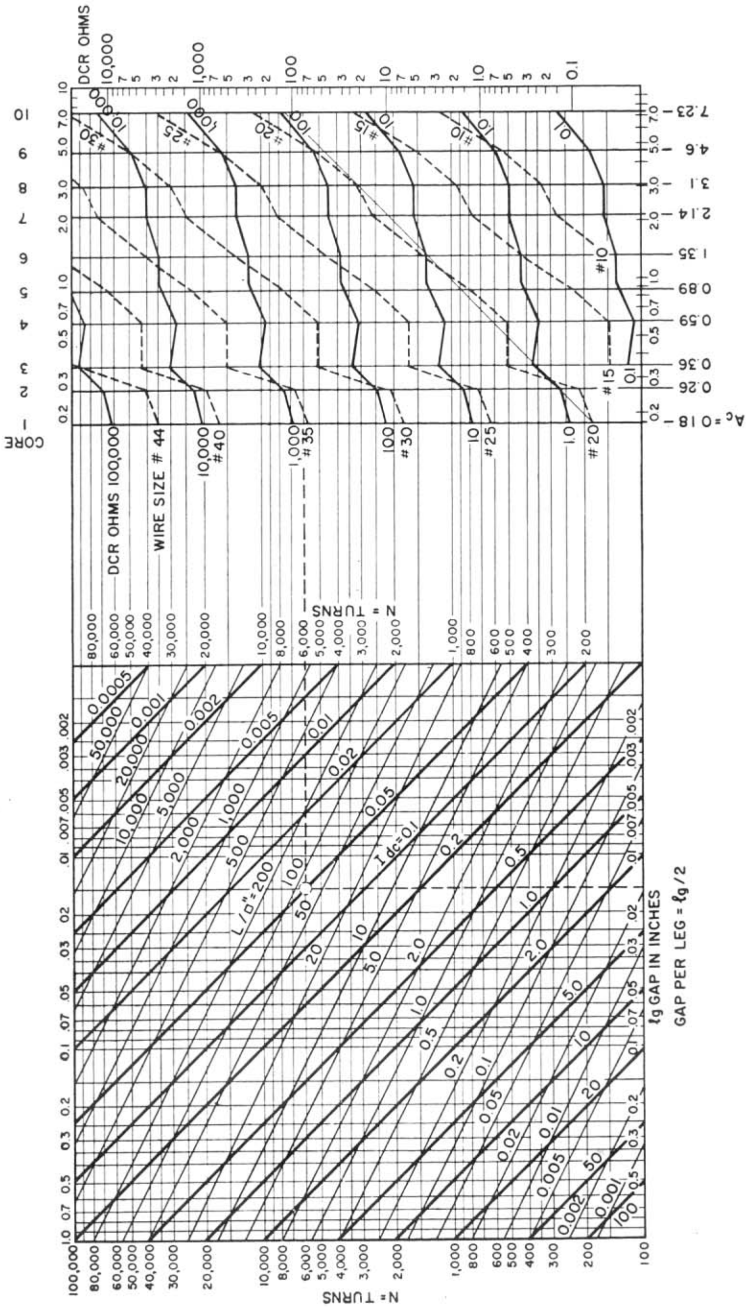

Fig. 71. Linear reactor design chart.

If the cores increased in each dimension by exactly the same amount, the lines in quadrant I would be straight. In an actual line of cores, several factors cause the lines to be wavy:

(*a*) Ratios of core window height to window width and core area deviate from constancy.
(*b*) Coil margins increase stepwise.
(*c*) Insulation thickness increases stepwise.

A-c flux density in the core may be calculated by equation 36, and B_{dc} by equation 35. If B_m materially exceeds 15,000 gauss, saturation is reached, and the reactor may become non-linear or noisy.

Instructions for Using Fig. 71.

1. Estimate core to be used.
2. Divide required inductance by area (A_c) of estimated core to obtain a value of L/sq in.
3. In second quadrant, locate intersection of L/sq in. and rated I_{dc}.
4. On this intersection, read total gap length (l_g) and number of turns (N). Gap per leg = $l_g/2$.
5. Project intersection horizontally into first quadrant to intersect vertical line which corresponds to estimated core. This second intersection gives d-c resistance and wire size.

Example. Required: 15 henrys at I_{dc} = 50 ma.
Estimate core No. 1.
L/sq in. = 84.3, l_g = 0.015 in., N = 6,000, DCR = 800 ohms.
Wire size = No. 36.
(Example shown starting with dotted circle.)

A similar chart may be drawn for silicon-steel laminations, but to maintain linearity lower values of flux density should be used.

41. Air-Gap Flux Fringing. In Section 39, equation 38 was developed for inductance of a linear reactor with an air gap. It is assumed that 85 per cent of the core flux is confined to the cross section of core face adjoining the gap. The remaining 15 per cent of the core flux "fringes" or leaves the sides of the core, thus shunting the gap. Fringing flux *decreases* the total reluctance of the magnetic path and *increases* the inductance to a value greater than that calculated from equation 38. Fringing flux is a larger percentage of the total for larger gaps. Very large gaps are sometimes broken up into several smaller ones to reduce fringing.

If it is again assumed that the air gap is large compared to l_c/μ, the reluctance of the iron can be neglected. For a square stack of punchings, the increase of inductance due to fringing is

$$\frac{L'}{L} = \left(1 + \frac{2l_g}{\sqrt{A_c}} \log_\epsilon \frac{2S}{l_g}\right) \tag{39}$$

Equation 39 is plotted in Fig. 72 with core shape $\sqrt{A_c}/S$ as abscissas and gap ratio l_g/S as parameter.[1] For large gaps, equation 35 also changes to $B_{dc} = 0.5\ L'NI_{dc}/l_gL$.

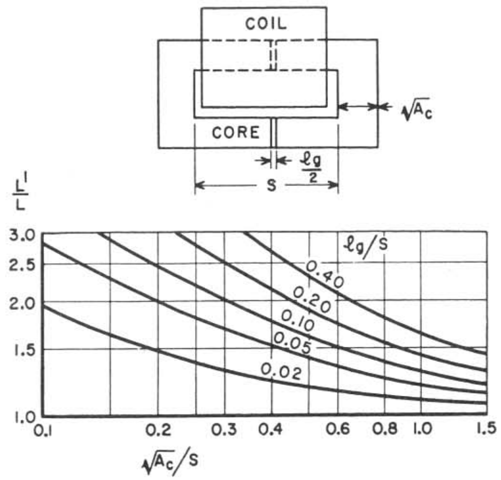

FIG. 72. Increase of reactor inductance with flux fringing at core gap.

If the air gap is enclosed by a coil, as at the top of Fig. 72, flux fringing is reduced because of the magnetizing force set up near the gap by the ampere-turns of the coil. A coil fitting tightly all around the core would produce no fringing at all. As the distance from inside of the coil to the core increases, so does the fringing. Fringing therefore depends upon the coil form thickness; if it materially exceeds the air gap per leg, fringing is nearly the same as it would be in a core gap which is not enclosed by a coil. Figure 72 is based on a thick coil form.

42. Similitude in Design. Charts such as Fig. 63 show that ratings are related to size in an orderly sequence, provided that certain proportions between core dimensions are maintained. Figure 63 is for 60

[1] See G. F. Partridge, *Phil. Mag.*, *22* (7th series), 675 (July–December, 1936).

cycles. If a transformer is desired for another frequency, its size may be estimated from Table VIII, provided that the same core proportions apply, and similar values of induction and temperature rise are used. If the new conditions are widely different, due allowance must be made for them or the estimate will not be accurate.

Table VIII and Figs. 63 and 71 are examples of *similitude*. If all variations between ratings are taken into account, similitude provides a very accurate basis for estimating new sizes; for the transformer designer there is no better basis for starting a new design.

43. Reactor Current Interruption. Sudden interruption of current through a reactor may cause high voltages to develop in the winding. This may be seen by considering the voltage across a reactor with linear inductance L and varying current i in the winding. Let current i be substituted for I_M in equation 37; it may be transposed to give

$$\phi = 10^8 Li/N \tag{37a}$$

where L is in henrys and i in amperes. If this expression for ϕ be substituted in equation 1, we obtain

$$e = -L\frac{di}{dt} \tag{40}$$

Equation 40 states that the magnitude of voltage across a reactor is equal to the inductance multiplied by the rate of current change with time. The sense or direction of this voltage is always such as to oppose the current change. Therefore, if current interruption takes place instantaneously, inductive voltage is infinitely large. In an actual reactor, losses and capacitance are always present; hence interruption of reactor current forces the reactor voltage to discharge into its own capacitance and loss resistance. The curves of Fig. 73 show how the reactor voltage e rises when steady current I flowing in the reactor is suddenly interrupted. The maximum value to which voltage e could rise under any condition is IR_2, where R_2 is the equivalent loss resistance. R_2 depends mostly on the reactor iron loss at the resonance frequency determined by reactor inductance L and capacitance C. This frequency is $1/T$, where T is $2\pi\sqrt{LC}$. Conditions for high voltage across the reactor occur with high values of k, the ratio of $\sqrt{L/C}$ to $2R_2$. If subject to sudden current interruptions, reactors must be insulated to withstand this voltage, or must be protected by spark gaps or other means. The curves of Fig. 73 are based on equation 41:

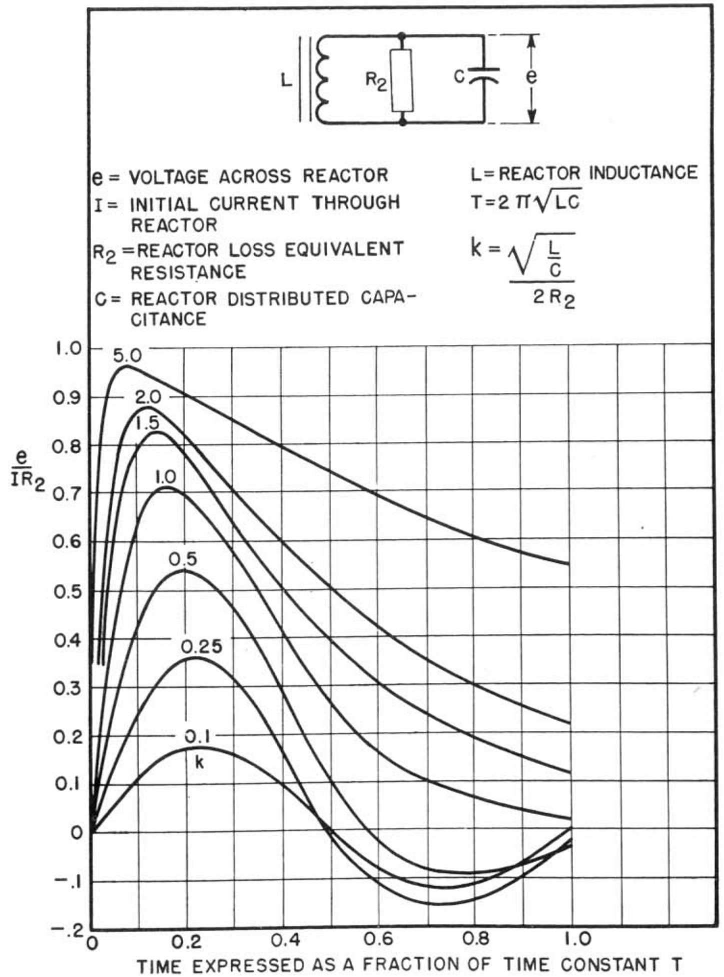

FIG. 73. Reactor voltage rise.

$$\frac{e}{IR_2} = \frac{k}{\sqrt{k^2 - 1}} (\epsilon^{m_2 t} - \epsilon^{m_1 t}) \tag{41}$$

where

$$m_1, m_2 = \frac{-2\pi}{T} (k \pm \sqrt{k^2 - 1})$$

If there is appreciable circuit or wiring capacitance shunting the reactor after it is disconnected, this contributes to the total reactor capacitance C.

44. Transformers with D-C Flux. When there is a net d-c flux in the core, as in single-phase half-wave anode transformers, the choice of core depends on the same principles as in reactors with large a-c flux. The windings carry non-sinusoidal load current, the form of which depends on the circuit. Winding currents may be calculated with the aid of Table I (p. 16). Generally the heating effects of these currents are large. Maximum flux density should be limited as described in Section 39. This precaution is essential in limited power supplies like aircraft or portable generators, lest the generator voltage wave form be badly distorted. On large power systems the rectifier is a minor part of the total load and has no influence on voltage wave form. The chief limitation then is primary winding current, and maximum induction may exceed the usual limits.

In single-phase half-wave transformers, air gaps are sometimes provided in the cores to reduce the core flux asymmetry described in Section 12. Transformers designed in this manner resemble reactors in that core induction is calculated as in Sections 37 to 41, depending on the operating conditions. Even in transformers with no air gap, there is a certain amount of incidental reluctance at the joints in both stacked laminations and type C cores. This small gap reduces the degree of core saturation that would exist in half-wave transformers with unbroken magnetic paths.

45. Power Transformer Tests. A power transformer is tested to discover whether the transformer will perform as required, or whether it will give reliable service life. Some tests perform both functions.

(*a*) *D-C Resistance.* This test is usually made on transformers at the factory as a check on the correctness of wire size in each winding. Variations are caused by wire tolerances, and by difference in winding tension between two lots of coils or between two coil machine operators. About 10 per cent variation can be expected in the d-c resistance of most coils, but this value increases to 20 per cent rather suddenly in sizes smaller than No. 40. The test is made by means of a resistance bridge or specially calibrated meter.

(*b*) *Turns Ratio.* Once the correct number of turns in each winding is established, correct output voltage can be assured for a coil of given design by measuring the turns. A simple way of doing this is by use of the turns-ratio bridge in Fig. 74. If the turns are correct, the null indicated by the meter occurs at a ratio of resistances

$$R_1/R_2 = N_1/N_2 \tag{42}$$

If there is an error in the number of turns of one winding, the null occurs at the wrong value R_1/R_2. A source of 1,000 cycles is preferable to one of 60 cycles for this test. The smaller current drawn by the transformer reduces IR and IX errors. Harmonics in the source obscure the null, and so the source should be filtered. The null is often made sharper by switching a small variable resistor in series with R_1 or R_2 to offset any lack of proportion in resistances of windings N_1 or N_2.

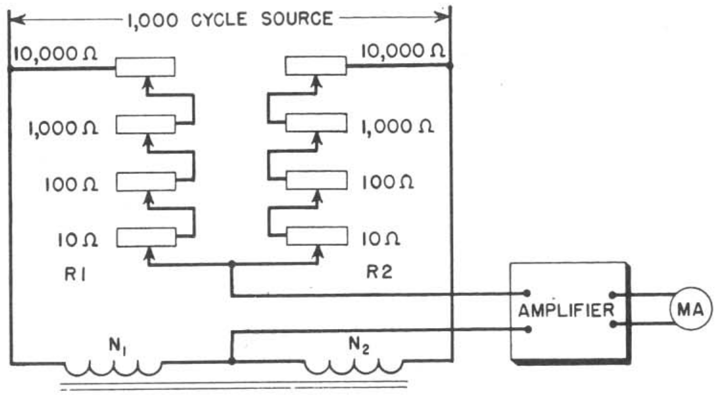

FIG. 74. Turns-ratio bridge.

An accuracy of 0.1 per cent can usually be attained with four-decade resistances. Polarity of winding is also checked by this test, because the bridge will not balance if one winding is reversed.

(c) *Open-Circuit Inductance (OCL)*. There are several ways of measuring inductance. If the Q (or ratio of coil reactance to a-c resistance) is high, the check may be made by measuring the current drawn by an appropriate winding connected across a source of known voltage and frequency. This method is limited to those cases where the amount of current drawn can be measured. A more general method makes use of an inductance bridge, of which one form is shown in Fig. 75.

If direct current normally flows in the winding, it can be applied through a large choke as shown. Inductance is then measured under the conditions of use. Source voltage should be adjustable for the same reason and should be filtered to produce a sharp null. R_c is provided to compensate for coil a-c resistance. Without it an accurate measurement is rarely attained. Enough data are provided by the test to calculate a-c resistance as well as inductance.

When Q is low, as it is in coils with high resistance, better accuracy is obtained with the Maxwell bridge, which is like the Hay bridge except that X_c and R_c are paralleled. Then the equations for bridge balance become

$$L_x = R_1R_2C \qquad R_x = R_1R_2/R_c \tag{43}$$

The Maxwell bridge has the further advantage that the null is independent of the source frequency.

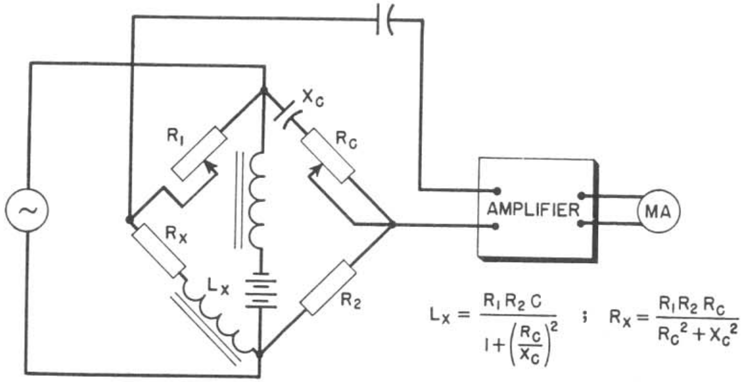

FIG. 75. Modified Hay bridge for measuring inductance.

(d) *Temperature Rise.* Tests to determine whether a transformer overheats are made by measuring the winding resistances before and after a heat run, during which the transformer is loaded up to its rating. Where several secondaries are involved, each should deliver rated voltage and current. Power is applied long enough to allow the transformer temperature to become stable; this is indicated by thermometer readings of core or case temperature taken every half hour until successive readings are the same. Ambient temperature at a nearby location should also be measured throughout the test. The average increase in winding resistance furnishes an indication of the average winding temperature. Figure 76 furnishes a convenient means for finding this temperature.

(e) *Regulation.* It is possible to measure voltage regulation by connecting a voltmeter across the output winding and reading the voltage with load off and on. This method is not accurate because the regulation is usually the difference between two relatively large quantities. Better accuracy can be obtained by multiplying the rated winding currents by the measured winding resistances and using equation 13. If the winding reactance drop is small this equation works well for resistive loads. To measure winding reactance drop, a short-circuit test is used. With the secondary short-circuited, sufficient voltage is applied to the primary to cause rated primary current to

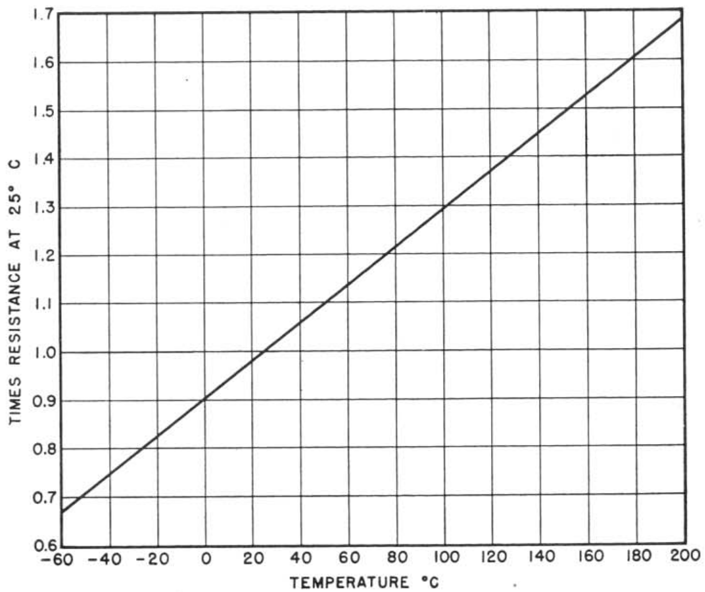

FIG. 76. Copper resistance versus temperature in terms of resistance at 25°C.

flow. The quotient E/I is the vector sum of winding resistances and reactances. Reactance is found from

$$X = \sqrt{Z^2 - R^2} \tag{44}$$

where R includes the resistance of both windings and the meter.

Sometimes it is more convenient to measure the leakage inductance with secondary short-circuited on a bridge and multiply by $2\pi f$.

(*f*) *Output Voltage.* Although the method described under (*e*) above is accurate for two-winding transformers, it is not applicable to multi-secondary transformers unless they are tested first with newly calibrated meters to see that all windings deliver proper voltage at full load. Once this is established, values of winding resistance and reactance thereafter can be checked to control the voltage. The interdependence of secondary voltages when there is a common primary winding makes such an initial test desirable. This is particularly true in combined filament and plate transformers, for which the best test is the actual rectifier circuit.

(*g*) *Losses.* Often it is possible to reduce the number of time-consuming heat runs by measuring losses. The copper loss is readily calculated by multiplying the measured values of winding resistance (corrected for operating temperature) by the squares of the respective rated currents. Core loss is measured with open secondary by means of a low-reading wattmeter at rated voltage in the primary circuit. If these losses correspond to the allowable temperature rise, the transformer is safely rated.

(*h*) *Insulation.* There is no test to which a transformer is subjected which has such a shaky theoretical basis as the insulation test. Yet it is the one test it must pass to be any good. Large quantities of transformers can be built with little or no insulation trouble, but the empirical nature of standard test voltages does not assure insulation adequacy. It has been found over a period of years that, if insulation withstands the standard rule of twice normal voltage plus 1,000 volts rms at 60 cycles for 1 minute, reasonable insulation life is usually obtained. It is possible for a transformer to be extremely under-insulated and still pass this test (see p. 44); conversely, there are conditions under which the rule would be a handicap. Therefore it can only be considered as a rough guide.

The manner of making insulation tests depends upon the transformer. Low-voltage windings categorically can be tested by short-circuiting the terminals and applying the test voltage from each winding to core or case with other windings grounded. Filament transformers with secondaries insulated for high voltage may be tested in similar manner. But a high-voltage plate transformer with grounded center tap requires unnecessary insulation if it is tested by this method. Instead, a nominal voltage of, say, 1,500 volts is applied between the whole winding and ground; after that the center tap is grounded and a voltage is applied across the primary of such value as to test the end terminals at twice normal plus 1,000 volts. Similar test values can be calculated for windings operating at d-c voltages other than zero. Such a test is called an induced voltage test. It is performed at higher than normal frequency to avoid saturation. An advantage of induced voltage testing is that it tests the layer insulation.

If insulation tests are repeated one or more times they may destroy the insulation, because insulation breakdown values decrease with time. Successive applications of test voltage are usually made at either decreased voltage or decreased time. In view of their dubious value, repeated insulation tests are best omitted.

Corona tests are not open to this objection. A voltage 5 per cent higher than normal is applied to the winding, and the leads are run through blocking capacitors to the input of a sensitive radio receiver as in Fig. 38.[1] RETMA standard noise values for this test are based primarily on radio reception, but they do indicate whether standard insulation practice is followed. See Table X.

TABLE X. CORONA VOLTAGE

RMS Working Voltage (kilovolts)	Corona Level (microvolts)
Up to 8.6	1,000
8.61 to 15	2,500

Transformers which are subjected to voltage surges may be given impulse tests to determine whether the insulation will withstand the surges. Power line surges are the most difficult to insulate for. The electric power industry has standardized on certain impulse voltage magnitudes and wave shapes for this testing.[2] The ratio of impulse voltage magnitude to 60-cycle, 1-minute insulation test voltage is called the *impulse ratio*. This ratio is much greater for oil-insulated transformers than for dry-type transformers, and is discussed further in Chapter 4.

[1] See RETMA Standard TR-102-B, "Power Transformers for Radio Transmitters."

[2] See ASA Standard C57.22-1948, paragraph 22.116.

4. RECTIFIER PERFORMANCE

46. Ripple. Filters used with rectifiers allow the rectified direct current to pass through to the load without appreciable loss, but ripple in the rectified output is attenuated to the point where it is not objectionable. Filtering sometimes must be carried out to a high degree. From the microphone to the antenna of a high-power broadcast station, there may be a power amplification of 2×10^{15}. The introduction of a ripple as great as 0.005 per cent of output voltage at the microphone would produce a noise in the received wave loud enough to spoil the transmitted program. A rectifier used at the low-power levels must be unusually well filtered to prevent noticeable hum from being transmitted.

Different types of rectifiers have differing output voltage waves, which affect the filter design to a large extent. Certain assumptions, generally permissible from the standpoint of the filter, will be made in order to simplify the discussion. These assumptions are:

1. The alternating voltage to be rectified is a sine wave.
2. The rectifying device passes current in one direction but prevents any current flow in the other direction.
3. Transformer and rectifier voltage drops are negligibly small.
4. Filter condenser and reactor losses are negligible.

47. Single-Phase Rectifiers. Single-phase half-wave rectified voltage across a resistive load R is shown in Fig. 77. It may be resolved by Fourier analysis into the direct component whose value is $0.318E_{pk}$ or $0.45E_{ac}$, and a series of alternating components. The fundamental alternating component has the same frequency as that of the supply.

Single-phase half-wave rectifiers are used only when the low average value of load voltage and the presence of large variations in this voltage are permissible. The chief advantage of this type of rectifier is its simplicity. A method of overcoming both its disadvantages is illustrated in Fig. 78 where a capacitor C shunts the load. By using the proper capacitor, it is often possible to increase the value of E_{dc} to