CHAPTER 5.

TRANSFORMERS AND IRON-CORED INDUCTORS

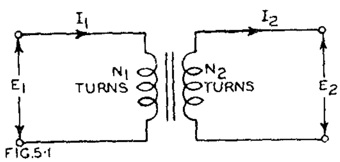
by G. Builder, B.Sc., Ph.D., F.Inst.P.,
I. C. Hansen, Member I.R.E. (U.S.A.), and
F. Langford-Smith, B.Sc., B.E.

SECTION 1 : IDEAL TRANSFORMERS

(i) Definitions (ii) Impedance calculations—single load (iii) Impedance calculations—multiple loads.

(i) Definitions

An **ideal transformer** is a transformer in which the winding reactances are infinite, and in which winding resistances, core loss, leakage inductances and winding capacitances are all zero. In such a transformer the voltage ratio between any two windings is equal to the turns ratio of the windings, under all conditions of loading, as illustrated in Fig. 5.1. Also, in such a transformer the currents in any two windings are inversely proportional to the ratio of turns in the windings under all load conditions.



$$\frac{E_2}{E_1} = \frac{N_2}{N_1}$$

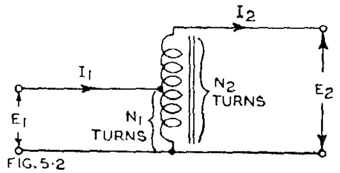
$$\frac{I_2}{I_1} = \frac{N_1}{N_2}$$

Fig. 5.1. Ideal two-winding transformer. E_1 and E_2 are alternating voltages (r.m.s.). I_1 and I_2 are alternating currents (r.m.s.) but the indicated directions of current flow are at a chosen instant and correspond to the direction of voltage at that instant. Similar remarks apply to Fig. 5.2.

Modern iron-cored transformers often approach so closely to perfection for their particular purposes that their analysis on the basis of ideal transformer theory may give useful practical approximations for design purposes.

A **double-wound transformer** is one in which, as illustrated in Fig. 5.1, separate primary and secondary windings are used to permit isolation of the primary and secondary circuits except through mutual inductive coupling.

Auto-transformers may be used with economy in some cases : a single winding is tapped to give the required turns ratio, which may be greater or less than unity, between primary and secondary. An ideal step-up auto-transformer is shown in Fig. 5.2.



$$\frac{E_2}{E_1} = \frac{N_2}{N_1}$$

$$\frac{I_2}{I_1} = \frac{N_1}{N_2}$$

Fig. 5.2. Ideal step-up auto-transformer.

An auto-transformer is always more economical to construct than a two winding transformer as indicated in the following tabulation :

Type of Winding	Ratios	Total Volt-ampere rating of windings	Transformer Output V.A.
Auto	10 : 9 or 9 : 10	20	100
Auto	2 : 1 or 1 : 2	100	100
Auto	3 : 1 or 1 : 3	133	100
Auto	5 : 1 or 1 : 5	160	100
Auto	10 : 1 or 1 : 10	180	100
Double	any ratio	200	100

The currents in the primary and secondary sections are exactly 180° out of phase, and the resultant current flowing through the common portion of the winding is the difference between the two. When the ratio is 1 : 2 or 2 : 1, the currents in the two sections of the winding are equal.

In ideal transformer theory there is no distinction between auto- and two-winding transformers and they need not therefore be considered separately in this section.

(ii) Impedance calculations—single load

When the secondary of a simple two-winding ideal transformer is loaded with a resistance R_2 as shown in figure 5.3, the equivalent or **transformed load R_1** as measured between the primary terminals is

$$R_1 = (N_1/N_2)^2.R_2 \qquad (1)$$

and since the voltage ratio between primary and secondary is equal to the turns ratio, this may be written as

$$R_1 = (E_1/E_2)^2.R_2 \qquad (2)$$

A transformer therefore merely transforms a load imposed on its secondary. Its primary does not in the ideal case impose any load unless a load is applied to the secondary. It is the turns ratio between primary and secondary, and not the number of turns in the primary, that governs the transformed or reflected load impedance.

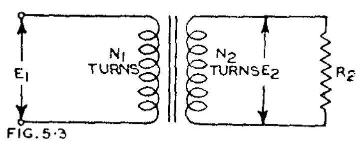


Fig. 5.3. Ideal two-winding transformer with loaded secondary.

If however a transformer is to approximate to the ideal, the number of turns in the primary must be sufficient to make the reactance of the primary high compared with the transformed value of the load impedance as measured across the primary terminals.

For an ideal transformer with a centre-tapped primary as shown in Fig. 5.4, the transformed load measured across the whole primary (between terminals, P, P) is equal to the transformed value R_1. *If however only one half of the primary is used* (between either of terminals P and terminal C.T.) the transformed load presented is $\frac{1}{4} R_1$.

$$\frac{E_1}{E_2} = \frac{N_1}{N_2}$$
$$R_1 = (N_1/N_2)^2.R_2 = (E_1/E_2)^2.R_2$$

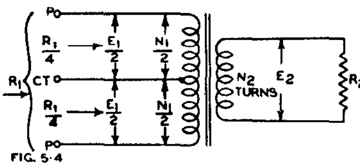
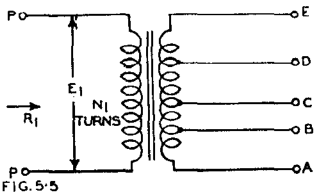


Fig. 5.4. Ideal transformer with primary centre-tap and loaded secondary.

As a practical example of the primary centre tap, consider the use of the transformer of Fig. 5.4 to feed a 500 ohm line (R_2 = 500 ohms). If the transformer has an impedance ratio $(N_1/N_2)^2$ equal to 10 : 1, the transformed load across the whole of the primary, e.g., when the primary is fed by a push-pull amplifier, is 10 × 500 = 5000 ohms. If, however, only one half of the primary were used for connection to a single-ended amplifier, the load presented to the amplifier would be 1250 ohms.

For an ideal transformer with a winding tapped for load matching, as shown in Fig. 5.5, the calculation of the tap to be selected for any particular load follows from the application of eqn. (1).



$$R_1 = (N_1/N_{AB})^2.R_{AB}$$
$$\text{or } R_2 = (N_1/N_{BD})^2.R_{BD}$$
and so on.

Fig. 5.5. Ideal transformer with secondary tapped for load matching.

In these equations a load connected across terminals A and B is denoted by R_{AB} and the number of turns in the secondary between these two terminals is given by N_{AB}, and corresponding designations apply for any pair of terminals across which the load is connected.

If, for example, the terminals A, B provide a match with a 10 ohm secondary load with a total of N_{AB} secondary turns, the number of turns N_{AD} required between terminals A and D to provide a similar match to a 500 ohm line is given simply by

$$(N_{AD}/N_{AB})^2 = 500/10 = 50, \text{ so that}$$
$$N_{AD}/N_{AB} = \sqrt{50} = 7.07.$$

The number of turns in the 10 ohm winding is approximately one-seventh of the number of turns for matching to 500 ohms. It is, of course, permissible to use any pair of secondary terminals such as B, C or C, D and so on, so that a wide range of transformation ratios is available from a transformer arrangement such as that shown in Fig. 5.5.

(iii) Impedance calculations—multiple loads

Where two or more loads are connected simultaneously to the windings of a transformer, the conditions for matching may be determined readily by the following methods.

Consider an ideal transformer having two secondaries of N_2 and N_3 turns connected to loads R_2 and R_3 respectively as shown in Fig. 5.6. It follows immediately from eqn. (2) that

$$E_2/E_1 = N_2/N_1\,;\ E_3/E_1 = N_3/N_1\,;\ E_3/E_2 = N_3/N_2 \tag{3}$$

and, since the transformer is an ideal one, these relations hold irrespective of the relative values of the loads.

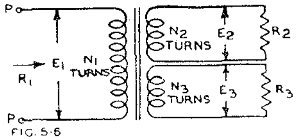
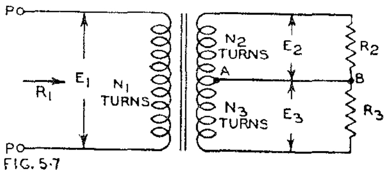


Fig. 5.6. Ideal transformer with multiple loads.

It is sometimes convenient to draw the diagram as in Fig. 5.7, which is equivalent in every way to Fig. 5.6. If, as a special case, $N_2/N_3 = R_2/R_3$, then the voltages of the two points A and B will be the same, and the link AB may be omitted without any effect. The currents in both sections of the transformer and in R_2 and R_3 will then be the same.

The value of the reflected or transformed load R_1 measured across the primary terminals can be obtained by considering the power in relation to the voltages specified in eqn. (3).

Fig. 5.7 Ideal transformer with multiple loads as in Fig. 5.6, but with the secondaries connected at points A, B.



Let W_1 = total input watts to the primary $= E_1{}^2/R_1$
W_2 = watts in the load R_2 $= E_2{}^2/R_2$
W_3 = watts in the load R_3 $= E_3{}^2/R_3$

Then $W_1 = W_2 + W_3$ (since there are no transformer losses)

so that
$$\frac{E_1{}^2}{R_1} = \frac{E_2{}^2}{R_2} + \frac{E_3{}^2}{R_3} = \frac{E_1{}^2}{R_2}\left(\frac{N_2}{N_1}\right)^2 + \frac{E_1{}^2}{R_3}\left(\frac{N_3}{N_1}\right)^2.$$

We therefore have

$$\frac{1}{R_1} = \frac{1}{R_2(N_1/N_2)^2} + \frac{1}{R_3(N_1/N_3)^2} \tag{4}$$

so that the total load R_1 presented by the primary is equal to the parallel combination of the two transformed loads $R_2(N_1/N_2)^2$ and $R_3(N_1/N_3)^2$.

If an additional winding of N_4 turns is connected to a load R_4 we obviously have in the same way

$$\frac{1}{R_1} = \frac{1}{R_2(N_1/N_2)^2} + \frac{1}{R_3(N_1/N_3)^2} + \frac{1}{R_4(N_1/N_4)^2} \tag{5}$$

and so on, for any number of loads. Such expressions are equally applicable whether the secondary windings used are separate windings or whether they form part of a single tapped secondary. For example if, in the transformer shown in Fig. 5.5, we have loads R_{AB}, R_{AC} and R_{BD} connected to terminals AB, AC, and BD respectively, we will have

$$\frac{1}{R_1} = \frac{1}{R_{AB}(N_1/N_{AB})^2} + \frac{1}{R_{AC}(N_1/N_{AC})^2} + \frac{1}{R_{BD}(N_1/N_{BD})^2}$$

and so on.

A typical practical case is one in which a known power output W_1 from an amplifier is fed into a known reflected impedance R_1 with two secondaries feeding loads R_2

and R_3 (such as two loudspeakers) which are required to operate with power inputs of W_2 and W_3 respectively, so that $W_1 = W_2 + W_3$. The required transformer turns ratios are then given by

$$R_1W_1 = E_1^2; \; R_2W_2 = E_2^2; \; R_3W_3 = E_3^2 \qquad (6a)$$

so that $R_2W_2/R_1W_1 = E_2^2/E_1^2 = N_2^2/N_1^2$ (6b)

and $R_3W_3/R_1W_1 = E_3^2/E_1^2 = N_3^2/N_1^2$ (6c)

For example if $W_2 = 3$ watts $R_2 = 500$ ohms
$W_3 = 4$ watts $R_3 = 600$ ohms
and $W_1 = 7$ watts $R_1 = 7{,}000$ ohms
we have

$$\frac{N_2^2}{N_1^2} = \frac{500 \times 3}{7000 \times 7} = \frac{1}{32.7}. \quad \text{Therefore } \frac{N_2}{N_1} = \frac{1}{5.7}$$

and similarly $N_3/N_1 = 1/4.5$.

Expressions such as (6) may be written in the more general form

$$N_n/N_1 = \sqrt{(R_n/R_1)(W_n/W_1)} \qquad (7)$$

where N_n = number of turns on secondary n,
N_1 = number of turns on primary,
R_n = load applied to secondary n,
R_1 = transformed total primary load,
W_n = watts in load R_n,
and W_1 = total watts input to primary,
but it must be noted that these relations hold only when all loads are connected so that the specified input conditions to the primary do exist. Eqn. (7) is also applicable to determine the turns ratios of two or more separate transformers feeding two or more loads from a common amplifier which delivers W_1 watts into a total load of R_1 ohms.

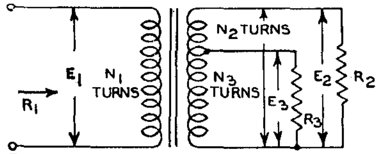


Fig. 5.8. *Ideal transformer with multiple loads and tapped secondary, which is effectively identical with Fig.* 5.6.

If a transformer is supplying power to two or more loads, such as loudspeakers, and one of these is switched out of circuit, the impedance reflected on to the primary will change due to the reduction of loading on the secondary. In order to avoid the resultant mismatching it is advisable to switch in a resistive load, having a resistance equal to the nominal (400 c/s) impedance of the loudspeaker, so as to take the place of the loudspeaker which has been cut out of circuit. In this case the resistance should be capable of dissipating the full maximum power input to the loudspeaker. Such an arrangement will also have the result that the volume from the remaining speakers will be unchanged

Alternatively if it is desired to switch off one loudspeaker and to apply the whole power output to a single speaker, it will be necessary to change the number of secondary turns so as to give correct matching. This change may generally be arranged quite satisfactorily by the use of a tapped secondary winding. In this case the loudspeaker would be used on the intermediate tap when both speakers are in use, and on the whole winding for single speaker operation.

It does not matter whether two or more secondary windings or a single tapped winding is employed. The arrangement shown in Fig. 5.8 is effectively identical with that of Fig. 5.6.

SECTION 2 : PRACTICAL TRANSFORMERS

(i) General considerations (ii) Effects of losses.

(i) General considerations

The treatment in section 1 based on ideal transformer theory is an extremely useful first approximation in design problems, particularly if the transformers to be used are so liberally designed that their general characteristics approximate to the ideal.

In practice it is usually necessary to take into account :

(a) The resistance of each winding
(b) The core loss
(c) The inductances of the windings
(d) The leakage inductances
(e) The capacitances between windings and between each winding and ground, and the self capacitance of each winding.

A useful equivalent circuit of a practical transformer is shown in Fig. 5.9.

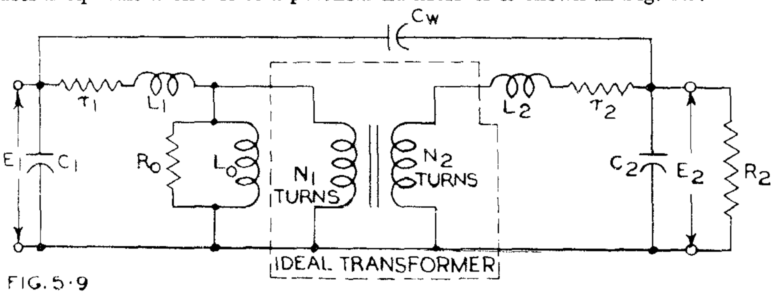


Fig. 5.9. Equivalent circuit of a practical transformer.

In this equivalent circuit we have an ideal transformer with a turns ratio N_2/N_1 (equal to the turns ratio of the actual transformer), with the incidental characteristics of the actual transformer represented by separate reactances and resistances.

r_1 = the resistance of the primary winding
L_1 = the equivalent primary leakage inductance
r_2 = the resistance of the secondary winding
L_2 = the equivalent secondary leakage inductance
R_0 = the equivalent core-loss resistance (including both hysteresis and eddy current losses)
L_0 = the inductance of the primary winding
C_1, C_2 = the primary and secondary equivalent lumped capacitances
C_w = the equivalent lumped capacitance between windings
R_1 = input resistance of transformer on load
and R_2 = the load resistance across the secondary.

Such an equivalent circuit is capable of representing a practical design with considerable accuracy, but actual calculations would be tedious and in some cases very difficult.

The reactances and resistances shown therein have varying effects on the input-output voltage ratio according to the frequency of the signal which the transformer is handling(Ref. A10). In general, the equivalent circuit can be presented in three distinct simplified forms for use when considering the transformer operating at low, medium and high frequencies respectively (Figs. 5.10B,C,D).

Audio transformers can be conveniently dealt with in this manner, whereas power transformers operating over a very limited frequency range can be more simply designed on the basis of Fig. 5.10A.

The whole of the equivalent circuit of Fig. 5.10A can be referred to the primary, as in Fig. 5.11 where the ideal transformer has been omitted and r_2 and R_2 multiplied by the square of the turns ratio. This is often a convenient way of making calculations.

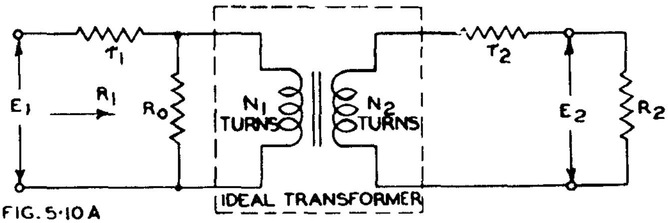
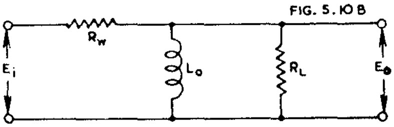
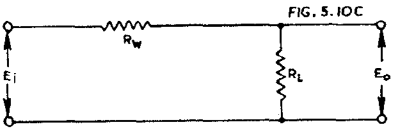


Fig. 5.10A. *Simplified equivalent circuit for calculating the effect of losses.*





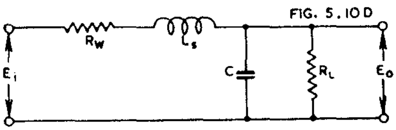


Fig. 5.10B. *Low frequency equivalent circuit.*

Fig. 5.10C. *Medium frequency equivalent circuit.*

Fig. 5.10D. *High frequency equivalent circuit.*

R_w = *Total winding resistance referred to primary*

L_s = *Total leakage inductance referred to primary*

C = *Total winding capacitance referred to primary*

R_L = *Secondary load resistance referred to primary*

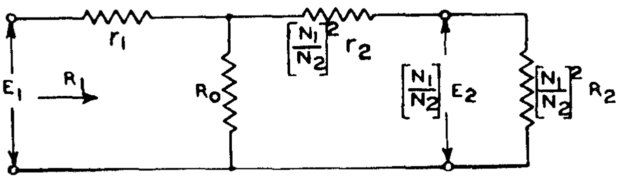


FIG. 5·11

Fig. 5.11. *Simplified equivalent circuit referred to the primary.*

(ii) The effects of losses

If a transformer is designed to deliver power into a resistance load, such as valve heaters, the primary and secondary copper losses are usually designed to be of the same order.

A particular transformer reaches its maximum efficiency when the copper losses have become equal to the iron losses (proof given in Ref. A10), although this does not necessarily occur at full load unless the transformer is so designed.

The **efficiency** is the ratio of the output power to the output power plus the losses. For audio and power transformers for radio purposes typical efficiencies range from 70 to 95 per cent, with the majority of power transformers falling between 80 and 90 per cent.

The **regulation** of a transformer is defined differently in American and British practice, the primary voltage being held constant throughout.

American definition. The percentage voltage regulation is the difference between the full-load and the no-load secondary voltages, divided by the **full-load** voltage and multiplied by 100.

British definition. The percentage voltage regulation is the difference between the full-load and the no-load secondary voltages, divided by the **no-load** secondary voltage and multiplied by 100 (B.S.205 : 1943).

However, the difference between the two definitions is quite small for small percentages of voltage regulation.

The regulation of audio transformers when operating over a limited frequency range (say 200-2000 c/s), that is the ratio of E_2/E_1, is affected mainly by the copper losses.

Thus $E_2/E_1 \approx \eta T$ and $I_2/I_1 \approx 1/T$, (1)
where η = efficiency
and T = turns ratio.

As a further consequence the impedance ratio is changed, and, making the same assumptions as above,

$$R_1 \approx r_1 + r_2/T^2 + R_2/T^2 \tag{2}$$

At low frequencies, the reflected impedance is altered by the shunt effect of the primary inductance, while at high frequencies, a similar change is caused by the leakage inductance and winding capacitances. This is covered in Sect. 3(iii).

The damping of a loudspeaker, connected as a load to the secondary of an output transformer, is also affected to some degree by the losses. Where an accurate indication of the damping factor is required, these losses should be taken into account. Refer to Chapter 21 Sect. 3.

SECTION 3 : AUDIO-FREQUENCY TRANSFORMERS

(i) General considerations (ii) Core materials (iii) Frequency response and distortion—(a) Interstage transformers (b) Low level transformers (c) Output transformers (iv) Designing for low leakage inductance (v) Winding capacitance (vi) Tests on output transformers (vii) Specifications for a-f transformers.

(i) General considerations (References A6 and A10)

Audio frequency transformers can be divided into three major categories,

(a) Low level input
(b) Medium level interstage
(c) Output.

For design purposes it is necessary to know

(1) Operating level, usually expressed in db above or below a reference level of 1 milliwatt (i.e. dbm),
(2) Frequency response, with permitted deviation, quoting reference level at which measurements are made,
(3) Permissible distortion, at specified operating levels and frequencies,
(4) Impedance, phase angle, and nature of source and load between which transformer is to be connected,
(5) Value of d.c. (if any) flowing through any winding,
(6) Hum reduction requirements,
(7) Phase shift permitted.

Operating level.

This restricts the choice of core materials and determines the physical size. Suitable core alloys include *MUMETAL* and *RADIOMETAL* together with silicon steels in various grades. As *MUMETAL* saturates at a comparatively low flux density, it is only suitable for low level transformers. Economic factors will probably dictate which material is finally used.

Frequency response

At low frequencies the response falls off due to the finite value of primary inductance. At high frequencies, the winding capacitance and leakage inductance are responsible for the response limitations.

Distortion (a) **Low frequency**

This is mainly dependent on the maximum operating flux density at the lowest frequency of interest. Distortion due to this cause falls off rapidly with increase in frequency. Other sources of distortion are observed when transformers operate in valve plate circuits. The drop in valve load impedance due to the shunting effect of the primary inductance may cause the valve to distort. Further, the load impedance will also change to a partially reactive one at low frequencies. The valve will therefore operate with an elliptical loadline and will introduce additional distortion unless care is taken.

A simple method of measuring the harmonic distortion in the cores of a-f transformers is described in Ref. C33.

Distortion (b) **High frequency**

At this end of the audio spectrum the load impedance changes again in magnitude and sign, thus causing an associated valve to generate distortion.

Source and load

Before the design of a transformer can be proceeded with, something must be known about the impedances between which the transformer is required to operate. Assuming that it is a low level unit, for example, it may be intended to operate from a ribbon microphone, a low impedance line, a gramophone pickup or the plate of a valve. It may have to feed the grid of a pentode or triode, a line or a mixer circuit. The secondary may be shunted by a resistance or a frequency correcting network. Unless these factors concerning the external circuits are known it is not possible to predict with any degree of accuracy, the ultimate performance of the transformer.

D.C. polarization

If an unbalanced d.c. component is present in one of the windings, this will cause a reduction in inductance over that attainable without d.c. This would require a larger transformer to meet the same performance specifications. In some cases where the d.c. magnetizing force is high, the use of high permeability alloys is not feasible. As far as practicable, unbalanced d.c. should be avoided in transformers, either by using a push-pull connection or shunt feed. For calculation of primary inductance with a d.c. component present, refer to Section 6.

Hum reduction

When the operating level is very low, it may be found desirable to shield the transformer to decrease the hum level to a suitable magnitude. This can be achieved in several ways, *MUMETAL* shields up to three in number being particularly effective. An outer case of sheet metal or cast iron is normally employed. The use of a balanced structure such as a core type instead of a shell type lamination will assist in reducing the effects of extraneous a.c. fields.

Phase shift

In certain applications it is desirable to apply negative feedback over an amplifier incorporating one or two transformers. To achieve stability with the desired amount of gain reduction, it is necessary to exercise careful control of phase shift over a frequency range very much wider than the nominal frequency range of the transformer (see Chapter 7 Sect. 3).

(ii) Core materials

High permeability alloys are now produced by several manufacturers under a variety of trade names. These are listed below, with silicon steel shown for comparison.

See also Ref. C34.

NICKEL IRON AND OTHER ALLOYS

Material	ρ	μ_0 d.c.	μ_{max} d.c.	(B-H) Sat. Gauss	Lines/sq.in.
Mumetal	62	30 000	130 000	8 500	55 000
Permalloy C	60	16 000	75 000	8 000	52 000
Radiometal	55	2 200	22 000	16 000	103 000
Permalloy B	45	2 000	15 000	16 000	103 000
Permalloy A	20	12 000	90 000	11 000	71 000
Cr-Permalloy	65	12 000	60 000	8 000	52 000
Mo-Permalloy	55	20 000	75 000	8 500	55 000
1040	56	40 000	100 000	6 000	39 000
Megaperm	97	3 300	68 000	9 300	60 000
Hipernick	46	3 000	70 000	15 500	100 000
45 Permalloy	45	2 700	23 000	16 000	103 000
Rhometal*	95	250—2 000	1 200—8 500	12 000	78 000
4% Silicon Steel	55	450	8 000	19 500	125 000

ρ = resistivity in microhm cm ; μ_0 = initial permeability
μ_{max} = maximum permeability obtainable.

For audio transformer work the first four are frequently used. *RHOMETAL* has a special field of application, namely for transformers handling ultrasonic and radio frequencies up to several megacycles.

Flux densities of the order of 22 000 lines per square inch can be used with *MUMETAL* and approximately double this value with *RADIOMETAL*, the upper limit being set by the permissible distortion. For further information on this point see Sect. 3(iii).

For higher power **output transformers,** high silicon content (up to 4½%) sheet steel is in general use. To retain high permeability at low flux densities, the strips or laminations should be annealed after shearing and punching. Spiral cores of grain-oriented silicon steel are of considerable use in this application.

As a general rule, the output transformer should have the largest core which is practicable or permissible having regard to cost or other factors. A large core of ordinary silicon steel laminations is usually better than a small core of special low-loss steel.

The weight of steel in the core is a function of the minimum frequency, the permissible distortion, the core material, and the maximum power output. As a rough guide, subject to considerable variation in practice, the core may be taken as having

Weight in lbs. = 0.17 × watts output
Volume in cu. ins. = 0.7 × watts output.

These are for normal typical conditions, and may be decreased for less extended low frequency response or for a higher permissible distortion. For good fidelity, an increase in core size to double these values is desirable.

Several new **core materials** are now available including *CASLAM* and *FERROXCUBE.*

CASLAM is a soft magnetic core material with finely laminated structure for use at frequencies from 50 c/s up to at least 10 Kc/s. It is composed of flake iron particles pressed into a compact mass of the desired shape in such a way as to produce innumerable thin magnetic layers aligned in the plane of the flux. By virtue of its dense compacted structure many of the assembly and fixing problems associated with the older stacking method are eliminated. Grade 1 is a low density material with a maximum permeability of 860. Grade 2 is a denser material with a higher maximum permeability of 1000. Grade 3 is similar to grade 1 in magnetic characteristics but has better strength and machining qualities.

For choke cores a pair of E's can be butted together but where minimum gap is required the block can be broken and then rejoined after the coil is positioned. Because of the fibrous laminated structure exposed by breaking, microscopic interleaving occurs when the join is remade in the correct manner.

*Properties depend upon different heat treatment deliberately given.

Rigid clamping is less important since there are no free laminations to vibrate under load. For this reason also, combined with the somewhat discontinuous nature of the material, the acoustic noise emitted by the block is considerably reduced, especially at higher frequencies.

PHYSICAL PROPERTIES OF CASLAM

Property	Caslam 1	Caslam 2	Caslam 3
Maximum permeability	860 at 4 Kg	1000 at 4 Kg	
Effec. permeability	500 at 10 Kg	830 at 10 Kg	Similar
Hys. loss for B_{max} = 10 Kg. at 50~	7000	7000 ergs/cc/cycle	to
Coercive force B_{max} = 10 Kg. at 50~	2.4	2.4 oersteds	grade 1
Sat. flux density	18 Kg	18 Kg	
Total a.c. loss, B_{max} = 10 Kg. at 50~	2.5	2.75 watts/lb.	
Density	7.0	7.4 gms/cc	
Max. oper. temperature	110°C	110°C	150°C
Resistivity (ohm-centimetre)			
(a) Normal to plane of laminae	0.04		0.03
(b) In plane of laminae	0.003		0.002

The a.c. loss is almost entirely hysteresis, the eddy current loss being less than 10% of the total. For this reason, the a.c. permeability at 50~ is approximately equal to the d.c. figure and at higher acoustic frequencies, blocks of *CASLAM* compare favourably in magnetic properties with stacks built up from normal silicon iron sheet.

Dust cores generally suffer from low permeability and, to reduce eddy currents, particle size has to be reduced ; this causes further reduction in permeability.

Other new ferromagnetic materials such as the ferromagnetic spinels and *FERROXCUBE* are described in Chapter 11 Sect. 3(v)E, and find their applications principally at frequencies above the audio range.

(iii) Frequency response and distortion

(a) Interstage transformers—Class A and B

At the mid-frequency* the amplification is very nearly equal to the amplification factor of the valve multiplied by the turns ratio of the transformer, where the secondary is unloaded.

At low frequencies the gain falls off due to the decrease in primary reactance. The ratio of the amplification at a low frequency A_{lf} compared with that at the mid-frequency A_{mf} can be expressed thus—

$$\frac{A_{lf}}{A_{mf}} = \frac{1}{\sqrt{1 + \left(\frac{R}{\omega L_0}\right)^2}} \qquad (1)$$

where L_0 = primary inductance
and R = plate resistance plus primary resistance.

The response will fall off 3 db at a frequency such that $\omega L_0 = R$. At a frequency such that

$$\omega L_0 = 2R \qquad (2)$$

the response will be down approximately 1 db from the mid-frequency level.

At high frequencies the leakage reactance and shunt capacitance, in conjunction with the plate and winding resistances, form a low Q series resonant circuit. Above this resonant frequency the gain will fall off rapidly. In the neighbourhood of resonance the change in gain will depend on Q_0, the Q of the resonant circuit. This factor can be varied by adding external resistance or by winding the secondary partly with resistance wire (Ref. C3). The resonant frequency can be varied by changing the value of the total leakage inductance, L_s or the interwinding capacitances. These are both functions of the transformer structure. See also page 518.

By careful choice of core material, lamination dimensions and method of sectionalizing the winding, it is possible to achieve a frequency response extending beyond the normal audio range (Refs. C2, C4, C5, C6 and Figs. 5.12 and 5.13A).

*The mid-frequency is the frequency at which maximum gain is obtained.

As the leakage inductance L_s is proportional to the square of the turns, N, it is possible to extend the high frequency response by reducing N. This, of course, reduces L_0 in the same proportion as L_s, but this effect can be overcome by the use of a high permeability alloy.

For low level working it is usual to assume an initial a.c. permeability of 350 for silicon steel. If a *RADIOMETAL* core with an initial a.c. permeability of say, 1600, and of similar dimensions is substituted, N could be reduced in the ratio of $\sqrt{1600/350}$ or 2.14/1 and L_s approximately 4.6/1 over that for the silicon steel transformer. Similarly if a *MUMETAL* core having an initial a.c. permeability of 10 000 and of similar dimensions is substituted, N could be reduced in the ratio of $\sqrt{(10\,000/350)}$ or 5.34/1 and L_s approximately 28.5/1.

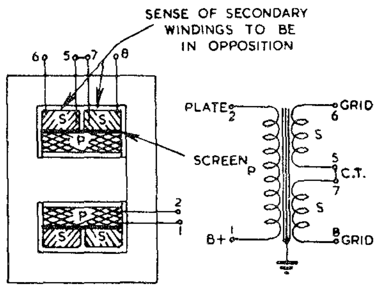


Fig. 5.12. *Transformer wound with balanced secondaries.*

In those examples, the primary inductance has remained constant, but the leakage inductance has decreased over 4 times and 28 times respectively without any sectionalizing of the winding or interleaving. This indicates the improvement possible with the use of high permeability alloys.

Care must be taken, with the reduction in turns, that the flux density in the core does not exceed safe limits from the point of view of distortion (see Refs. C7, C8, C9, C10).

In the case of transformers working at low levels and hence, generally, a low flux density, the distortion may not be a consideration. It is possible to achieve an improved high frequency response, while keeping the distortion at the same level at low frequencies, by increasing the core cross section and proportionately reducing the turns. This however, increases the mean length of turn and also L_s, so the nett reduction of L_s is not as great as would be anticipated. In addition, the cost of the transformer increases, particularly if a high permeability alloy is used. Class B input (driver) transformers are usually called upon to handle appreciable amounts of power during part of the audio frequency cycle, but are designed on the basis of open circuit working when considering primary inductance. Winding resistance and leakage inductance must be kept low to avoid distortion (Refs. C21, C22).

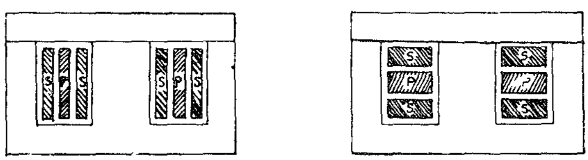


Fig. 5.13A. *Typical coil arrangements in audio transformers.*

(iii) (b) Low level transformers

Transformers working about or below a zero level of 1 milliwatt(0 dbm) usually employ a *MUMETAL* core. In most instances hum shielding is necessary to attain the desired signal to hum ratio. This is generally required to be in excess of 60 db. Satisfactory shielding can be obtained by using around the transformer one or more shields made of high permeability alloy. Where a.c. fields are strong, the outer shield is normally designed to be made of an alloy which has a high saturation flux density. Thus the inner shields of, say, *MUMETAL*, are working at maximum efficiency in a low a.c. field where their permeability is a maximum. Sometimes sheet steel or thick cast iron outer cases are used. More often a special alloy, such as Telcon 2129, is used. This material has magnetic properties similar to *RADIOMETAL*, but is suitable for deep drawing. Inner shields are often of copper to give shielding from electric fields (Refs. C11, C12). Some improvement is possible if a core type structure is used in place of a shell type, owing to the cancellation of stray voltages induced in the winding by external fields.

As an example of the use of high permeability alloys, the following design problem is presented. Calculate the primary turns for a 50 ohm to 50 000 ohm transformer working at zero level (0 dbm). The frequency response must not fall more than 1db below mid-frequency response, at 50 c/s Distortion must not exceed 1% at zero level at 50 c/s. The source impedance is 50 ohms resistive and the secondary is unloaded. Core material to be used is *MUMETAL*.

1st Step. Calculate primary inductance.

For 1 db attenuation, $\omega L_0 = 2R$ (Eqn. 2).

$L_0 = 2R/\omega = 2 \times 50/2 \times \pi \times 50 = 0.32$ henry.

2nd Step. Calculate primary turns.

Assume square stack of Magnetic and Elec. Alloys No. 35 lamination. Length of magnetic path, $l = 4.5$ inches

Cross sectional area = 0.56 square inches.

$$L_0 = \frac{3.2A\ \mu N^2}{10^8 \times l} \qquad \text{Therefore } N^2 = \frac{L_0 \times 10^8 \times l}{3.2 \times A \times \mu}$$

$$N^2 = \frac{0.32 \times 10^8 \times 4.5}{3.2 \times 0.56 \times 10^4} \qquad \text{Therefore } N = 90 \text{ turns.}$$

3rd Step. Calculate working flux density.

Primary voltage $E = \sqrt{W.R}$

where W = input power

Therefore $E = \sqrt{1 \times 10^{-3} \times 50} = 0.224$ V

and $$B = \frac{0.224 \times 10^8}{4.44 \times 50 \times 90 \times 0.56}$$

= 2000 lines per square inch

= 310 Gauss.

4th Step. Determine percentage distortion.

Referring to Fig. 5.13B and using curve for $\omega L_0 = 2R$, it will be noted that the percentage distortion is approximately 0.75%. This assumes that the permeability value, μ, is still 10 000 at the operating flux density. In practice μ may exceed this figure and thus the distortion as calculated above may be larger than would be measured in a finished transformer. In this problem, for simplification, no account has been taken of the stacking factor, which would modify the result slightly.

When the secondary of the transformer is loaded, R in eqn. 1 then becomes R_A as in eqn. 3 and R in Fig. 5.13B is read as R_A. The calculation for distortion then follows in a similar manner to the unloaded secondary example worked earlier.

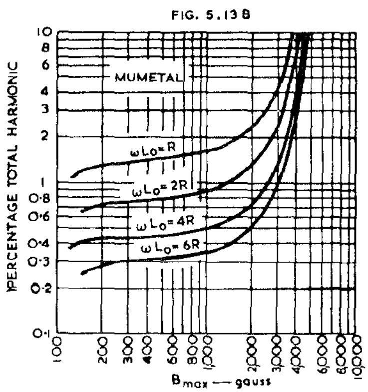


Fig. 5.13B. Total harmonic distortion plotted against flux density for different ratios of primary reactance to R where R = plate resistance + primary resistance (Ref. C9).

(iii) (c) Output transformers

The factors affecting the frequency response of output transformers (Ref. C13) are similar to those affecting interstage transformers. Refer to Figs. 5.10B, C and D.

The response falls off from the mid-frequency gain by 3 db at a low frequency such that

$$\omega L_0 = R_A \tag{3}$$

Attenuation is 1 db when

$$\omega L_0 = 2R_A \tag{4}$$

where $R_A = \dfrac{(r_p + R_W)R_L}{r_p + R_W + R_L}$

r_p = plate resistance of valve

R_L = load resistance referred to the primary

and R_W = total winding resistance referred to the primary.

The response falls off from the mid-frequency gain by 3 db at a high frequency such that

$$\omega L_S = R_B \tag{5}$$

Attenuation is 1 db when

$$\omega L_S = 0.5\, R_B \tag{6}$$

where $R_B = r_p + R_W + R_L$

and L_S = total leakage inductance referred to primary.

$$\text{The gain at the mid-frequency} = \frac{E_0}{E_i} = \frac{\mu R_L}{R_B} \tag{7}$$

where E_i = voltage input to grid of output valve

E_0 = voltage output across R_L

and μ = amplification factor of output valve.

It is thus possible to specify the primary and leakage inductances permissible when the frequency response requirements are known. An example will illustrate this.

Determine the minimum primary inductance and maximum leakage inductance permitted in an output transformer designed to match a pair of Class A 2A3 triode valves with a 5 000 ohm load. The response is to be within 1 db from 50 to 10 000 c/s. The plate resistance of each valve is 800 ohms. Neglect R_W.

Make all calculations from plate to plate.

For a fall of 1 db at 50 cycles per second,

$$\omega L_0 = 2R_A.$$

$$\text{Now } R_A = \frac{5000 \times 1600}{5000 + 1600} = 1200 \text{ ohms approx.}$$

$$\omega L_0 = 1200 \times 2 = 2400$$

$$L_0 = 2400/2 \times \pi \times 50$$

$$= 7.6 \text{ henrys approx.}$$

This is the value that would be measured on a bridge at low induction.

For a fall of 1 db at 10 000 cycles per second,

$$\omega L_S = 0.5\, R_B$$

Now $R_B = 5000 + 1600 = 6600$ ohms.

$$\omega L_S = 0.5 \times 6600 = 3300$$

$$L_S = 3300/2 \times \pi \times 10\,000$$

$$= .052 \text{ henry approx.} = 52 \text{ millihenrys.}$$

Note particularly that distortion requirements may necessitate an increase in L_0 and a decrease in L_S.

The following table indicates the relationship between the low frequency attenuation and the ratio of $\omega L_0/R_A$:

TABLE 1

Loss	Relative amplification	$\omega L_0/R_A$
0.5 db	0.94	2.76
1.0 db	0.89	1.94
2.0 db	0.79	1.30
3.0 db	0.71	1.00
6.0 db	0.50	0.58

The inductances required for various values of R_A, for a bass response loss of 1 decibel are as follows, correct to two significant figures :

TABLE 2

Value of R_A ohms	Bass response down 1 db at 150 c/s	100 c/s	50 c/s	30 c/s
800	1.7 H	2.6 H	5.1 H	8.5 H
1200	2.6 H	3.8 H	7.6 H	13 H
1500	3.2 H	4.8 H	9.5 H	16 H
2000	4.3 H	6.4 H	13 H	21 H
3000	6.4 H	9.5 H	19 H	32 H
4000	8.5 H	13 H	26 H	42 H
5000	11 H	16 H	32 H	53 H
7500	16 H	24 H	48 H	80 H
10 000	21 H	32 H	64 H	110 H
15 000	32 H	48 H	95 H	160 H
20 000	43 H	64 H	130 H	210 H
50 000	110 H	160 H	320 H	530 H

where R_A is approximately equal to the load resistance R_L in parallel with the effective plate resistance of the valve ; however see comments below. When the plate resistance is very high, as for a pentode without feedback, R_A may be taken as being approximately equal to R_L.

For loss in bass response to be reduced to 0.5 decibel these inductance values should be increased by a factor of 1.9 times. For a reduction of 2 db the factor becomes 0.67.

Since the permeability of the core material varies with induction, the frequency response will vary with signal level—the limiting low frequency will usually extend lower as the signal level is increased. It is therefore desirable that the inductance values tabulated above be calculated or measured at low signal levels. In the case of feedback from the secondary, this effect should be taken into account.

Table 2 gives the value of inductance to provide a nearly constant output voltage, but this is only one of several requirements to be satisfied. The **core distortion** is a function of the ratio $\omega L_o/R_A$ (this may be derived from equations 9 and 10) and for low distortion a high ratio of inductive reactance to R_A is required, and this is equivalent to having a low value of bass attenuation (see Table 1). The bass attenuation in Table 2 (1 db at specified minimum frequencies) is based on $\omega L_o/R_A = 1.94$, resulting in fairly low core distortion. Still lower core distortion would be achieved by increasing the inductance and hence, incidentally, decreasing the bass attenuation.

The other important secondary effect resulting from a finite value of inductance is the **phase angle of the load** presented to the output valve(s). If the value of inductance from Table 2 is used for bass response down 1 db at a specified frequency, the phase angle of the load presented to the output valves will be between 45° and 90° for a triode either without feedback or with negative voltage feedback. This will cause a pronounced elliptical load-line, normally resulting in severe valve distortion at this frequency, at full power output.

If R_A in Table 2 is taken as being equal to R_L, the maximum phase angle of the load will be less than 28° for bass response 1 db down at the specified frequencies. If the factor 0.67, for bass response down 2 db, is applied to Table 2, the maximum phase angle of the load will be less than 38°. If the factor 2.0, for bass response down about 0.5 db, is applied to Table 2, the maximum phase angle of the load will be about 15°. This appears to be a reasonable value for good fidelity.

Table 2 may therefore be used as a general guide to the choice of inductance values where more exact calculations are not required—see below.

Summary of general application of Table 2

To give low core distortion, nearly constant output voltage and a total load impedance effective on the valve which is not too reactive :—

1. Take $R_A = R_L$.
2. Apply Table 2 as printed for ordinary use.
3. Multiply inductance values by a factor of 2 for good fidelity.
4. The specified frequencies in Table 2 are to be interpreted as the minimum frequencies of operation for the transformer.

Where the source impedance is high, the high frequency fall off is determined largely by L_S and the winding capacitances as in the case of interstage transformers.

Similar devices to that employed in the construction of interstage transformers can be used to extend the range of output transformers. The use of *RADIOMETAL*, specially annealed high silicon content steel (such as *SUPERSILCOR*), and grain oriented steels, are all common in better quality output transformers (Refs. C14, 15).

It is feasible to " build-out " a transformer into a half section filter and thus maintain the impedance, viewed either by the source or the load, constant over a wide frequency range. In addition the phase angle variation is reduced towards the extremes of the range. This reduces distortion and maintains full power output to a greater degree than otherwise possible.

This idea can be applied to interstage transformers and output transformers quite successfully (Refs. C19, C20). Even large modulation transformers and class B driver transformers for broadcast equipment have provided improved performance when treated in this way (Refs. C16, 17, 18).

In radio receivers and record players advantage can be taken of this " building-out " procedure, to limit the high frequency response to any given point, say 6000 c/s, with rapid attenuation thereafter. This usually involves only one extra component ; a condenser across the secondary winding of the output transformer. This is quite an effective " top " limit, more so than the normal tone control. The output transformer is designed to have the necessary amount of leakage inductance for the network to function as intended.

The **winding resistances** are not of major importance in interstage transformers, but assume greater significance in output transformers. An appreciable amount of power may be lost unless the resistances of both primary and secondary are kept to reasonable proportions. In the normal good quality transformer the total resistance reflected into the primary side is approximately double the measured d.c. resistance of the primary. The total winding resistance (referred to the primary) will vary between 10 and 20 per cent of the load resistance, which means an insertion copper loss of 0.5 to 1 db. This extra resistance must be considered when choosing the turns ratio to reflect the correct load (see Eqn. 2, Sect. 2), otherwise an impedance error of 10 to 20 percent will occur. Core losses will not materially affect the calculation as these losses do not reach their maximum except at full power at the lowest audio frequency of interest.

Distortion in output transformers

When a transformer has its primary connected to an audio frequency source of zero impedance, the waveform of the voltage on the secondary will be the same as that of the source—in other words there is no distortion.

When a transformer is connected in the plate circuit of a valve, the latter is equivalent to a resistance r_p in series with the source and the transformer primary. If the secondary of the transformer is also loaded by a resistance R_2, this is equivalent (as regards its effect on distortion) to a total primary series resistance R,

where $\dfrac{1}{R} = \dfrac{1}{r_p} + \dfrac{1}{R_2(N_1/N_2)^2}$

N_1 = primary turns

and N_2 = secondary turns.

In the following treatment the symbol R is used to indicate the total effective primary series resistance, whether it is caused by r_p alone or by a combination of this and secondary loading.

The resistance R in series with the primary causes a voltage drop proportional to the current flowing through it, which is the magnetizing current. Now the form of the magnetizing current is far from being sine-wave, since it is distorted by the non-linear B-H characteristic of the core material.

This distorted current waveform has no bad effect when R is zero, but results in distortion of the voltage waveform which becomes progressively greater as R is increased, for any one fixed value of B_{max}.

The resulting harmonic distortion with silicon steel has been calculated by Dr. N. Partridge, and his results are embodied in the formula which follows (Refs. C24, C25, C26, C27).

$$\frac{V_h}{V_f} = S_H \cdot \frac{10^9}{8\pi^2} \cdot \frac{l}{N^2 A} \cdot \frac{R_A}{f}\left(1 - \frac{R_A}{4Z_f}\right) \qquad (8)$$

This formula can be modified to include the core stacking factor, 90%, and to use inch units instead of centimetres. It then becomes

$$\frac{V_h}{V_f} = \frac{5.54 S_H l R_A}{N^2 A f}\left(1 - \frac{R_A}{4Z_f}\right) \qquad (9)$$

where V_h = the harmonic voltage appearing across the primary,
V_f = the fundamental voltage across the primary,
S_H = the distortion coefficient of the magnetic material,
l = the length of the magnetic path,
N = number of primary turns,
R_A = resistance (or equivalent resistance) in series with the primary, (refer under eqn. 4, Sect. 3),
A = cross-sectional area of core,
f = frequency of fundamental in cycles per second,
Z_f = impedance of primary at fundamental frequency $\approx 2\pi f L$
and L = inductance of primary in henrys at chosen flux density.

In most cases the final term $(1 - R_A/4Z_f)$ can be omitted with a further simplification.

The right hand side of this equation gives the value of the fractional harmonic distortion ; the percentage harmonic distortion may be obtained by multiplying this value by 100. The formula holds only for values of R/Z_f between 0 and 1 ; this limits its application to output circuits having a maximum attenuation of 3 db.

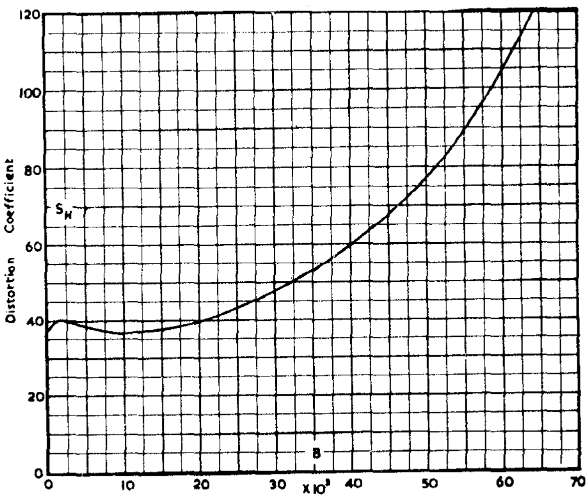


Fig. 5.13C. The distortion coefficient of Silcor 2 as a function of B (Ref. C25).

Values of S_H for Silcor 2 can be obtained from Fig. 5.13C. Similarly Fig. 5.13D can be used to determine the permeability of Silcor 2 and hence the inductance of the primary from the formula

$$L = \frac{2.88 N^2 A \mu}{10^8 l} \text{henrys} \qquad (10)$$

where μ = permeability at operating flux density and core stacking factor is 90%.

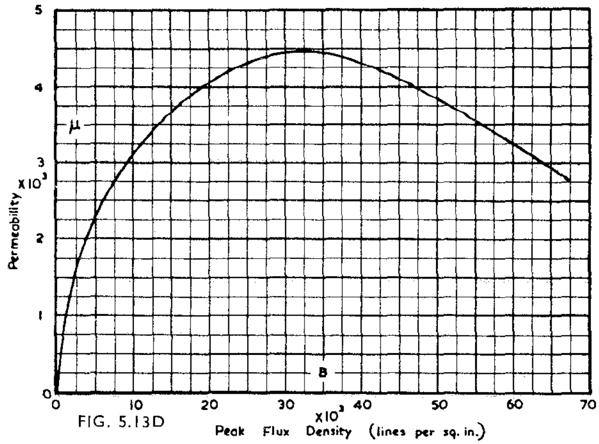


Fig. 5.13D. *Variation of permeability with B at 50 c/s of Silcor 2 laminations,* 0.014 *in. thick* (Ref. C25).

Both of these figures have been adapted from those published by Partridge (Ref. C25).

An example is quoted to demonstrate the use of these formulae.

Determine the transformer distortion produced when a pair of KT66 valves (very similar to type 6L6-G) connected as triodes are operated in conjunction with a transformer having the following characteristics.

Lowest frequency of operation	50 c/s.
Maximum flux density	40 000 lines/sq. in.
Core stack of M.E.A. 78 Pattern lams. Silcor 2	2 ins.
(Refer to Sect. 5 and also Fig. 5.18C for lamination data).	
Operating conditions of KT66 valves	
Plate voltage	400 volts
Plate resistance (per valve)	1450 ohms
Load resistance (total)	10 000 ohms
Power output	14 watts

1st step. Calculate primary voltage

$$E = \sqrt{WR} = \sqrt{14 \times 10\,000} = 374 \text{ V.}$$

2nd step. Calculate primary turns

$$N = \frac{E \times 10^8}{4 \times fBA} \quad \text{(allowing 90\% core stacking factor)}$$

$$= \frac{374 \times 10^8}{4 \times 50 \times 4 \times 10^4 \times 1.25 \times 2} = 1870 \text{ turns.}$$

3rd step. Determine μ from Fig. 5.13D

$\mu = 4300.$

4th step. Calculate Z_f ($\approx 2\pi fL$)

$$Z_f = \frac{2\pi f\, 2.88\, N^2 A\mu}{10^8 l}$$

$$= \frac{2 \times \pi \times 50 \times 2.88 \times 1.87 \times 1.87 \times 10^6 \times 1.25 \times 2 \times 4300}{10^8 \times 7.5}$$

$= 45\,300$ ohms.

5th step. Calculate R_A (parallel resistance of plate and load resistance).

$$R_A = \frac{(r_p + R_W)\, R_L}{r_p + R_W + R_L} \quad \text{(see eqn. 4, Sect. 3)}$$

assuming $R_W = 400$ ohms

then $R_A = \frac{(2900 + 400)\ 10\,000}{13\,300} = 2480$ ohms.

6th step. Calculate $1 - (R_A/4Z_f)$

$1 - (R_A/4Z_f) = 1 - (2480/4 \times 45\,300) = 0.986.$

Thus this factor can be neglected without serious error.

7th step. Determine S_H from Fig. 5.13C

$S_H = 60.$

8th step. Calculate fractional distortion

$$\frac{V_h}{V_f} = \frac{5.54\ S_H l R_A}{N^2 A f} = \frac{5.54 \times 60 \times 7.5 \times 2480}{1.87 \times 1.87 \times 10^6 \times 2.5 \times 50}$$

$= .014.$

Thus percentage distortion is $.014 \times 100$ or 1.4%.

Note : High fidelity output transformers may be designed with distortion less than 0.05% at maximum power output at 50 c/s.

The value of the distortion coefficient S_H is constant for any given material operating at any one value of B_{max}. The value of S_H will be different for each harmonic, but its value for the third harmonic (S_3) is very close to the r.m.s. sum of all harmonics (S_{rms}) when there is no direct current component.

It will be noticed that the curve in Fig.5.13Creaches a minimum at about $B_{max} =$ 10 000 lines/sq. in., and that the distortion coefficient rises at both lower and higher values of B_{max}, although the lower rise is only slight. Actually the lower part of the curve (below the knee) drops away rapidly and eventually reaches zero at $B_{max} = 0$. Some alloy core materials have appreciable values of the distortion coefficient even when approaching $B_{max} = 0$ (Ref. C9).

When there is a direct polarizing field, even as well as odd harmonic distortion are both evident. Under these conditions, below 65 000 lines/sq. in., the r.m.s. sum of the second and third harmonic currents S_{2+3} approximates to the r.m.s. sum of all the harmonic currents.

It will be seen from this analysis that it is desirable to use a low impedance source, (e.g., triodes), and a high inductance primary for best results at low frequencies.

Distortion at high frequencies

At high frequencies distortion is produced, apart from normal valve distortion due to non-linearity of characteristics, by the leakage inductance and winding capacitances which change the magnitude and phase angle of the load impedance.

The load on the secondary, if a loudspeaker, also complicates this trouble. Here again, a low impedance source is desirable ; a high impedance source will accentuate the distortion due to this effect. With Class B amplifiers, it is essential that the leakage inductance between each half of the primary be as small as possible, otherwise there will not be proper cancellation of even order harmonics, and higher order harmonics will be generated (Ref. C23). A static shield between primary and secondary will prevent any stray even harmonics being fed into the secondary by capacitive coupling. This shield will alter the winding capacitances and increase the leakage inductance, hence it must be employed judiciously.

In Class B output transformers, high leakage inductance and winding capacitances cause distortion and decrease in power output. To a considerable degree, these remarks on distortion at high frequencies also apply to Class AB transformers.

(iv) Designing for low leakage inductance*

Assuming that the turns and winding dimensions are kept constant, leakage inductance can be progressively reduced by interleaving the winding structure until the limit is reached when

$$a/3N^2 < c \tag{11}$$

where a = total thickness of all winding sections

N = number of leakage flux areas

and c = thickness of each insulation section.

*This treatment follows Crowhurst, N. H. (Ref. C28).

From eqn. (11) it will be seen that the insulation between the sections is the limiting factor. Fig. 5.13E shows that the largest value of N^2 for a given number of sections is achieved when there is a half section at the end of each winding structure. Although series connections are shown, similar results can be obtained by paralleling the sections. In this case all the turns in each paralleled section must be equal. An example of the use of Fig. 5.13H to determine leakage inductance follows :

A push-pull output transformer is to be wound on a former 1.875 inches long and have a winding height of 0.4 inch, allowing for clearance. The mean length of turn is 9.5 inches. The primary and secondary turns are 2700 and 120 respectively. Insulation between sections is 0.015 inch. Assuming the seventh winding arrangement in left hand column of Fig. 5.13E, it is required to determine the leakage inductance between the whole primary and secondary. The total primary winding height is 0.24 inch and the total secondary winding height is 0.16 inch.

$a = 0.4$ inch. $\qquad N^2 = 16.$

Referring to Fig. 5.13H, intercepts of a and N^2 give 0.008.

Adding $c = 0.015$ gives 0.023.

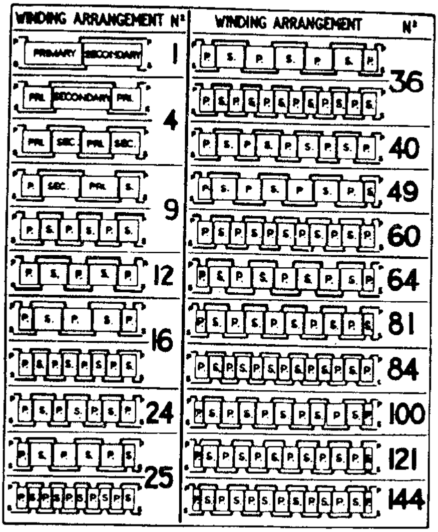


Fig. 5.13E. *Table of winding arrangements (Ref. C28).*

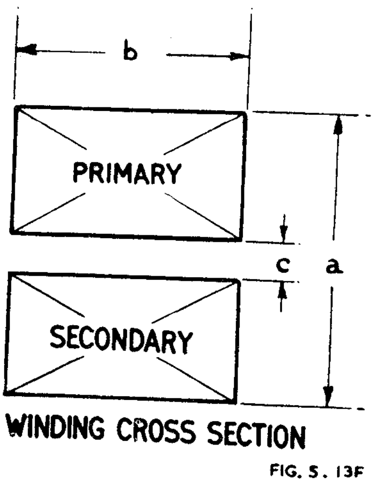


Fig. 5.13F. *Dimensions used in chart (Fig.* 5.13H).

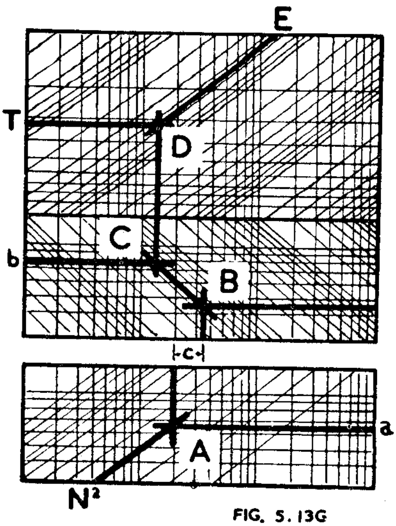


Fig. 5.13G. *Showing use of chart (Fig.* 5.13H).

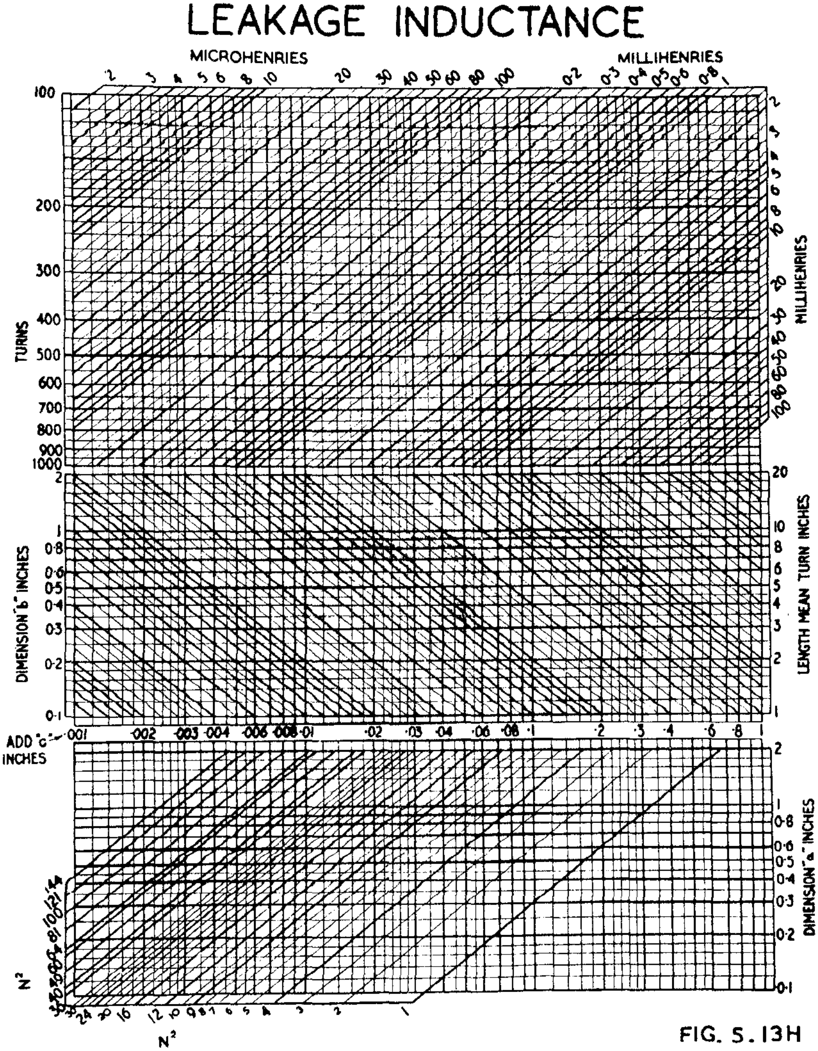


Fig. 5.13H. *Leakage inductance chart* (*Ref.* C28).

Intercepts with the mean length of turn, 9.5 inches and winding width, *b*, of 1.875 inches gives a vertical of 0.012.

Assuming 270 turns, leakage inductance = 0.14 millihenry, or 14 millihenrys for 2700 turns.

(v) Winding capacitance*

This information assists in computing the various winding capacitances of multi-layer windings used in the construction of audio-frequency transformers, chokes and other equipment. They apply to windings in which turns are wound on layer by layer, either interleaved or random wound, so that the P.D. between adjacent turns belonging to consecutive layers will be much greater than that between adjacent turns

*Reprint of an article by Crowhurst, N. H., in Electronic Eng. (Ref. C29).

in the same layer. The capacitance effect between adjacent turns in the same layer is neglected, only that between layers being considered. Capacitance between winding and core, and electrostatic screens, if used, must also receive attention.

Fig. 5.13J illustrates a cross section of a piece of winding in which layer interleaving is used. It is seen that the dielectric between adjacent conductors in consecutive layers is complex both in shape and material. The turns on the top layer shown fall so that each turn drops in the space between two turns on the second layer, corrugating the interleaving material with a slight resulting increase in capacitance compared with that between the middle and bottom layers shown. Due to the spiral form of each layer, the position of turns in consecutive layers to one another will change at different points round the direction of winding, thus the capacitance between any pair of layers will automatically take up the average value. The composite dielectric is made up of conductor insulation, most commonly enamel, interleaving material and the triangular shaped spaces left between adjacent turns and the interleaving material. These spaces will be filled with dry air if the windings are dried out and hermetically sealed, or with impregnating compound if the windings are vacuum impregnated. The latter procedure will give rise to a somewhat higher capacitance.

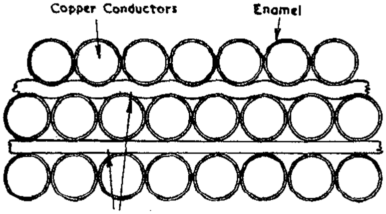


Fig. 5.13J. *Section through layer interleaved winding (Ref.* C29).

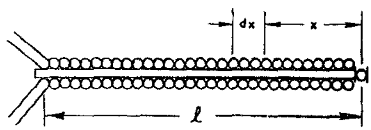


Fig. 5.13K. *Effective capacitance of a pair of layers (Ref.* C29).

In practice the major controlling factor determining the total capacitance between two adjacent layers of winding is the thickness of the interleaving material, thus distributed capacitance may conveniently be estimated in terms of the thickness of interleaving material, giving this material a value of dielectric constant empirically obtained, allowing for the average effect of the other dielectrics in the composite arrangement.

Effective layer to layer capacitance

Take a winding having only two adjacent layers, its turns distributed uniformly throughout the two layers as shown in section at Fig. 5.13K. Consider an element dx at a distance x from the end of the winding where the conductor steps up from one layer to the other. In the complete winding the elemental capacitance due to the section dx will be transformed so that it can be represented as an equivalent value across the whole winding. If the length of the whole layer is l and the capacitance per unit layer length C, then the effective capacitance of the element referred to the whole winding will be $(x/l)^2C.dx$. The capacitance due to the whole winding will be

$$\int_0^1 (x/l)^2 C.dx = \frac{1}{3}\, l.C.$$

Thus the effective capacitance of such a two-layer winding is one-third of the capacitance between two layers measured when their far ends are unconnected.

Take now a winding consisting of n whole layers : there will be $(n - 1)$ adjacent pairs of layers throughout the winding and the effective capacitance of each pair of layers, referred to the whole winding, will be a capacitance of $(2/n)^2$ times their capacitance referred to the high potential end and considered as a pair. Thus the capacitance of n whole layers referred to the whole winding becomes

$$\frac{4}{3} \cdot \frac{(n-1)}{n^2} \cdot Cl$$

For large values of n the capacitance becomes inversely proportional to the number of layers. Fig. 5.13Q illustrates a typical winding shape together with the dimensions as used in the related diagrams. The capacitance per layer, given in the foregoing formula as Cl, is proportional to the product of the length per mean turn L_{mt}, and the length of layer L_w.

Vertical sectionalizing

It is sometimes of advantage to sectionalize the winding as shown at Fig. 5.13L, each vertical space being filled completely before proceeding to fill the next one.

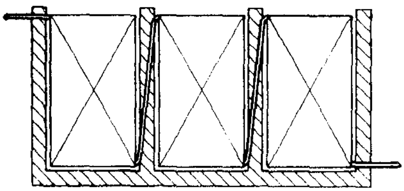


FIG. 5.13L

Fig. 5.13L: Sectionalizing to reduce capacitance (Ref. C29).

Although physically the winding will have the same overall cross-sectional dimensions its self-capacitance will be equivalent to that of a winding having $1/N \times$ layer length and $N \times n$ layers, where N is the number of vertical sections. The distributed capacitance of the winding due to such sectionalizing thus is reduced by the factor $1/N^2$. Note that this rule applies only to referred interlayer capacitance and does not apply to capacitance between the top and bottom of the winding and adjacent windings or screens. The various reduction factors for vertical sectionalizing are given in Table 4.

TABLE 4

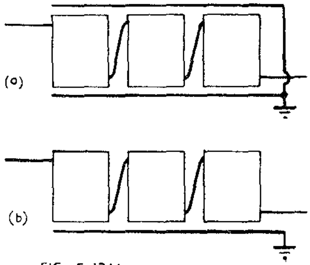
Number of vertical sections	Distributed capacitance component	Winding to screen capacitance arrangement as Figure 5.13M		
		(a) One side earthy	(a) Centre point earthy	(b) One side earthy
1	1	1	0.5	0
2	0.25	0.75	0.25	0.125
3	0.111	0.704	0.185	0.185
4	0.0625	0.6875	0.1875	0.219
5	0.04	0.68	0.168	0.24
6	0.0278	0.676	0.176	0.255
∞	0*	0.667	0.167	0.333

*The method of winding is here changed so that for the purposes of this column, the dimensions L_w and T_w will change places.

Effect of mixing windings

In the design of a transformer it is often necessary to mix the primary and secondary windings in order to reduce leakage inductance. This arrangement will generally be a disadvantage as regards minimizing winding capacitance, since it exposes greater surface area of winding in proximity to either the other winding or an earthed screen. If the ratio of the transformer is fairly high, then from the high impedance winding the whole of the low impedance winding appears at common audio potential, usually earthy. But if the ratio of the transformer is not very high, capacitance between points at differing audio potentials in the two windings may have serious effects, and it is generally best to arrange the windings so as to avoid such capacitance.

Fig. 5.13N shows a cross-section suitable for an inter-valve transformer designed to operate two valves in push-pull from a single valve on the primary side. The H.T. end of the primary is earthy and is therefore diagrammatically earthed. The high potential end of the primary is adjacent to the earthy end of one of the half secondaries so that the capacitance between windings at this point is effectively from anode to earth. The two high potential ends of the secondary are remote from the primary and so minimize the possibility of unbalanced capacitance transfer from primary to one half secondary.



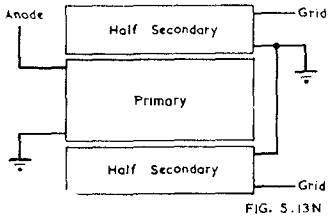


Fig. 5.13M. *Arrangements using vertical sectionalizing. See Table* 1 (*Ref. C*29).
Fig. 5.13N. *Push-pull secondary intervalve transformer secondary arrangements* (*Ref. C*29).

Another problem which often arises is in the design of push-pull output transformers, particularly for Class AB or Class B circuits, where it is essential that each half of the primary be well coupled to the whole secondary. From the viewpoint of leakage inductance and winding resistance, it is unimportant whether the secondary sections are connected in series or parallel. Fig. 5.13P illustrates three arrangements for a transformer of this type, each of which may be best suited under different circumstances. At (a) is an arrangement which gives minimum primary capacitance, but suffers from the defect that leakage inductance and winding resistance are unequal for the two primary halves. For Class A operation using valves requiring an optimum load of high impedance this arrangement is sometimes the best. At (b) is an arrangement intended to equalize winding resistance and leakage inductance from each half-primary to the whole secondary as well as primary self-capacitance. This arrangement is particularly suited to circuits employing low loading Class AB or Class B operation. The alternative arrangement shown at (c) results in a slightly lower referred capacitance across one half only of the primary. In general this unbalance is not desirable, but if leakage inductance is adequately low, the coupling between all the windings may be so good that the reduction in capacitance may be apparent across the whole primary.

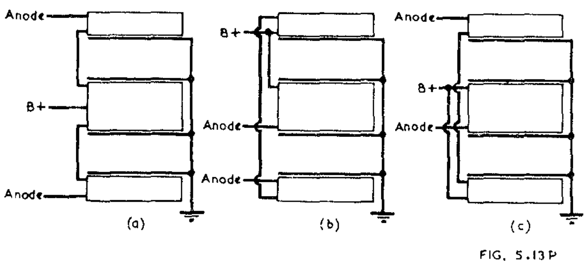


Fig. 5.13P. *Arrangements for push-pull output transformer* (*Ref. C*29).

TABLE 5.

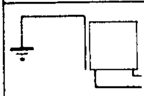
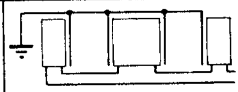
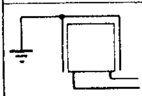
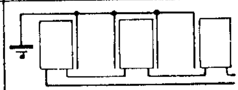
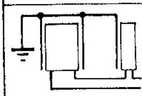
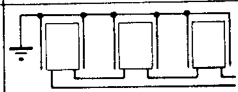
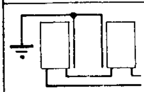
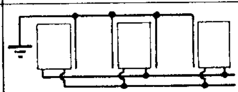
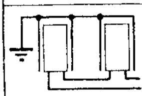
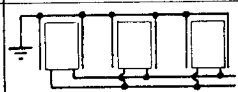
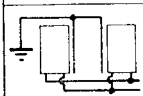
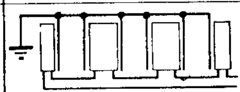
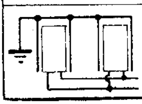
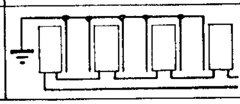
WINDING & SCREEN ARRANGEMENT & CONNECTION	CAPACITANCE FACTOR		WINDING & SCREEN ARRANGEMENT & CONNECTION	CAPACITANCE FACTOR	
	One side earthy	Centre point earthy		One side earthy	Centre point earthy
	—	·25		1·25	·25
	1	·5		1·11	·11
	·22	·055		2·11	·611
	·5	—		* 1	1
	1·5	·5		3	1·5
	* —	·5		1·94	·44
	2	1		1·75	·25

Table 5 gives a pictorial representation of various ways in which high impedance windings may be arranged in relation to earthy points shown as screens. The table is equally applicable if these points are earthy low impedance windings. The capacitance factors for alternative connexions of the windings, with either one side or the centre point at earth potential, are given relative to the average capacitance between one end layer of the winding and one screen. The two arrangements marked with an asterisk indicate that it is necessary to reverse the direction of winding in order tc achieve the capacitance factor shown.

Random winding

In what is known as random winding, no interleaving material is used. For ideal

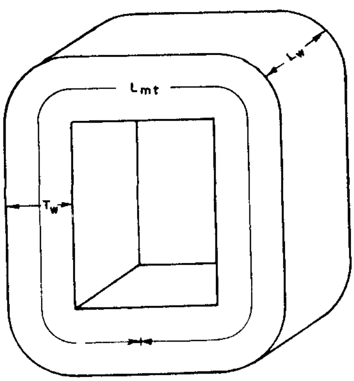


FIG. 5.13Q

Fig. 5.13Q. *Dimensions used in figures* 5.13R, 5.13S, 5.13T, 5.13U (*Ref.* C29).

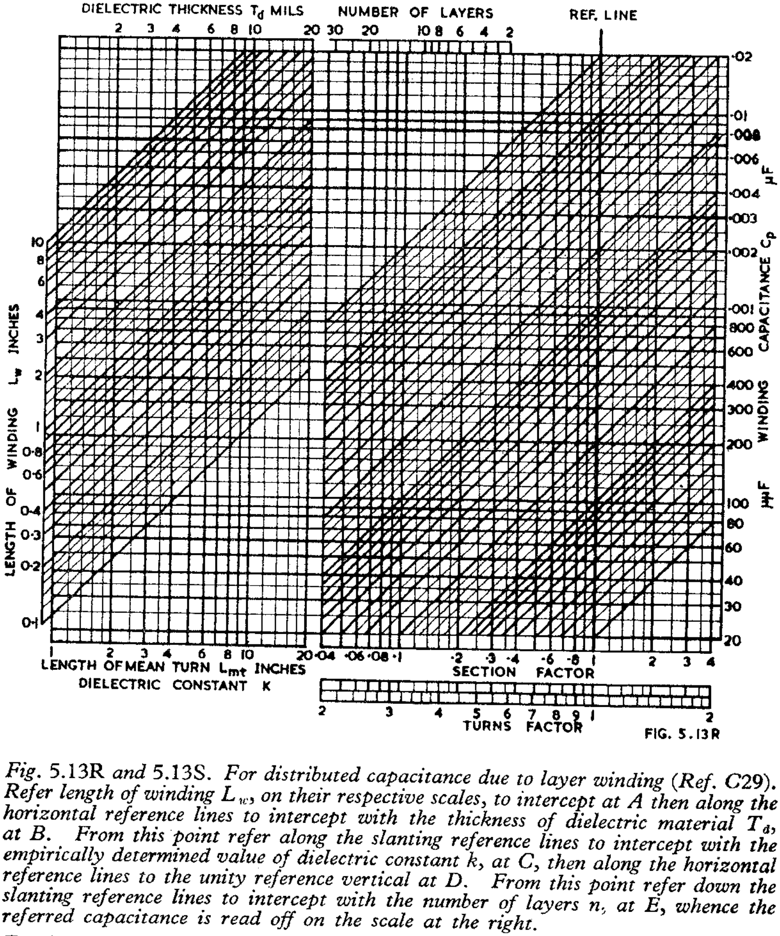


FIG. 5.13R

Fig. 5.13R *and* 5.13S. *For distributed capacitance due to layer winding (Ref. C29). Refer length of winding L_w, on their respective scales, to intercept at A then along the horizontal reference lines to intercept with the thickness of dielectric material T_d, at B. From this point refer along the slanting reference lines to intercept with the empirically determined value of dielectric constant k, at C, then along the horizontal reference lines to the unity reference vertical at D. From this point refer down the slanting reference lines to intercept with the number of layers n, at E, whence the referred capacitance is read off on the scale at the right.*

For interwinding, or winding to screen capacitance: As above, for the winding dimensions and dielectric thickness to a point corresponding to D, whence, for winding to screen capacitance, or interwinding capacitance when the other winding is all at low potential, refer down the slanting reference lines to intercept with the section capacitance factor obtained from Table 4. For interwinding capacitance where there is appreciable potential in the adjacent portions of both windings, individual attention will be necessary for each interwinding space, and the turns factor scale will assist here.

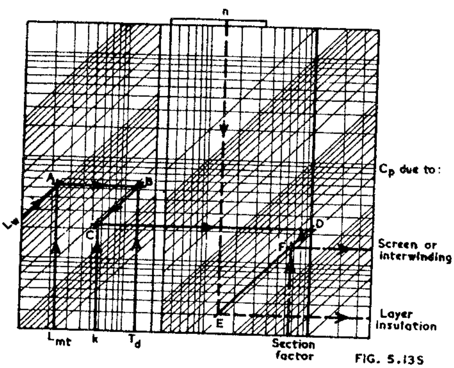


FIG. 5.13S

random winding the layers should be built up so that at all times during winding the top surface is level. Failure to do this will not greatly affect the self-capacitance, but will result in increased danger of breakdown due to electrical or mechanical stresses. As the number of layers is always large, it is convenient to reduce the calculation to simple terms of the winding dimensions.

The essential variables are: length of winding L_w, length of mean turn L_{mt}, winding thickness T_w (see Fig. 5.13Q), and number of turns, T. Considering variation of each of these quantities in turn, the others being taken as constant: variation of L_w will vary the referred capacitance per layer as before, and additionally the effective number of layers will vary inversely as $L_w^{\frac{1}{2}}$; variation of L_{mt} simply varies the

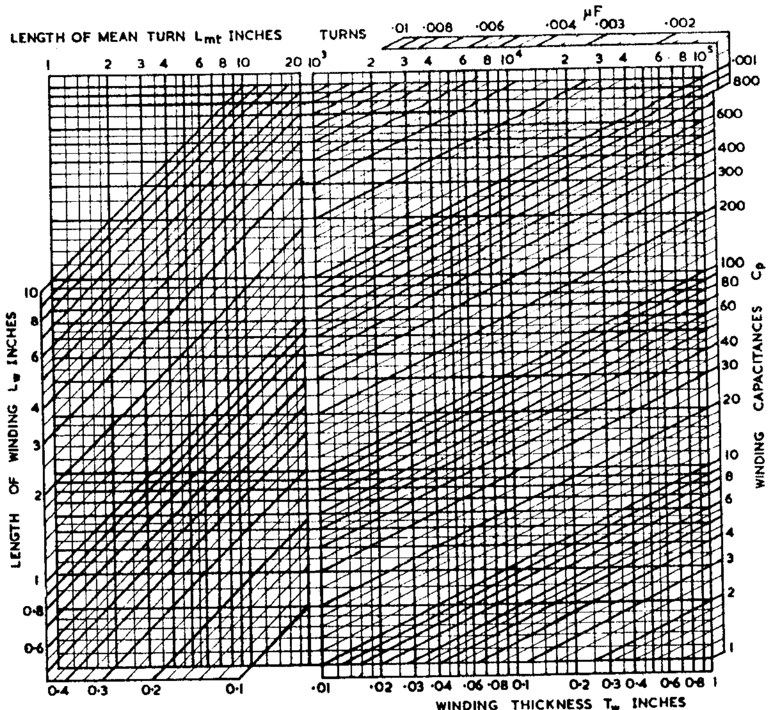


FIG. 5.13T

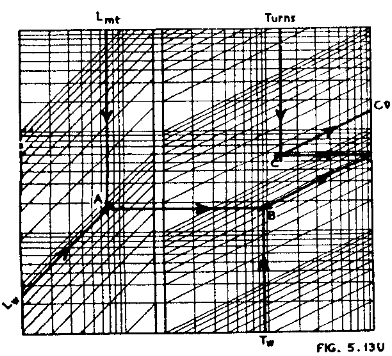


FIG. 5.13U

Fig. 5.13T and 5.13U (Ref. C29). For distributed capacitance due to random winding. Refer length of winding L_w, and length of mean turn L_{mt}, on their respective scales, to intercept at A then along the horizontal reference lines to intercept with the vertical line for thickness of winding T_w, at B. From here refer up the slanting reference lines to the right-hand edge of the data sheet, and then back along the horizontal lines to intercept with the vertical lines for the number of turns at C from which referred capacitance is read off along the slanting lines at C_p.

layer size, as before ; variation of both T_w and T varies the number of layers in direct proportion to $T_w^{\frac{1}{2}}$ or T. Thus the whole expression for variation of capacitance can be written,

$$C_p \propto \frac{L_{mt} L_w^{3/2}}{T_w^{\frac{1}{2}} T^{\frac{1}{2}}}$$

Figures 5.13T and 5.13U are based on this relation and empirical values obtained from average results with random windings.

Example 1

A push-pull output transformer is arranged as at Fig. 5.13N : primary winding has a total of 12 layers, $T_d = 3$, $k = 2$: insulation between primary and screen, $T_d = 15$, $k = 5$; main dimensions, $L_w = 2$ in., $L_{mt} = 8$ in.

Distributed capacitance, using Figs. 5.13R and 5.13S, $L_w = 2$ in., $L_{mt} = 8$ in., $T_d = 3$, $k=2$ and $n = 12$, is $C_p = 240$ $\mu\mu$F.

Capacitance from each primary to screen, substituting $T_d = 15$, $k = 5$, is $C_p =$ 1200 $\mu\mu$F. The capacitance factors for the arrangement of Fig. 5.13N are : (a) 0.25 ; (b) 0.5 ; (c) 0.375. Thus the total capacitance referred to the whole primary for each method of connexion is,

(a) 300 $\mu\mu$F + 240 $\mu\mu$F = 540 $\mu\mu$F.
(b) 600 $\mu\mu$F + 240 $\mu\mu$F = 840 $\mu\mu$F.
(c) 450 $\mu\mu$F + 240 $\mu\mu$F = 690 $\mu\mu$F.

Example 2

An intervalve transformer, to operate from push-pull to push-pull, uses a simple arrangement having both windings all in one section : $L_w = 0.6$ in, $L_{mt} = 2.5$ in, T_w (each winding) = 0.1 in. ; insulation between windings, T_d, $k = 3$; Turns, 4 000 c.t./12 000 c.t.

Distributed capacitance, using Figs. 5.13T and 13U $L_w = 0.6$ in., $L_{mt} = 2.5$ in., $T_w = 0.1$ in., T (primary) = 4 000, is $C_p = 58$ $\mu\mu$F. For secondary, $T = 12\,000$, $C_p = 34$ $\mu\mu$F.

Capacitance coupling between one-half primary and one-half secondary, using Figs. 5.13R and 13S, actual capacitance, $L_w = 0.6$ in., $L_{mt} = 2.5$ in., $T_d = 10$, $k = 3$, is $C_p = 100$ $\mu\mu$F. Both windings wound in same direction, turns factor across this capacitance referred to whole primary is $\frac{1}{2} + 1\frac{1}{2} = 2$, so referred capacitance is 400 $\mu\mu$F. Windings wound opposite directions, turns factor referred to primary is $1\frac{1}{2} - \frac{1}{2} = 1$, so referred capacitance is reduced to 100 $\mu\mu$F. Referring these two values to secondary, turns factors are $1/2 + 1/6 = 2/3$ or $1/2 - 1/6 = 1/3$, giving capacitance values referred to whole secondary of 45 $\mu\mu$F or 11 $\mu\mu$F respectively.

A complete analysis would need to consider leakage inductance between each primary and each half secondary, and separate source and load impedances applied to each half. For this purpose, the primary and secondary shunt capacitances across each half would be 116$\mu\mu$F and 68 $\mu\mu$F respectively, while the capacitance coupling would be 1,600 $\mu\mu$F or 400 $\mu\mu$F referred to half primary. The secondary shunt capacitances referred to the half primaries would be 610 $\mu\mu$F.

Example 3

A direct coupled inter-valve transformer is arranged as at Fig. 5.13M : $L_w = 1.5$ in., $L_{mt} = 6$ in., $T_d = 20$, $k = 1$ (air spaced) ; T_w (each whole winding) = 0.2 in. ; Turns 4 000/12 000 c.t.

Distributed capacitance, using Figures 5.13T and 5.13U, $L_w = 1.5$ in., $L_{mt} = 6$ in., $T_w = 0.2$ in., T (primary) = 4 000, gives $C_p = 400$ $\mu\mu$F. Secondary, $T = 12\,000$ gives $C_p = 225$ $\mu\mu$F (or 450 $\mu\mu$F per half secondary).

Interwinding capacitance, using Figs. 5.13R and 5.13S, actual capacitance, $L_w =$ 1.5 in., $L_{mt} = 6$ in., $T_d = 20$, $k = 1$, give RC_p just over 90 $\mu\mu$F say 100 $\mu\mu$F.

Vertical sectionalizing will reduce the distributed component in each case, but will also vary the interwinding capacitance. Using the information in Table 4 the results may be presented as in Table 6.

In practice three sections for the primary and four or five sections for each half-secondary will be best, remembering capacitance reduction is more important in the secondary. By making the earthy end of primary and secondary at opposite ends of the vertical groups, interwinding capacitance coupling effects are minimized.

TABLE 6

Number of vertical sections	Primary capacitance			Half secondary capacitance		
	Distributed	Interwinding	Total	Distributed	Interwinding	Total
1	400	100	500	450	—	450
2	100	75	175	112.5	12.5	125
3	44.4	70.4	115	50	18.5	68.5
4	25	69	94	28	22	50
5	16	68	84	18	24	42
6	—	—	—	12.5	25.5	38

(vi) Tests on output transformers

Summary of R.M.A. Standard SE-106A (Sound systems)

The distortion shall be measured with a zero-impedance source of voltage in series with a pure resistance R_{TG} of value 0.4 ($\pm$ 5%) times the square of the standard distribution voltage, V_{TR}, from which the tap is designed to work, divided by the manufacturer's rating, W_{TR}, for the power drawn by that tap at that distribution voltage :

$$R_{TG} = (0.4\ V_{TR}{}^2/W_{TR})(1 \pm 0.05).$$

Measurements shall be made at the lowest frequency of the rated frequency response or 100 c/s, whichever is the higher ; at 400 c/s and at 5000 c/s if within the rated frequency response.

The power-handling capacity of a speaker matching transformer is the maximum r.m.s. power drawn by the transformer at which the specified distortion (which shall be not more than 2%) is not exceeded. The power drawn by the transformer, W_T, shall be determined by dividing the square of the actual voltage measured across the primary terminals, V_{TP}, by the square of the standard distribution line voltage, V_{TR}, from which the primary tap is designed to work, and multiplying this quotient by the manufacturer's rating for the power, W_{TR}, drawn by that tap at that distribution voltage :

$$W_T = W_{TR}\ (V_{TP}/V_{TR})^2$$

The rating shall be determined as follows : The power at the stated distortion at 100 c/s shall be multiplied by 2. This figure shall be compared with the other measurements made at the other test frequencies, and the lowest power figure shall be taken as the power-handling capacity. In case the transformer is provided with more than one primary tap, the rating shall be given for the tap drawing the highest power at the rated distribution voltage. If there is more than one secondary tap, the power rating shall be given for the tap with the lowest measured power rating when properly terminated.

If the transformer is to be used in a system employing emphasized bass, a transformer must be chosen which has a rating higher than the nominal power to be handled in proportion to the bass emphasis employed. Likewise, transformers to handle organ music must have a rating at least four times the nominal power to be handled.

The frequency response of a speaker matching transformer is the variation of output voltage as a function of frequency, with a constant source voltage in series with a known impedance connected to the primary, expressed as a variation in db relative to the output voltage at 400 c/s.

For measurement, the transformer shall be connected as for distortion measurement (see above). The frequency response shall be measured using a constant source voltage, which will deliver one-half rated input power, W_{TR}, to the transformer at 400 c/s

The loss of a speaker matching transformer is the inverse ratio of the power delivered by the secondary of the transformer to a pure resistance equivalent to the rated load impedance, to the power delivered by the same source if the transformer is replaced by an ideal transformer of the same impedance ratio, expressed in db. For measurement, the transformer shall be connected as for distortion measurement (see above).

The impedance presented by the ideal transformer to the source shall be taken as

$$R_{T1} = V_{TR}^2/W_{TR}.$$

The power delivered to the secondary load is (V_{TS}^2/R_{TL}) where V_{TS} is the voltage across the load resistance R_{TL}.

The power delivered to the ideal transformer is (V_{T1}^2/R_{T1}) where V_{T1} is the voltage across the load resistance R_{T1}.

The loss is given then by

$$\text{Loss} = 10 \log_{10} \frac{V_{T1}^2/R_{T1}}{V_{TS}^2/R_{TL}}.$$

The loss shall be measured at 400 c/s and at a value of source voltage at which rated power is delivered to the ideal transformer. The loss shall not exceed 2 db (equivalent to 63% minimum efficiency).

(vii) Specifications for a-f transformers

The following details are suggested for forming part of a specification for a transformer. See also Ref. C32.

In all cases, it is desirable to submit a circuit showing the transformer application when writing a specification.

(1) Input transformers

(a) Operating level; this should be quoted in db above or below specified reference level, usually 1 milliwatt.

(b) Frequency response with permissible variation in db from a reference frequency, generally 1000 c/s.

Conditions of measurement must be specified—usually normal operating conditions.

(c) Impedance ratio or turns ratio.

(d) Positions of any taps should be stated.

(e) Source and source impedance.

(f) Load and load impedance. This may be the grid of a following amplifier valve. The secondary winding may also be shunted by a frequency-correcting network; if so, full details should be given.

(g) Total r.m.s. harmonic distortion; this should be measured at max. output at the lowest frequency of interest.

(h) Minimum resonant frequency.

(i) Insertion loss in db—frequently quoted at 400 c/s.

(j) Permissible phase characteristics at lowest and highest frequencies of interest.

(k) Direct currents in windings.

(l) Magnetic and electrostatic shielding.

(2) Interstage transformers

In general, as for input transformers, with the addition that the type of valves used in the preceding stage, together with their operating conditions, should be specified. Where push-pull input is intended, the maximum out-of-balance current should be stated.

(3) Output transformers

As (2) above. Where multiple secondary windings are employed the power to be delivered to each should be stated.

SECTION 4 : MAGNETIC CIRCUIT THEORY

(i) Fundamental magnetic relationships (ii) The magnetic circuit (iii) Magnetic units and conversion factors.

(i) Fundamental magnetic relationships

Just as we have an electrical circuit, so the core of a transformer can be regarded as a magnetic circuit through which a flux passes, its value depending on the magnetomotive force producing it and on the nature of the magnetic circuit. These are related by an equation resembling Ohm's Law for electrical circuits :

$$F = \phi R \text{ or } \phi = F/R \text{ or } R = F/\phi \qquad (1)$$

where F = magnetomotive force (m.m.f.) in **gilberts** (c.g.s. electromagnetic units) analogous to electromotive force (e.m.f.) in electrical circuits,

ϕ = total flux = the total number of lines of flux (or maxwells ; 1 maxwell = 1 line of flux),

R = reluctance (equivalent to resistance in electrical circuits). The c.g.s. electromagnetic unit of reluctance is the reluctance of 1 cubic centimetre of vacuum, which is very closely that of 1 cubic centimetre of air. Reluctances are combined in series or parallel like resistances; when in series they are added.

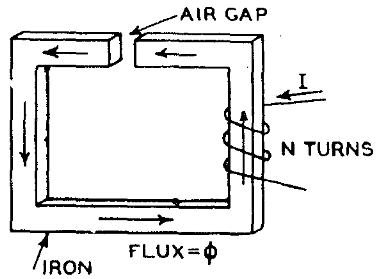


Fig. 5.14. Typical magnetic circuit with iron core and air-gap.

A typical magnetic circuit is shown in Fig. 5.14 where we have the greater part of the magnetic circuit through iron, and a short length (the air gap) through air. In this case the total reluctance is the sum of the reluctances of the iron-circuit and the air gap. The flux ϕ is caused by N turns of wire carrying a current I amperes and producing a magnetomotive force given by

$$F = (4\pi/10)\, NI \approx 1.257\, NI \qquad (2)$$

where F = magnetomotive force in gilberts,

N = number of turns,

and I = current in amperes.

Thus 1 ampere turn $\approx$ 1.257 gilberts. (3)

Instead of considering the total flux, it is often more convenient to speak of the **flux density,** that is the number of lines (maxwells), per square inch or per square centimetre (c.g.s. electromagnetic unit = 1 gauss = 1 maxwell per square centimetre). The symbol for flux density is B. Thus

$$\phi = BA \qquad (4)$$

where ϕ = total flux,

B = flux density, either in lines per square inch or in gauss (maxwells per square centimetre),

and A = cross sectional area of magnetic path (practically equal to the cross sectional area of the core in the iron section), in square inches or square centimetres respectively.

The **magnetizing force** (also known as the magnetic potential gradient or the magnetic field intensity) is defined as the magnetomotive force per unit length of path :

$$H = F/l \tag{5}$$

where H = magnetizing force in oersteds (or gilberts/centimetre),
F = magnetomotive force in gilberts,
and l = length of path in centimetres.

Alternatively, if F is expressed in ampere-turns, H may be expressed in ampere-turns per inch or per centimetre.

The **permeability** (μ) is defined by the relationship ;

$$\mu = B/H \tag{6}$$

where B = flux density in gauss (maxwells per square centimetre),
μ = permeability*,
and H = magnetizing force in oersteds.
In air, H is numerically equal to the flux density (B).

Permeability is the equivalent of conductivity in electrical circuits. Permeability in iron cores is not constant, but varies when the flux is varied. The relationship between B, H and μ is shown by the "BH characteristics" of the iron, as shown for example in Fig. 5.15. The value of μ at any point is the value of B divided by the value of H at that point.

For example, the permeability at point C is equal to OE/OD, where OE represents the flux density at point C, and OD represents the magnetizing force at point C. The permeability is therefore the slope of the line OC. A curve may be drawn indicating the value of μ for any value of H, and this curve may be plotted on the same graph with, of course, the addition of a μ scale (dashed curve in Fig. 5.15).

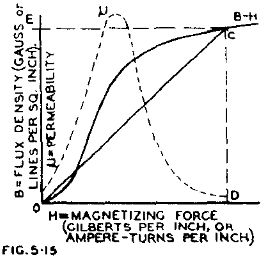


FIG. 5·15

Fig. 5.15. *BH characteristics of a typical transformer steel. The dashed curve is the permeability μ where $\mu = B/H$.*

The **incremental permeability** is the permeability when an alternating magnetizing force is superimposed on a direct magnetizing force.

The **initial permeability** is the permeability at values of H approaching zero.

The **hysteresis loop** of a typical magnetic material is shown in Fig. 5.16. When the magnetizing force increases from zero (demagnetized condition) to the positive peak, the B-H characteristic is followed from O to A, where A represents the maximum (peak) values of H and B. As the value of H decreases to zero and then increases in the opposite direction, the path followed is along the curve ACDE, where E is the negative equivalent to A, occurring half a cycle later than A. From E, the path followed is along the curve EFGA†. The area of the curve ACDEFGA represents power loss, known as hysteresis loss, in the magnetic material.

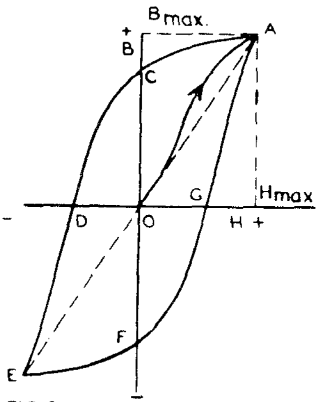


FIG. 5·16

Fig. 5.16. *Hysteresis loop of a typical magnetic material.*

Point C is the value of B for zero magnetizing force (i.e., $H = 0$) and it represents the residual magnetism ; this value of B is called the **remanence,** or **remanent flux density**‡.

*Strictly speaking, μ is measured in gauss per oersted, but it is common practice in engineering work to speak of the permeability as a pure number which is the ratio between μ and μ_o where μ_o is the permeability of vacuum, and has a value of unity.

†Actually this "cyclic condition" is not reached until after a number of cycles have occurred. In the early cycles the position of A falls slightly each cycle.

‡This strictly applies only to the initial cycle ; the term residual flux density is used for symmetrical cyclically magnetized conditions. The latter term is also sometimes used when it is not desired to distinguish between the initial and cyclic conditions.

Point D is that at which the applied negative magnetizing force brings the value of the residual magnetism to zero.

The length OC is called the **coercive force** (strictly this applies only for symmetrical cyclically magnetized conditions).

The **average permeability** is the slope of the straight line EOA (shown dashed in Fig. 5.16).

The locus of the extremities (A or E) of the normal hysteresis loops of a material is called its **normal magnetization curve** ; this is the same as the *B-H* curve of Fig. 5.15.

When alternating current is passed through the winding, the iron will pass through the hysteresis curve ACDEFGA (Fig. 5.16) each cycle. The maximum value of H is called H_{max}, and the corresponding value of B is called B_{max}, the curve being normally symmetrical in the positive and negative directions in the absence of a direct current component.

When dealing with alternating currents, it is usual to quote ampere-turns per inch in r.m.s. values ; the corresponding values of flux density may be quoted either in terms of B or B_{max}. It is obvious that $H_{max} = \sqrt{2}H$, but B_{max} is normally greater than $\sqrt{2}B$.

(ii) The magnetic circuit

A typical magnetic circuit is shown in Fig. 5.14. Certain assumptions are generally made for simple theoretical treatment, including—

1. That the flux confines itself entirely to the iron over the whole length of the iron path (in practice there is always some leakage flux, which is more serious when there is an air gap).

2. That the flux is uniformly distributed over the cross-sectional area of the iron.

In Fig. 5.14 we therefore have :

Total magnetomotive force $\quad F \approx 1.26\, NI$ from (2)

Total reluctance $\quad R = R_{iron} + R_{air}$

where R_{iron} = reluctance of iron path

and R_{air} = reluctance of air path.

Total flux $\phi = F/R = F/(R_{iron} + R_{air})$ from (1).

The reluctance of the air gap is given by

$$R_{air} = F/\phi = Hl/BA = kl/A \qquad (7)$$

where l = length of air gap,

A = equivalent area of air gap, allowing for " fringing "

$\approx (a + l)(b + l)$

and a, b = actual dimensions of pole faces.

Values of k are given by the table below :

H	oersteds	ampere turns/cm	ampere turns/inch
B	gauss	gauss	lines/sq. inch
l	centimetres	centimetres	inches
A	sq. centimetres	sq. centimetres	square inches
k	1	0.796	0.313

The reluctance of the iron path is not constant, so that the best approach is graphical.

The magnetic **potential difference** (U) is the equivalent of potential difference in electrical circuits. The sum of the potential differences around any magnetic circuit is equal to the applied magnetomotive force.

Applying this to Fig. 5.14,

$$F = U_{iron} + U_{air} \qquad (8)$$

where F = total magnetomotive force,
U_{iron} = magnetic potential difference along the whole length of iron,
and U_{air} = magnetic potential difference across the air gap.

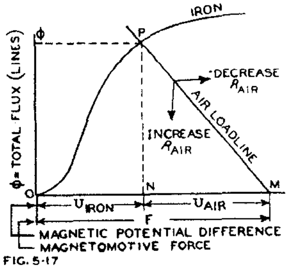


Fig. 5.17. Total flux versus magnetic potential difference, with air loadline.

This is applied in Fig. 5.17 where OM represents the applied magnetomotive force ($F \approx 1.26\ NI$), and the curve shows the total flux (ϕ) plotted against F for iron only. The shape of the curve is the same as that of the B-H characteristic of Fig. 5.15, but the vertical scale $\phi = Ba$, and the horizontal scale $F = lH$.

The " loadline " through M represents the effect of the air-gap ; it follows the equation

$$U = F - \phi R_{air} \tag{9}$$

and its slope is $-(1/R_{air})$. The intersection of the " loadline " and the curve at point P gives the operating point. Therefore ON represents the magnetic potential difference along the iron path, while NM represents the magnetic potential difference across the air gap.

It will thus be seen that variation of the air gap merely changes the slope of the " loadline " PM and moves point P, without changing the base line OM.

In order to be more generally applicable, the scales of Fig. 5.17 may be changed—
vertical scale : from ϕ to B (Note $\phi = Ba$)
horizontal scale : from F to H (Note $F = lH$)
from U to U/l (i.e., magnetic potential difference per inch)
as in Fig. 5.18A.

(iii) Magnetic units and conversion factors

The basic units generally adopted are the c.g.s. electromagnetic units such as the gauss, the oersted and the gilbert, with the centimetre as the unit of length. Practical units such as the lines per square inch and ampere-turns per inch are widely used in engineering design. More recently, the Giorgi M.K.S. system, with its webers and webers per square metre, has achieved considerable popularity. The full range of these various systems of units is given in Chapter 38 Sect. 1.

The following table of conversion factors will be helpful in converting from one system to another.

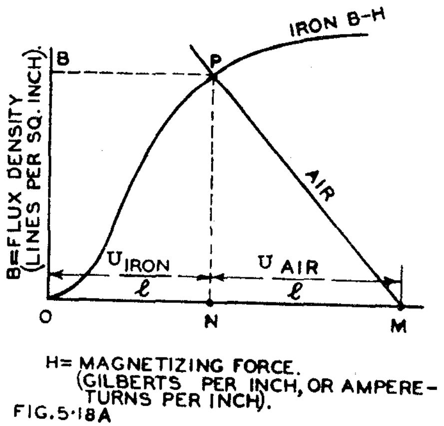


Fig. 5.18A. Flux density versus magnetizing force, with air loadline.

Magnetic Units—Conversion Factors

Multiply	by	to obtain
F in ampere-turns	$0.4\pi = 1.257$	F in gilberts
F in gilberts	$1/0.4\pi = 0.796$	F in ampere-turns
F in pragilberts*	0.1	F in gilberts
F in gilberts	10	F in pragilberts*
F in ampere-turns	$4\pi = 12.57$	F in pragilberts*
F in pragilberts*	$1/4\pi = 0.0796$	F in ampere-turns
H in ampere-turns/in.	$0.4\pi/2.54 = 0.495$	H in oersteds
H in oersteds	$2.54/0.4\pi = 2.02$	H in ampere-turns/in.
H in praoersteds*	10^{-3}	H in oersteds
H in oersteds	10^{3}	H in praoersteds*
H in ampere-turns/in.	495	H in praoersteds*
H in praoersteds*	0.00202	H in ampere turns/in.
B in maxwells/sq. in.	$1/6.45 = 0.155$	B in gauss
B in gauss B in maxwells/sq. cm.	6.45	B in maxwells/sq. in.
B in webers/sq. metre*	10^{4}	B in gauss
B in gauss	10^{-4}	B in webers/sq. metre*
B in maxwell/sq. in.	$10^{-4}/6.45 = 0.155 \times 10^{-4}$	B in webers/sq. metre*
B in webers/sq. metre*	6.45×10^{4}	B in maxwells/sq. in.
ϕ in maxwells ϕ in lines of flux	10^{-8}	ϕ in webers*
ϕ in webers*	10^{8}	ϕ in maxwells

SECTION 5 : POWER TRANSFORMERS

(i) General (ii) Core material and laminations (iii) Primary and secondary turns (iv) Currents in windings (v) Temperature rise (vi) Typical design (vii) Specifications for power transformers.

(i) General

The general design principles of power transformers have been dealt with in detail elsewhere (Refs. D1, D2, D3, D4) but an outline is given below of the design procedure for small power transformers for use in radio and electronic equipment. For transformers of this kind, efficiencies ranging from 80 to 90 per cent are common. The ratio of copper to iron losses is usually about 2 to 1. Winding capacitances have little effect on circuit operation and are usually neglected. For radio receivers it is common practice to provide some form of electrostatic screening between the primary and the other windings. This can be achieved in one of several ways (Ref. D5) :

(a) By using a turn of shim copper or brass of full winding width between primary and secondary, taking care to insulate the ends to avoid a shorted turn. This shield is then earthed.

(b) By winding the earthed low voltage filament between primary and secondary.

(c) By winding the high voltage secondary in two separate halves, one on top of the other, over the primary. The innermost and outermost leads are joined and become the centre tap which is then earthed. The two leads from the middle of the winding then become the high tension outers, and are connected to the rectifier plates. In this way the capacitance from primary to secondary is made very small. It is important in this method of construction to ensure that adequate insulation is used between the two plate leads within the winding, as the whole of the potential difference of the high tension winding appears between them.

With the normal method of construction the leakage inductance between either half of the secondary and the primary is unequal. In large transformers this becomes

*M.K.S. unit : See Chapter 38 Sect. 1.

important and, to maintain balance, the primary is wound between the halves of the secondary or the latter are wound side by side over the primary.

(ii) Core material and size

Various grades of core material are available all differing in silicon content from 0.25 to 4.5 per cent (Refs. D6, 7). They feature lower loss, increasing cost and brittleness with increase in silicon. Several manufacturers make somewhat similar grades and these are listed hereunder (Ref. D8).

TRANSFORMER STEELS

Silicon Content*	M. and E.A.	Baldwin	Sankey	Allegheny	Armco
4%	Silcor 1	Quality 5	Super Stalloy	Transf. C	Trancor 2
3½%	Silcor 2	Quality 4	Stalloy	Transf. D	Trancor 1
2¾%	Silcor 3	Quality 3B	42 Quality	Electrical	Spec. Elec.
1%	Silcor 4	Quality 1	Lohys	Armature	Armature

*This applies exactly to Silcor (Magnetic and Elec. Alloys Ltd.).

The core losses for the various Silcor grades measured at 50 c/s. with 0.014 inch sheet, at two different values of flux density are shown below.

	Watts lost per pound			
B_{max}	Silcor 1	Silcor 2	Silcor 3	Silcor 4
10 Kilogauss	0.59	0.63	0.89	1.32
13 Kilogauss	1.04	1.07	1.51	2.24

It will be observed that for small changes in flux density the core loss varies as the square of the flux density. For radio power transformer work, core materials similar to Silcor 2 are commonly used. Measurements of losses can be made with a low power factor wattmeter, or by the three ammeter method. A system suitable for mass production testing has been described recently (Ref. D9).

Cold rolled, grain oriented 2.7 per cent silicon steels are becoming of increasing importance. Typical trade names for this material are " Hipersil " and " Crystalloy " (Ref. 10). Flux densities in excess of 110 000 lines per square inch (17 kilogauss), can be employed without high core loss. Owing to low core losses at such high flux densities, the application of this material to small transformers for electronic equipment is increasing. It can be used in strip form, being wound around the winding in one method of assembly. In another method, the strip core is sawn in half and the two halves clamped together around the winding. Either method results in a considerable saving in material and labour over the present method of punching laminations and hand stacking the winding with them.

The most popular laminations in current use are the *E* and *I* " scrapless " variety (Refs. D11, 12, 13, 14). These are so dimensioned that the *I* is punched from the window of the *E* thus avoiding wastage. The usual ratio of dimensions of these laminations are as follows : Window height 1, tongue 2, window width 3, *I*-length 6, magnetic path 12.

Similar laminations are stamped by several different firms :

Standard Lamination Sizes

	Pattern Number				
Tongue	M.E.A.	Baldwins	Sankey	Allegheny	Chicago
9/16 in.	18	—	—	EI-56	—
⅝	145	392	—	EI-625	F000
¾	35	217	70	EI-75	D000
⅞	147	—	—	EI-11	B000
1	29	430	111	EI-12	1000
1⅛	—	—	158	EI-112	13000
1¼	78	420	133	EI-125	14000
1⅜	152	362	—	EI-138	—
1½	120	—	149	EI-13	—

Laminations for small power transformers are generally 0.014 inch thick ; thicker laminations up to 0.025 inch are occasionally used, but result in increased losses and shortened lamination die life.

To determine the core size, it is first necessary to estimate the power requirements. Then we may apply the empirical relation

$$A = \frac{\sqrt{V.A}}{5.58} \qquad (1)$$

where A = cross-sectional area in square inches
and VA = voltamps output.

For ease of production an approximately square stack is desirable. Thus having determined A, the tongue size can be estimated from $\sqrt{A}$. Where high voltage windings, or windings operating above ground, or at a high potential between other windings, are used, it becomes necessary to employ a lamination which has a different ratio of dimensions from those quoted earlier, the window height being increased to allow room for extra insulation. The input current taken by a transformer on no-load is commonly called the *magnetizing current*. In fact, it consists of two components in phase quadrature. The in-phase component, usually small, is the *iron* loss plus a very small copper loss. The quadrature portion is the true magnetizing current. Their vector sum does not normally exceed about a third of the full load current.

(iii) Primary and secondary turns

The primary turns required can be determined from the fundamental transformer equation.

$$N = \frac{E \times 10^8}{4.44\, fBA} \qquad (2)$$

where N = primary turns, f = frequency in c/s
B = max. flux density
and A = cross-sectional area of core in sq. ins.

Assuming an average stacking factor of 90 per cent, the factor 4.44 becomes 4 in the denominator. The stacking factor reduces the apparent height of the stack, the reduction being caused by insulation, scale and burr due to die wear. With very thin laminations, the stacking factor decreases to 80 per cent approximately. For a flux density of 64 500 lines per square inch (10 kilogauss) eqn. (2) becomes :

For 240 volts 50 c/s	230 volts 50 c/s	117 volts 60 c/s
$N = \frac{1860}{A}$	$N = \frac{1780}{A}$	$N = \frac{755}{A}$

where A = gross cross-sectional area of core in sq. ins.

The flux density employed depends on the application, the power rating, the core material and the frequency. For oscilloscopes and pre-amplifiers, densities of 40 000 to 50 000 lines per square inch (about 7 000 gauss) are used. For small transformers below about 50 watts, densities up to 90 000 lines per square inch (14 000 gauss) are used, gradually decreasing to about 65 000 lines (10 000 gauss) as the transformer size increases to several hundred watts.

Having chosen a suitable flux density, the turns required for each secondary winding may be calculated.

It can be shown that, approximately,

$$N_2 = \frac{E_2 \cdot N_1}{E_1 \cdot \sqrt{\eta}} \qquad \text{where } \eta = \text{transformer efficiency.}$$

As a first approximation it may be assumed that $\eta = 0.85$ and $\sqrt{\eta} = 0.92$.

The values thus obtained for the secondary turns may be checked by more detailed calculations once wire gauges have been chosen and winding resistances calculated. For this purpose the equivalent circuit of Figure 5.11 may be used.

If a secondary is to feed a rectifier and the d.c. output of the valve is specified, reference should be made to Chapter 30 or a valve data book to determine the required secondary voltage. Allowance must be made for any voltage drops due to the d.c. resistance of the rectifier filter.

(iv) Currents in windings

To enable the wire gauges to be chosen and to assess the copper losses in the windings, it is necessary to estimate the current in each winding. Where windings are used to supply valve heaters and resistance loads, the winding current is the same as the load current. In a secondary winding feeding a rectifier the winding current must be estimated from a knowledge of the type of rectifier and its associated filter and their characteristics. For normally-loaded full wave rectifiers the following values of secondary current may be used as a fairly close guide for design purposes.

Condenser input filter : The r.m.s. current in each half of the transformer secondary may be taken approximately as 1.1 times the direct current to the load. For further details see Chapter 30 Sect. 2.

Choke input filter : The r.m.s. current in each half of the transformer secondary may be taken approximately as 0.75 times the direct current to the load. For further information, see Chapter 30, Sect. 4 (also Sect. 3). With half-wave rectification there will be a d.c. component of the current which will affect the transformer design if the total d.c. ampere turns are considerable. For half-wave battery chargers using bulb rectifiers, this must be taken into consideration when selecting core size and wire gauges (Ref. D15).

In the case of transformers supplying full-wave rectifiers, the full load primary current can be estimated by calculating the total secondary loading in voltamps, allowing an efficiency of 85% as a first approximation. This is then the primary input in voltamps which, when divided by the primary voltage, will give the desired current. Where secondaries feed a resistive load, the loading is the product of the voltage and current. Where a secondary feeds a full-wave rectifier the load is the product of the direct current and the direct voltage output from the rectifier plus the power lost in the rectifier. This latter can be calculated, for a condenser input filter, from data presented in Chapter 30, Sect. 2. It should be noted that indirectly heated, close-spaced rectifiers such as the 6X4, are more efficient than types such as the 5Y3-GT, with its heavier filament power and lower plate efficiency. This is one reason why the former are almost exclusively used in the majority of small a.c. radio receivers.

The primary input current as calculated above should be accurate to within 10% for small transformers.

Wire gauges and copper losses

For the usual type of radio receiver power transformer it will be safe to choose wire gauges on the basis of 450-800 circular mils* per ampere (i.e. 2830-1590 amperes per square inch), but values up to 1000 circular mils per ampere (i.e. 1270 amperes per square inch) may be desirable in larger units, or if a high flux density is used with high loss lamination steel. The latter figure is easy to remember as 1 circular mil per milliamp.

In practice, wire gauges may be chosen arbitrarily on a basis of (say) 700 circular mils per ampere (i.e. approximately 1800 amperes per square inch) and check calculations should then be made to see that

(a) the build of the winding (i.e. winding height) is satisfactory for the window space available,
(b) the copper loss is not so high as to cause excessive temperature rise,
(c) the voltage regulation is satisfactory.

Wire tables in this Handbook will simplify these calculations (Chapter 38 Sect. 19).

(v) Temperature rise

This is dependent upon the cooling area, the total loss and the ratio of iron to copper loss. To avoid deterioration of the insulating materials, it is necessary to limit the working temperature to 105°C for Class A insulation which includes paper, cotton,

*A circular mil is the area of a circle 1 mil (1/1000 inch) in diameter.

silk, varnish and wire enamel (Refs. D16, 17). The temperature rise in the winding, as measured by the change of resistance method, will be about 10°C lower than the maximum (hot spot) temperature. Thus with an ambient temperature of 40°C (104°F) plus a margin of 10°C for the difference between measured and hot spot temperature, it will be seen that the maximum permissible rise is 55°C, as measured by the change of resistance.

It is common practice to allow 10°C margin for change in line voltage, frequency, or operation in situations with restricted ventilation. Thus 45°C is generally accepted as the maximum permitted rise above ambient when measured by resistance change.

The temperature difference between winding and core varies between 10° and 20°C according to the distribution of losses. This means that even with an ambient temperature of 25°C (77°F), the core temperature may be 60° (140°F). This will feel quite hot to the touch, although the internal temperature may be well under the permitted maximum. Measurement of the core temperature may be made with a spirit thermometer if good thermal contact is maintained between the core and the thermometer bulb.

The winding temperature rise can be calculated by measuring the cold resistance, R_0 at an observed temperature T_1. After a heat run at full load the hot resistance R is measured and the ambient temperature T_2 is again measured.

Taking the temperature coefficient of resistivity (a) of copper as 0.003 93, the temperature rise T is found from the formula

$$T = \frac{R - R_0}{R_0 a}.$$

To correct for any changes in ambient temperature during the heat run it is necessary to subtract the difference $T_2 - T_1$ from T to find the actual rise. An example will illustrate this.

The primary resistance of a transformer measured when cold was 30 ohms at an ambient temperature of 20°C. After an 8 hour full load run, the resistance was 36 ohms, while the ambient temperature was then 18°C. It is required to find the winding temperature rise.

$$T = \frac{36 - 30}{30 \times 0.00393} = \frac{6}{30 \times 0.00393} = 51°\text{C}.$$

Now $T_2 - T_1 = (18 - 20)°\text{C} = -2°\text{C}$.

Winding temperature rise $= [51° - (-2)]°\text{C} = 53°\text{C}$.

Standard methods of testing small radio receiver power transformers have been published (Refs. D18, D22). The cooling area of a transformer for the " scrapless " type laminations with standard ratio of dimensions is

$$A = T(7.71T + 11S)$$

where A = cooling area

T = width of tongue

and S = height of stack.

For a square stack $S = T$ and $A = 18.71T^2$ (Ref. 11).

Having calculated the iron and copper losses and the total cooling area, it is possible to estimate the temperature rise before a sample transformer is wound. The watts lost per square inch of cooling area is first calculated. Reference to Fig. 5.18B will then show the temperature rise to be expected with an accuracy of approximately ± 10%.

(vi) Typical design

(Refs. D19, D20, D21, D23).

Specifications :

Primary	240 V 50 c/s
Secondary (i)	6.3 V 2.6 A
Secondary (ii)	300 V + 300 V r.m.s. for full wave 6X4 rectifier with condenser input filter to deliver 60 mA d.c.

Secondary loading

Secondary (i)	6.3 V 2.6 A	or 16.5 watts
D.C. output	343 V 0.06 A	or 20.5 watts

Rectifier loss (see Chap. 30) 2.0 watts approx.
Total loading 39.0 watts
Primary input (assuming an efficiency of 85%) 46.0 watts.

Thus the input current is 46/240 or 0.19 A. Core cross-section : $A = \sqrt{39}/5.58 = 1.12$ square inches. Choosing pattern EI-112 lamination of Silcor 2 material, a suitable stack for calculated cross section is 1.125 inches, that is, a square stack.
Primary turns :

With a flux density of 13 kilogauss (84 000 lines per square inch) and a stacking factor of 0.9,

$$N_p = 240 \times 10^8/4 \times 50 \times 84\,000 \times 1.125 \times 1.125$$
$$= 1130.$$

Secondary (ii) turns :

$$N_1 = E_1N_p/E_p\sqrt{\eta} = 300 \times 1130/240\sqrt{0.85}$$
$$= 1530 + 1530.$$

Turns per volt $= N_1/E_1 = 1530/300 = 5.1$.
Secondary (i) turns $= (6.3 \times 5.1) = 32$.
Wire gauges :

Assume a current density of 600-700 circular mils per ampere. Referring to wire tables, a suitable primary gauge is 29 A.W.G. enam.

Turns per layer = 95. Refer Figs. 5.18C and wire tables in Chapter 38 Sect. 19.
No. of layers = 1130/95 = 12.
R.M.S. current in secondary (ii) is (0.06 × 1.1) = 0.066 A.

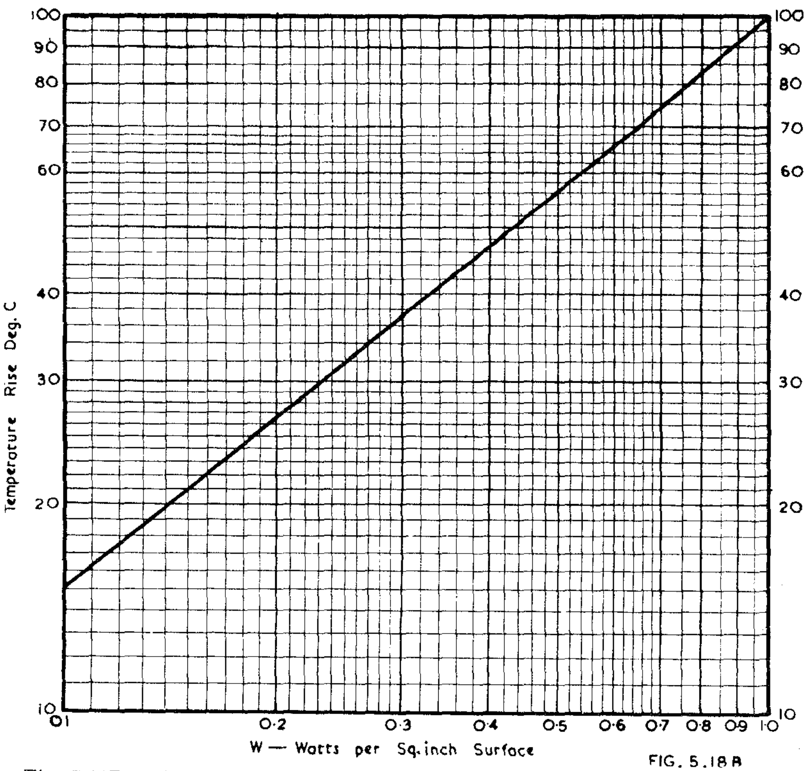


Fig. 5.18B. *Temperature rise versus loss per square inch of cooling surface for ambient temperature* 25°C (*Ref.* D.1.)

Suitable wire gauge is 34 A.W.G. enam.
Turns per layer = 167.
No. of layers = 3060/167 = 19.
Suitable secondary (i) gauge is twin 20 A.W.G. enam.
Turns per layer of twin wire = 16.
No. of layers = 32/16 = 2.
A twin wire is used in preference to a single wire of 17 or 18 A.W.G. in order to save winding height.

Winding Build :

Primary build = No. of layers × (enam. wire diam. + interlayer insulation)
= $12 \times (12.2 + 2) \times 10^{-3}$ inches
= 0.170 inch.

Secondary (ii) build = $19 \times (6.9 + 1) \times 10^{-3}$ inch
= 0.150 inch.

Secondary (i) build = $2 \times (32.4 + 5) \times 10^{-3}$ inch
= 0.075 inch.

Allowing 50 mil former thickness and 10 mil insulation between windings and over outer winding, total build is (0.170 + 0.150 + 0.075 + 0.050 + 0.040) inch
= 0.485 inch.

Winding height = 0.562 inch.

Build expressed as a percentage of window height
= 0.485/0.562 × 100 = 86.5%.

In this particular design, the heater winding is wound between primary and high tension windings to serve as a static shield.

Mean length of turn calculations—see Fig. 5.18D.

A = Former build plus insulation	= 0.050 + 0.010 = 0.060 inch
B = Primary build plus insulation	= 0.170 + 0.010 = 0.180 inch
C = Secondary (i) build plus insulation	= 0.075 + 0.010 = 0.085 inch
D = Secondary (ii) build plus insulation	= 0.150 + 0.010 = 0.160 inch
4S = 2 × (stack + tongue)	= 4.5 inches
Primary mean length of turn	= 5.44 inches
Secondary (i) mean length of turn	= 6.27 inches
Secondary (ii) mean length of turn	= 7.04 inches.

Winding resistance :

If each value for the mean length of turn is multiplied by the number of turns in its own winding and then divided by twelve, the resulting quantity will be the number of feet of wire in each winding. By referring to the wire tables, the resistance in ohms per thousand feet for any particular gauge can be found. Dividing the wire length by a thousand and multiplying by the resistance per thousand feet, will determine the winding resistance.

Primary resistance	= 5.44 × 1130 × 81.8/12 × 1000	= 42 ohms
Secondary (i) resistance	= 6.27 × 32 × 12/2 × 12 × 1000	= 0.09 ohms
Secondary (ii) resistance	= 7.04 × 3060 × 261/12 × 1000	= 470 ohms

Copper loss :

Primary copper loss	= $(0.19)^2 \times 42$	= 1.52 watts
Secondary (i) copper loss	= $(2.6)^2 \times 0.09$	= 0.61 watts
Secondary (ii) copper loss	= $(0.066)^2 \times 470$	= 2.05 watts
Total copper loss		= 4.18 watts

Iron loss :

Loss per pound of Silcor 2 at flux density of 13 kilogauss = 1.07 watts.
Weight of core = 2.17 lbs.
Therefore total iron loss = (1.07 × 2.17) = 2.32 watts.

Temperature rise :

Total of iron and copper loss = (2.32 + 4.18) = 6.5 watts.
Cooling area of square stack = $18.71 \times (1.125)^2$ = 23.6 sq. inches.
Therefore watts lost per sq. inch = 6.5/23.6 = 0.275 watts.
From Fig. 5.18B temperature rise = 35°C.

Efficiency :

Efficiency = Power output/(power output plus losses)
$= 39/(39 + 6.5) \times 100 = 86\%$.

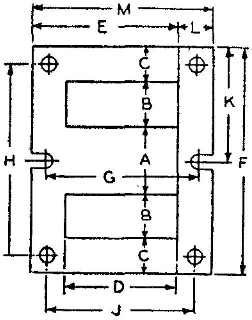
Regulation, calculated from British definition :

Primary voltage drop $= 0.19 \times 42 = 8$ V

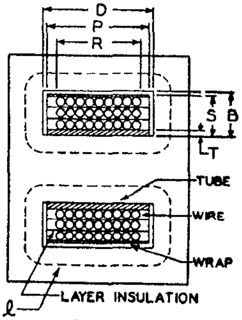
Regulation due to resistance of primary $= 8 \times 100/240 = 3.3\%$.

FIG. 5.18C

LAMINATION AND COIL DATA



EXAMPLE:
*24E WIRE ON SQUARE 1.25" CORE
LAMINATION
SIZE ------ 1.25"
OVERALL ---- 3.75" X 3.12"
4 HOLE MTG. -- 3.12" X 2.5"
CORE
WEIGHT ---- 2.95 LBS.
AREA ----- 1.56 SQ."
MAGNETIC PATH -- 7.5"
COIL - *24E
WIRE --- 0.578 AMPS MAX
WIRE --- 61 TURNS/LAYER
WIRE --- .0213" DIAMETER
INSULATION - .004" KRAFT
TUBE --- .050" PAPER
BUILD --- .538" MAX.



Lamination Size	Width Center Leg, In.	A.	0.500	0.625	0.750	0.875	1.000	1.125	1.250	1.375	1.500
Window Dimensions	Area, Sq. In.	B x D	0.250	0.292	0.422	0.573	0.750	0.950	1.17	1.42	1.69
	Length, In.	D	0.812	0.937	1.125	1.31	1.50	1.69	1.87	2.06	2.25
	Width, In.	B	0.312	0.312	0.375	0.437	0.500	0.562	0.625	0.687	0.750
Coil Dimensions	Build, Max. In.	S	0.256	0.256	0.315	0.367	0.430	0.483	0.538	0.590	0.646
	Length Overall, In.	P	0.750	0.875	1.06	1.25	1.44	1.56	1.75	1.94	2.12
	Length Winding, In.	R	0.500	0.625	0.750	1.00	1.12	1.25	1.37	1.56	1.75
Core Tube	Thickness, In.	T	0.035	0.035	0.050	0.050	0.050	0.050	0.050	0.050	0.060
"E" & "I" Piece Overall Dimensions	Length, In.	F	1.62	1.87	2.25	2.62	3.00	3.37	3.75	4.12	4.50
	Width, In.	M	1.31	1.62	1.87	2.19	2.50	2.81	3.12	3.44	3.75
4 Hole Mounting Dimensions	Length, In.	H	...	....	....	....	2.50	2.81	3.12	3.44	3.75
	Width, In.	J	....	....	....	....	2.00	2.25	2.50	2.75	3.00
	Diameter Hole, In.	...	....	....	....	....	0.219	0.219	0.219	0.219	0.281
2 Slot Mounting Dimensions	Width, In.	G	....	....	1.62	1.95	2.16	2.44	2.78	2.94	3.12
	From End, In.	K	....	....	1.12	1.31	1.50	1.69	1.87	2.06	2.25
"E" Piece Dimensions	Width, Overall, In.	E	1.06	1.25	1.50	1.75	2.00	2.25	2.50	2.75	3.00
	Width, Outer Leg, In.	C	0.250	0.312	0.375	0.437	0.500	0.562	0.625	0.687	0.750
"I" Piece	Width, In.	L	0.250	0.312	0.375	0.437	0.500	0.562	0.625	0.687	0.750
For Square Stack Center Leg	Volume, Cu. Inches	Area x l	0.810	1.46	2.54	4.03	6.00	8.55	11.70	15.60	20.20
	Weight, Pounds	Vol. x Density	0.21	0.37	0.64	1.05	1.50	2.14	2.95	3.90	5.06
	Area, Sq. Inches	A x A	0.250	0.390	0.562	0.766	1.000	1.27	1.56	1.90	2.25
Magnetic Path	Length, In.	l	3.25	3.75	4.50	5.26	6.00	6.75	7.50	8.24	9.00

Fig. 5.18C. *Lamination and coil data (Ref. D20).*

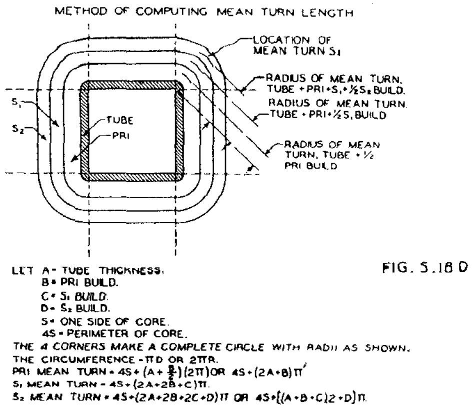


Fig. 5.18D. *Method of computing mean length of turn* (*Ref.* D20).

The percentage regulation of any secondary winding is calculated by dividing the full-load voltage drop of the winding by the open-circuit winding voltage and multiplying by 100. By adding the percentage regulation of the primary winding, the overall regulation for the secondary winding under examination is found.

Secondary (i) voltage drop	$= 2.6 \times 0.09 = 0.234$ V
Regulation, overall (i)	$= (0.234 \times 100/6.8) + 3.3 = 6.7\%$
Secondary (ii) voltage drop	$= 0.066 \times 470 = 31$ V
Regulation, overall (ii)	$= (31 \times 100/650) + 3.3 = 8.1\%$
Full load secondary (i) voltage	$= 6.35$ V
and full load secondary (ii) voltage	$= 300$ V $+ 300$ V.

This should be a satisfactory design.

As the temperature rise is low, it might be feasible to effect economies by re-design. For example, a higher loss core material could be used. Alternatively the flux density could be increased by decreasing the turns or the size of the stack. A new design could also be tried on the next smaller size lamination.

(vii) Specifications for power transformers

The following may be incorporated into a specification for a power transformer.

(a) Input voltage and frequency ; should these quantities vary, the range of variation must be stated.
(b) Secondary full load voltages and currents. If tapped windings are required this should be indicated. Tolerances on voltages should be stated.
(c) Regulation of secondary voltages other than for rectifier plate circuit.
(d) Power factor of each secondary load.
(e) Rectifier system, if d.c. output is required ; e.g. full wave, half wave, etc.
(f) Type of rectifier valve to be used.
(g) Type of filter circuit ; e.g. choke or condenser input, etc.
(h) Capacitance of input condenser, or inductance and resistance of input choke.
(i) D.C. full load current and voltage at filter input.
(j) D.C. regulation.
(k) External voltages between windings, or between windings and ground. Should transformer windings be interconnected externally, this should be indicated.
(l) Static shields. Number and position of each should be indicated.
(m) Any limiting dimensions together with mounting and lead terminations.
(n) Ambient temperature in which the transformer is to operate.

A circuit should be supplied showing intended use of transformer.

In all cases it is desirable for the design of the whole rectifier system, comprising transformer, rectifier and filter system, to be carried out by the one engineer.

SECTION 6 : IRON-CORED INDUCTORS

(i) General (ii) Calculations-general (iii) Effective permeability (iv) Design with no d.c. flux (v) Design of high Q inductors (vi) Design with d.c. flux (vii) Design by Hanna's method (viii) Design of inductors for choke-input filters (ix) Measurements (x) Iron-cored inductors in resonant circuits.

(i) General

Iron cored inductors fall into several different categories depending upon the circuit requirements. In some applications these inductors may have to carry a.c. only, in others, both a.c. and d.c. They may have to work over a wide range of frequencies, or at any single frequency up to the ultrasonic range. Iron-cored inductors are employed as smoothing and swinging chokes in power supplies, as equalizer elements in audio frequency equipment, as modulation chokes and as filter elements in carrier equipment, to mention a few varied applications.

To design such an inductor it is therefore necessary to know some or all of the following specifications—

(a) inductance, or range of inductance if variable,
(b) alternating voltage across the coil,
(c) direct current through the coil,
(d) frequency of operation,
(e) maximum shunt capacitance,
(f) minimum frequency of self-resonance,
(g) minimum Q over frequency range and/or
(h) frequency of maximum Q,
(i) d.c. resistance,
(j) shielding,
(k) temperature rise,
(l) size and weight limitations,
(m) insulation requirements.

(ii) Calculations—general

The inductance L of an iron-cored coil may be calculated using the relation

$$L = \frac{3.2 \times N^2 \times \mu \times a}{10^8 \times l} \text{ henrys} \quad (1)$$

where N = number of turns
a = effective cross sectional area of coil in square inches
l = length of magnetic circuit in inches
and μ = effective permeability.*

The effective permeability depends on the type of steel used, on the a.c. and d.c.

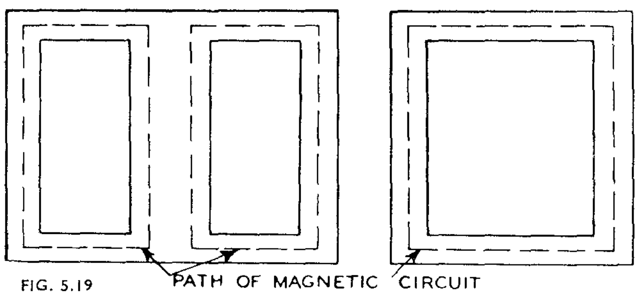


Fig. 5.19. *Magnetic cores, showing method for calculation of length of magnetic path.*

*A more precise term is "inductance ratio."

flux densities in the core, and on whether the core laminations are interleaved or whether there is an air gap in the magnetic circuit.

The length of the magnetic circuit is the length measured around the core at the centre of cross section of each magnetic path. Referring to Fig. 5.19, the path taken will be along the centre of each leg except where there are two windows, when each path through the centre leg will be along a line one-quarter of the way across the leg. In this latter case only a single path round the window is considered in calculating the magnetic circuit length.

Eqn. (1) is properly applicable only when the cross-sectional area of the core is uniform throughout the magnetic circuit. Where the cross section is non-uniform, a conservative value of L will usually be obtained by using for a the minimum value of the cross-sectional area. For more accurate calculations in such cases reference should be made to a suitable text-book (e.g. Ref. E1).

(iii) Effective permeability

For any given sample of lamination steel the effective permeability depends primarily on the a.c. and d.c. flux densities in the core.

Irrespective of the presence of air gaps in the magnetic circuit, the maximum a.c. flux density B_{max} in the core is determined by the cross section of the core, the number of turns in the coil, the alternating voltage across the coil and the frequency, and may be calculated directly from the relation derived from eqn. (2) in Sect. 5,

$$B_{max} = \frac{E \times 10^8}{4.44 fNa} \tag{2}$$

The d.c. magnetizing force in the core depends on the total number of d.c. ampere-turns, less the number of ampere-turns absorbed in any air gap, divided by the length of the magnetic circuit in inches (see Sect. 4).

Fig. 5.20 shows the variation of effective permeability with variation of a.c. flux density and d.c. magnetizing force for typical electrical sheet steel. It will be observed that the effective permeability increases up to a maximum as the a.c. flux density increases, and then drops rapidly due to saturation of the core. For any particular value of a.c. flux density the effective permeability decreases as the d.c. magnetizing force is increased.

Where the a.c. flux in the core is much less than the d.c. flux, it is convenient to refer to the effective permeability as the **incremental permeability** since it depends on the characteristics of the core material in relation to small changes in the total flux. This term is therefore generally used in dealing with filter chokes and other inductance coils having a relatively large number of turns and carrying a relatively large direct current.

(iv) Design with no d.c. flux

If there is no d.c. flux in the core the calculation of inductance is straightforward because the effective permeability depends only on the number of turns, the frequency, the core cross section, and the applied voltage. This case arises when the windings of a coil or transformer do not carry direct current or when the windings carrying direct current are so arranged that the d.c. flux in the core is zero as, for example, with transformers used in balanced push pull amplifiers.

Design of a coil to obtain a required value of inductance is therefore also quite straightforward, and the following steps may be followed—

(1) Choose an available core and calculate from eqn. (1) the number of turns required to give the necessary inductance assuming a value of effective permeability of between 1 000 and 5 000 depending on a very rough estimate of the a.c. flux density in the core.

(2) Calculate the a.c. flux density for this number of turns in relation to the known frequency and voltage across the coil.

(3) Correct the estimate of the number of turns using a revised value of permeability for the calculated flux. If the first estimate was far out, repeat steps (2) and (3).

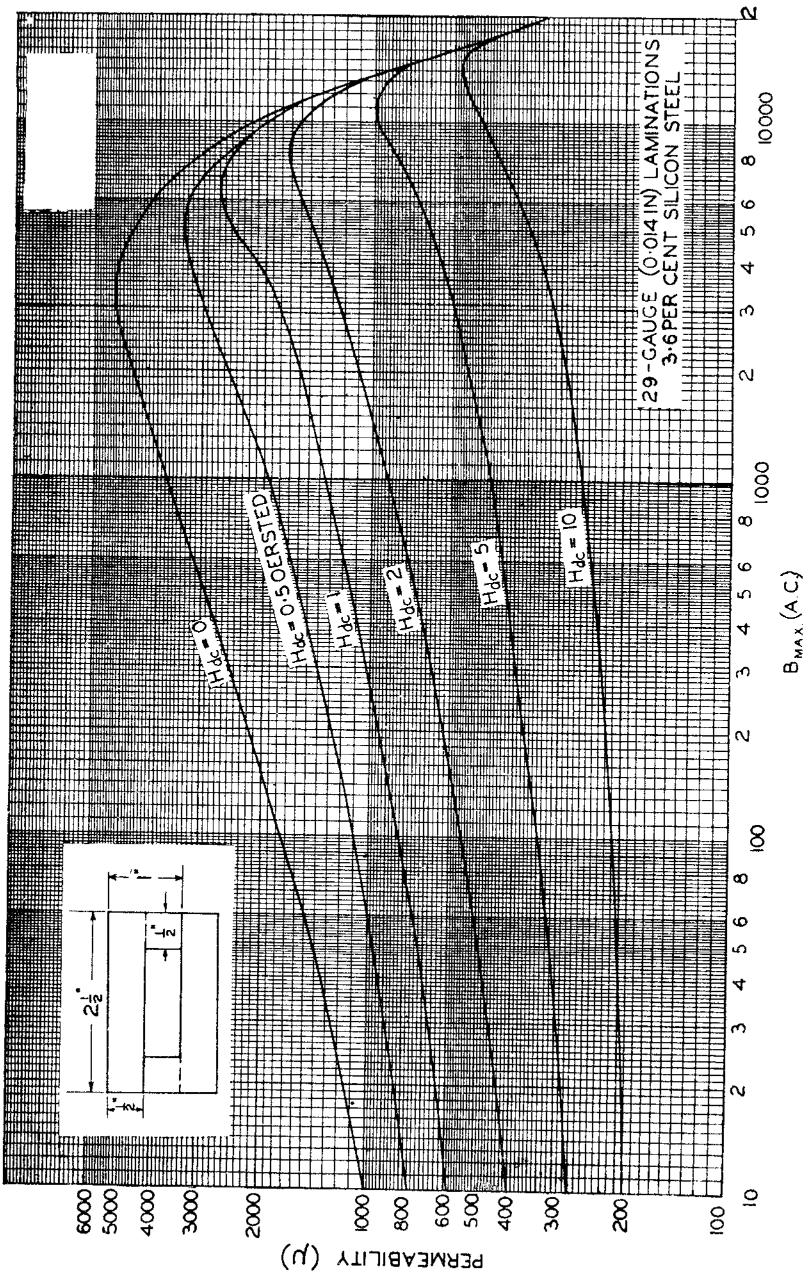


Fig. 5.20. *Effective permeability* (μ) *of silicon steel versus a.c. maximum flux density, for various values of d.c. magnetising force* H_0 *(by courtesy of Allegheny Ludlum Steel Corporation).*

(4) If the flux density is excessive (i.e. above about 50 000 lines per square inch) *increase the number of turns until the flux density is satisfactory.*

In case step 4 is necessary, the magnetic circuit must be broken by an air gap and the gap adjusted until the required value of inductance is obtained ; but if the larger value of inductance is satisfactory or useful the air gap may be omitted.

(5) Calculate the wire gauge required, and the copper and iron losses as in the design of a power transformer. Check also that the d.c. resistance of the winding is not excessive for the purpose for which the coil is to be used.

If an air gap of length α inches is used in the core, the value of inductance becomes

$$L = \frac{3.2N^2 a}{10^8 . l} \times \frac{1}{(1/\mu + \alpha/l)} \tag{3}$$

and this expression may be used for calculations. Alternatively, it will be clear that if the air-gap is large, so that $1/\mu \ll \alpha/l$, eqn. (3) approximates to

$$L = 3.2N^2 \, . \, a/10^8\alpha \tag{4}$$

FIG. 5.21

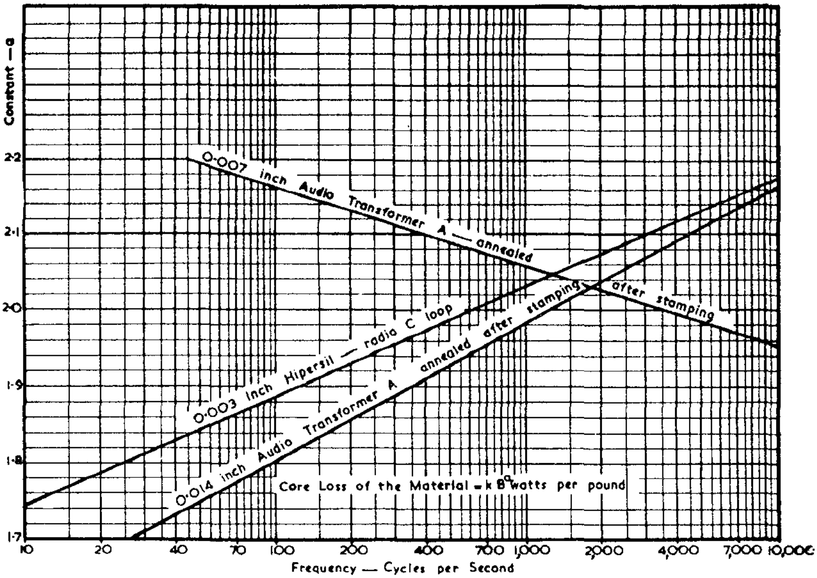


Fig. 5.21. Experimentally determined curve showing values of " a " as a function of frequency for several magnetic materials (Ref. E2).

(v) Design of high Q inductors

This sub-section follows S. L. Javna (Ref. E2) and is set out in a step-by-step form for ease of working. The following table defines the symbols used by Javna in the formulae presented here.

A = Effective cross-sectional area of magnetic-flux path, in sq. in.
A_g = Gross cross-sectional area of magnetic-flux path, in sq. in.
a = Empirical constant, see Fig. 5.21.
B = Maximum flux density in the core, in lines per sq. in.
F = Fraction of core-window area occupied by copper wire of coil
f = Frequency, in cycles per second.
g = Actual gap length, in inches.
k = Empirical constant, see Fig. 5.22.
L = Inductance of iron-cored coil, in henrys.
l = Mean length of magnetic path, in inches.
m = Mean length of a turn of the coil, in inches.

$\mu_\triangle$ = Incremental permeability of core with respect to air at operating frequency.
N = Number of turns in coil.
n = A.W.G. (B. and S.) wire gauge number of conductor.
$R_{a.c.}$ = Apparent a.c. coil resistance caused by core loss, in ohms.
$R_{d.c.}$ = Copper loss resistance, in ohms.
s = Total lamination window area, in sq. in.
V = Voltage across coil, in volts.
w = Weight of core, in pounds.

A typical problem will be solved to illustrate the design procedure.

A 5 henry inductor is to be designed on a one inch square stack of Allegheny pattern EI-12 audio transformer A silicon steel annealed laminations. The voltage across the coil will be 10 V a.c. at 1000 c/s. It is required to determine the turns and gauge of wire, the gap width and Q at the operating frequency. Assume a stacking factor of 0.9.

Tabulation :

$$\left.\begin{aligned} l &= 6.0 \text{ ins.} \\ w &= 1.5 \text{ lbs.} \\ A_g &= 1.0 \text{ sq. in.} \\ A &= 0.9 \text{ sq. in.} \\ s &= 0.75 \text{ sq. in.} \end{aligned}\right\} \quad \text{from Fig. 5.18C}$$

m = 5.5 ins. from Fig. 5.18D
$F \approx$ 0.3, a typical value for this lamination
k = 1.3×10^{-8} from Fig. 5.22.
a = 1.987 from Fig. 5.21.
$\mu_\triangle$ is found from Fig. 5.20.

Calculation :

$$B = \left(\frac{1.74mV^4 10^7}{akwf^4L^2A^2sF}\right)^{1/(a+2)} = 48.6 \text{ lines/sq. in.}$$

$$R_{a.c.} = 39.5kB^a wf^2L^2/V^2 = 430 \text{ ohms}$$

$$R_{d.c.} = \frac{3.44mV^2 10^8}{sFB^2A^2f^2} = 440 \text{ ohms}$$

$$Q = \frac{2\pi fL}{R_{a.c.} + R_{d.c.}} = 36$$

$$N = \frac{V\,10^8}{4.44\,fBA} = 5150 \text{ turns.}$$

$$n = 49.8 + 9.96 \log (1.2R_{d.c.}/mN) = 33 \text{ A.W.G.}$$

$$g = \left(\frac{1.59N^2A_g}{L10^8} - \frac{lA_g}{2\mu_\triangle A}\right) = 0.08 \text{ in.}$$

Coil build check :
Turns/layer using 33 A.W.G. enam. = 133 from winding data chart,
thus number of layers = 5150/133 = 39.
Interlayer insulation from winding data chart = 0.0015 in.,
thus winding build = 39 (0.0078 + 0.0015) = 0.363 in.
Former thickness = 0.05 in.
Insulation under and over winding = 0.02 in.,
thus total coil build = (0.363 + 0.05 + 0.02) = 0.433 in.
Max. build from Fig. 5.18C = 0.43 in.,
thus the winding should fit the window satisfactorily.

It should be noted that $R_{a.c.}$ and $R_{d.c.}$ are usually very nearly equal and where a square stack of scrapless laminations is used, m will approximate l. Due to the large air gap the inductance will remain practically independent of voltage and frequency. The basic design equations may be used and applied to almost any magnetic material. The constants a and k for other magnetic materials can be obtained from graphs drawn from measurements made in accordance with the method mentioned in the

FIG. 5.22

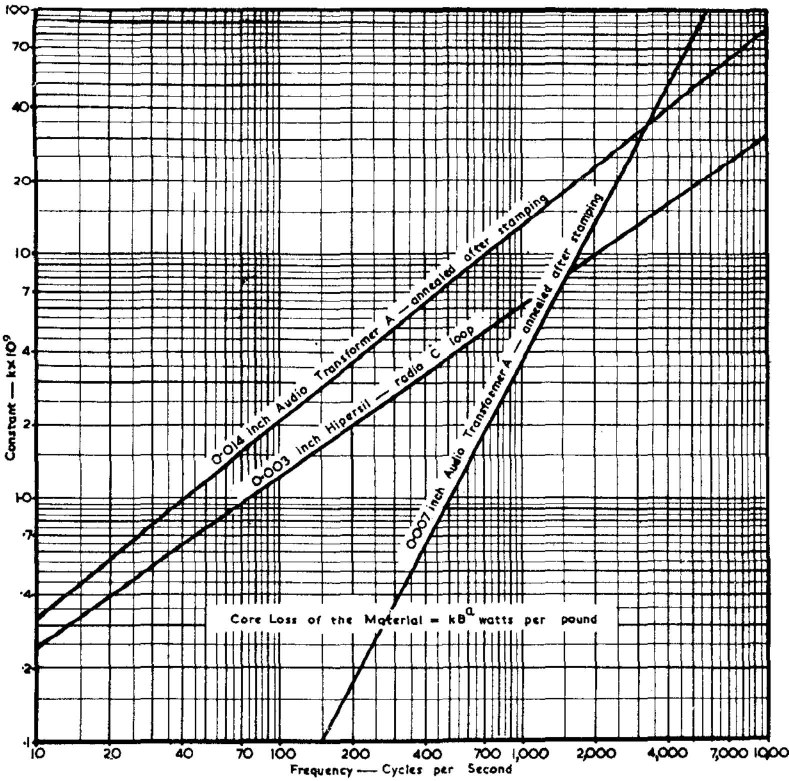


Fig. 5.22. *Experimentally determined curve showing values of " k " as a function of frequency for several magnetic materials (Ref. E2).*

original reference. Further information on iron-cored inductor design is contained in Refs. E3, E4, E5, E6, E7, E8, E9.

(vi) Design with d.c. flux

(Ref. E10).

When a d.c. flux is set up in the core by unbalanced direct currents in the windings, the effective permeability is decreased, as is shown in the curves of Fig. 5.20.

The amount of d.c. flux set up in the core depends on the applied d.c. magnetizing force (i.e. the total unbalanced ampere-turns in the windings), on the length of the magnetic circuit, and on the length of any air gap in the magnetic circuit.

The effective permeability depends on

(i) the d.c. flux density in the core

(ii) the a.c. flux density in the core

and (iii) the length of air gap in the magnetic circuit.

There is an optimum air gap giving a maximum value of the effective permeability for any particular value of total d.c. magnetizing force and a.c. flux density (Ref. E11).

For very low values of a.c. flux density, the following table gives the variation of incremental permeability with d.c. magnetizing force (A.T./inch of total length of magnetic circuit) and the ratio of the length of air gap to the total length of the magnetic circuit.

Incremental Permeability of High Silicon Steel
Total d.c. magnetizing force

Gap Ratio	1.0 AT/in.	2.0 AT/in.	5 AT/in.	10 AT/in.	20 AT/in.	30 AT/in.
0	1000	820	490	340	250	140
0.0005	720	700	680	560	360	170
0.0010	530	520	510	490	410	250
0.0015	—	425	410	400	370	270

Such data may be obtained using the methods given in " Magnetic Circuits and Transformers " (Ref. A3) Ch. 7, which sets out the general methods suitable for calculation of inductance with d.c. flux in the core.

The method developed by Crowhurst (Ref. E18) makes a considerable saving in time by the use of charts, particularly when a number of inductors are to be designed.

However, when the a.c. flux density is very small, as for example in many filter chokes, the method developed by Hanna may be used.

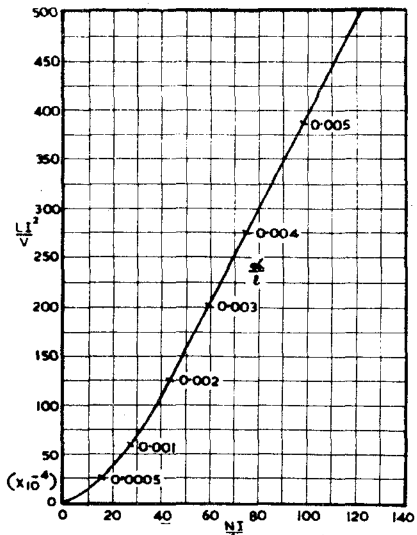


Fig. 5.23. LI^2/V plotted against NI/l with various gap ratio intercepts (Ref. E13).

(vii) Design by Hanna's Method

(Ref. E12)

Strictly, the method is applicable only when the core to be used is of constant cross section throughout the length of the magnetic path, but it gives a useful first estimate in cases where the cross section is not uniform.

Let N = number of turns in coil
I = direct current through coil
L = Inductance in henrys (at low a.c. flux density)
l = length of magnetic gap (inches)
α = length of air gap (inches)
α/l = air-gap ratio
a = cross-sectional area of core (square inches)
and V = $l.a$ = Volume of core (cubic inches).

Then, assuming in a typical case that L is 12 henrys and I is 80 mA proceed thus—

(a)	choose lamination size and stack	e.g. EI-11 $\frac{7}{8}$ in.
(b)	find V from Fig. 5.18C	4.03 cub. in.
(c)	calculate LI^2/V	190×10^{-4}
(d)	find NI/l from Fig. 5.23	58
(e)	find l from Fig. 5.18C	5.26 in.
(*f*)	calculate N	3800 turns
(g)	assume suitable wire gauge	33 A.W.G.
(h)	find turns per layer from winding data chart	118
(i)	calculate number of layers	33
(j)	calculate coil build using 0.0015 in. insulation	0.31 in.
(k)	calculate winding build using 0.05 in. former	0.36 in.
(l)	find maximum build from Fig. 5.18C so that winding will fit window satisfactorily	0.367 in.
(m)	calculate mean length of turn from Fig. 5.18D	4.8 ins.
(n)	calculate coil resistance	315 ohms
(o)	calculate power lost in coil	2 watts
(p)	calculate cooling area (see Sect. 5)	14.3 sq. ins
(q)	calculate dissipation in watts/sq. in.	0.14 watts/sq. in.
(r)	find temperature rise from Fig. 5.18B	20°C.
(s)	find gap ratio α/l from Fig. 5.23	0.0029
(t)	calculate air gap $(\alpha/l)\times l$	0.0075 in.

With any particular core, the highest inductance is obtained with the largest possible number of turns, as limited by the window space available, the permissible value of d.c. resistance, and heating of the winding with the specified current.

(viii) Design of inductors for choke-input filters

(Ref. E13)

(a) The input choke

Chapter 30 Sect. 3 Eqn 1, 2 or 3 gives the minimum value required for the inductance of the input choke of a choke-input filter for any particular values of load resistance and frequency.

The required inductance for any particular value of load resistance may be achieved by designing the choke by Hanna's method as given above. It is only necessary to check that the value of the a.c. flux density in the core is small, so that Hanna's method will be applicable. For a full-wave rectifier the a.c. voltage applied to the choke is of twice the supply frequency and its peak amplitude is approximately two-thirds of the direct voltage at the input to the filter.

If, however, good regulation is required over a wide range of loads, it is necessary for the inductance of the input choke to be sufficiently high at the highest value of load resistance, that is, at the minimum value of load current. Some minimum load current greater than zero must be provided because the inductance would need to be infinite to maintain regulation down to zero load.

For a wide range of load variation (say 10 : 1) it would be very uneconomical to design a choke to have a constant inductance at all load currents as large as that required at *minimum* load. The choke is therefore designed so that its inductance will vary with the direct current through it, in such a way that the inductance is sufficiently high at all values of the load current. This method is discussed fully by Dunham (Ref. E14) and Crowhurst (Ref. E19) and reference should be made to their papers for detailed design considerations.

For many purposes, the following simple design procedure is adequate.

(1) From Eqn. (1) (2) or (3) of Chapter 30 Sect. 3, determine the required values of the choke inductance at minimum and maximum d.c. load currents.

(2) Select an interleaved core and, assuming a permeability of (say) 1000, design a choke to give the inductance required at *minimum d.c. load current*. For this purpose use the procedure of Sect. 6(iv) above but correct the assumed value of permeability using Fig. 5.20 and taking into account the number of d.c. ampere-turns

per inch of core path length. The peak a.c. voltage, of double mains frequency, across the choke, may be taken to be about two thirds of the d.c. voltage at the input to the filter.

(3) Using Fig. 5.20 note the change in permeability μ when maximum d.c. load current flows through the choke and check that the value of inductance under these conditions is sufficient. If it is not sufficient an increase in turns may be satisfactory and this can be checked by repeating the procedure above. If this is not satisfactory a small air gap may be necessary, but calculations then become complex.

A choke so designed is commonly referred to as a **swinging choke** because its inductance varies with the direct current through it.

(b) **The second choke**

The design of the second choke of a choke input filter is straightforward, and Hanna's method may be used after determining the required value of inductance by the methods set out in Chapter 31 Sect. 1.

(ix) Measurements

Measurements of inductances of iron-cored coils must be made under conditions similar to those under which the coils are to be used, because the value of inductance depends to a marked degree on the a.c. flux density in the core and also on any d.c. flux set up in the core by direct currents in the windings.

Bridge methods are desirable for accurate measurements of inductance but must be arranged to simulate the operating conditions of the coil being measured. Owen and Hay bridges are widely used when there is a large flux due to direct current (Refs. E15, E16).

For many purposes it is satisfactory to determine the effective inductance (or more accurately, the impedance) of a coil by measuring the current through the coil when the rated a.c. voltage is applied to it ; but this method is not usually feasible when a d.c. component is also present.

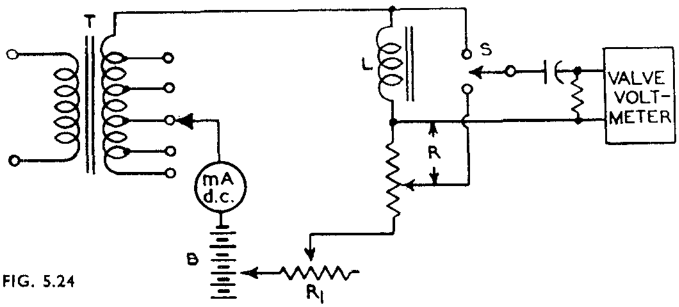


FIG. 5.24

Fig. 5.24. Determination of the impedance of an inductor L carrying direct current.

When a d.c. flux must be produced in the core to simulate operating conditions, the circuit arrangement of Fig. 5.24 may be used to determine the a.c. impedance of the coil. A valve voltmeter is used to adjust the a.c. voltage drop across the known variable resistance R to equality with the a.c. voltage drop across the inductance L. The required value of direct current through the coil is obtained by adjusting the tapping on battery B and the rheostat R_1 (or by adjusting any other variable direct current source), the current being measured by a d.c. ammeter or milliammeter. The required alternating voltage across the choke coil is then adjusted by varying the tapping on transformer T, the voltage across the coil being read on the valve voltmeter ; this adjustment is not usually critical. The resistance R is then varied until the same reading is obtained on the valve voltmeter when it is connected across R

by operating the switch S. If the valve voltmeter responds to direct voltages, it must be connected to the circuit through a blocking condenser and grid leak as shown in the diagram.

The value Z obtained for the impedance of the coil will differ from its reactance owing to coil losses, but for most purposes it will be satisfactory to assume that it is equal to the reactance. If the frequency used is f, the value of inductance is given approximately by

$$L \approx Z/2\pi f.$$

or

$$L \approx Z/314 \text{ Henrys, for } f = 50 \text{ c/s.}$$
$$\approx Z/376 \text{ Henrys for } f = 60 \text{ c/s.}$$

where Z is numerically equal to the value of R for balance.

For choke coils carrying direct current, and operating with a high a.c. flux density, reference should be made to the method of approximate measurement given by F. E. Terman (Ref. E15) Fig. 40 and pp 57-58.

(x) Iron-cored inductors in resonant circuits

The performance of iron-cored inductors in resonant circuits cannot be calculated mathematically owing to the immense complexity. While most work is carried out empirically, it is helpful to have a general grasp of the problem, and this may perhaps best be carried out graphically (see Ref. E17).

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