SECTION 7

TRANSFORMERS

SECTION 7

CHAPTER 1

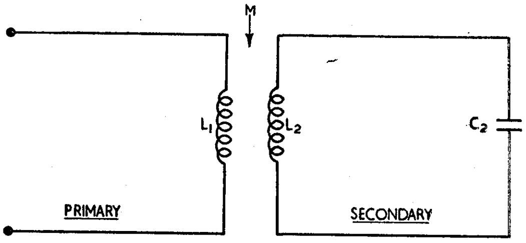
COUPLED CIRCUITS

PART 1, SECTION 7, CHAPTER 1

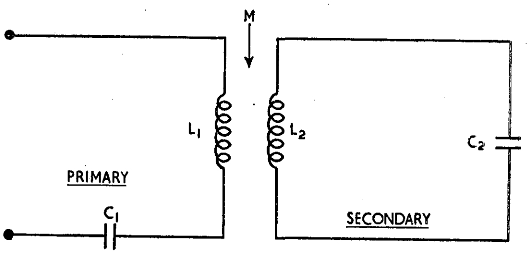
COUPLED CIRCUITS

Introduction

1. Two a.c. circuits are said to be *coupled* when they are so linked that energy is transferred from one circuit to the other. For example, when mutual inductance exists between coils that are in separate circuits, these circuits are 'inductively coupled'. The effect of the mutual inductance is to make possible the transfer of energy from one circuit to the other by *transformer* action. That is, an alternating current established in the first or *primary* circuit produces magnetic flux which is linked with, and induces a voltage in, the coupled or *secondary* circuit. This does not, of course, apply to d.c. circuits since the flux must be *changing* for electromagnetic induction to occur. Two examples of inductively coupled circuits commonly encountered in radio equipments are shown in Fig. 1.



(a) TUNED SECONDARY



(b) TUNED PRIMARY, TUNED SECONDARY

Fig. 1—INDUCTIVELY COUPLED CIRCUITS

Coupling Coefficient

2. It was shown in Sect. 2, Chap. 2 that the proportion of the flux from one circuit which is linked with another determines the extent of the coupling between them. The greatest possible mutual linkage between two circuits occurs when *all the flux* from each embraces every turn of the other. With two coils of self-inductance L_1 and L_2 having mutual inductance M:—

$$M = k\sqrt{L_1 L_2}$$

$$\therefore k = \frac{M}{\sqrt{L_1 L_2}} \quad \cdot\cdot \quad (1)$$

The ratio $\frac{M}{\sqrt{L_1 L_2}} = k$ is termed the *coupling coefficient* between two mutual inductively coupled circuits. It is always less than unity since when $k = 1$, $M = \sqrt{L_1 L_2}$ and there is no leakage of flux between the two circuits ; this is a condition which cannot exist in practice.

Mutual Inductive Coupling

3. Fig. 2 shows two series resonant circuits in which the inductances L_1 and L_2 are placed sufficiently close together to be coupled via the mutual inductance M between them. The values of L and C in each circuit are such that both circuits have the same resonant frequency.

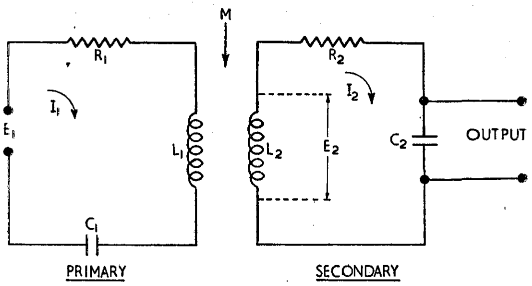


Fig. 2—TWO SERIES TUNED CIRCUITS COUPLED BY MUTUAL INDUCTANCE

4. With the secondary on open circuit, the primary current is dependent entirely on the primary impedance, i.e., $I_1 = \frac{E_1}{Z_1}$. With the secondary on load the effect of the secondary current is to modify the primary circuit by the addition of a 'reflected impedance'. Thus when an e.m.f. of r.m.s. value E_1 is applied to the primary circuit an alternating current of r.m.s. value I_1 will be

established and will set up an alternating magnetic flux around the coil L_1. Some of this flux links with the turns of L_2 and induces an e.m.f. of r.m.s. value E_2 in it, so that a current $I_2 = \frac{E_2}{Z_2}$ is established in the secondary when it is on load. The secondary current in turn induces a voltage back into the primary in such a direction as to tend to oppose the primary voltage (Lenz's law). The two circuits thus interact so as to affect each other to an extent depending on the coupling between them.

Reflected Impedance

5. The current in the primary circuit is not the same as it would be in the absence of the secondary current. The effect of the presence of the secondary is as though an impedance $\frac{(\omega M)^2}{Z_2}$ had been added in series with the primary, where $\omega = 2\pi f$, M = mutual inductance, and Z_2 = series impedance of the secondary circuit considered by itself $= \sqrt{R_2^2 + X_2^2}$. The equivalent impedance $\frac{(\omega M)^2}{Z_2}$ is termed the *reflected impedance*, and since Z_2 contains resistance R_2 and reactance X_2 so also does the reflected impedance. The effect of the reflected impedance is:—

(*a*) To *increase* the primary *resistance*.

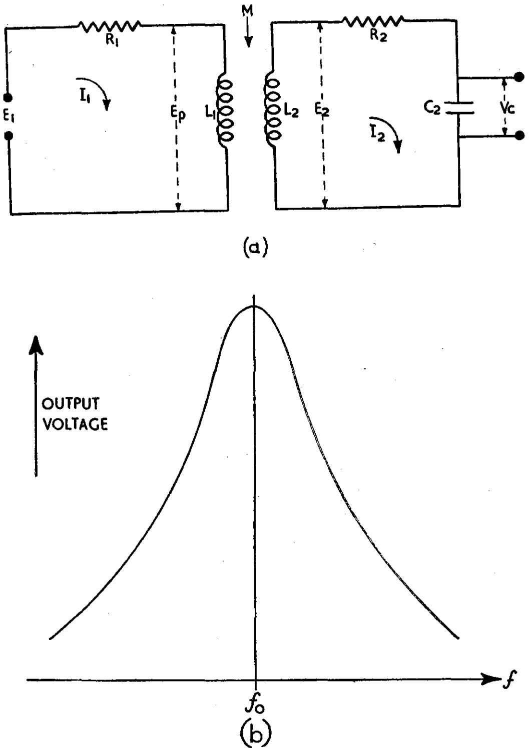
(*b*) To *reduce* the primary *reactance*.

6. When M is small or Z_2 is large the reflected impedance $\frac{\omega^2 M^2}{Z_2}$ will be small and the primary circuit will be virtually unaffected by the secondary. When M is large or Z_2 is small, the reflected impedance will be large, and the secondary circuit will affect the primary to a considerable extent. This is especially the case when the secondary is at resonance to the applied frequency for then $Z_2 = R_2$ and $\frac{\omega^2 M^2}{Z_2}$ is a maximum.

7. Off resonance, the effect of the reflected impedance is to increase the primary resistance and to reduce the primary reactance. When both circuits are resonant with the applied frequency, the reflected impedance is purely resistive and its effect is to increase the primary resistance only.

Untuned Primary and Tuned Secondary

8. This is a type of circuit often used in the initial (r.f.) stages in receivers and is as shown in Fig. 3(a). Transformer action results in the primary current inducing a voltage in the secondary, the magnitude of which depends on the coupling coefficient. The resulting secondary current will depend on the secondary impedance and is given by $I_2 = \frac{E_2}{Z_2}$. It varies as in a normal series tuned circuit, so that the variation of output voltage ($V_c = I_2 X_{c_2}$) as the frequency varies about resonance will be as shown in Fig. 3(b).



(a)

(b)

Fig. 3—R.F. TRANSFORMER—UNTUNED PRIMARY, TUNED SECONDARY

9. In practice, the input of E_1 volts to the r.f. transformer of Fig. 3(a) would produce an output voltage V_c of about the same value. This is because the high internal resistance of the supply makes the p.d. across L_1 much less than the e.m.f. E_1. The secondary voltage E_2 would be smaller still (since loose coupling is usual in r.f. transformers) and the Q of the secondary would

cause the output voltage to be of the same order as E_1. The output voltage can be made larger than the supply voltage by tuning the primary as well as the secondary.

Tuned Primary and Tuned Secondary

10. This type of circuit (Fig. 4) is used extensively in radio equipments. When two resonant circuits having equal Q values are tuned to the same frequency and coupled together, the resulting behaviour depends very largely upon the value of the co-efficient of coupling k.

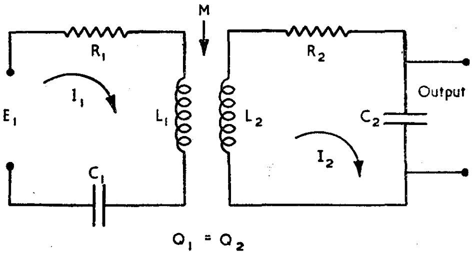


Fig. 4—R.F. TRANSFORMER—TUNED PRIMARY, TUNED SECONDARY

11. **Loose coupling.** This is the term used to denote a low value of k, when there is little interaction between the two circuits. As the frequency of the applied voltage is varied through the resonant point of the two circuits, the current in the primary increases to a maximum and falls off again according to the normal series resonance curve. The voltage induced in the secondary by this primary current will vary in the same manner and the secondary current will vary even more sharply, since the selectivity of the secondary is proportional to the *product* of the Q values of primary and secondary. Its peak value is however small because of the low value of k. These points are illustrated in Fig. 5. Loose coupling is usual in r.f. voltage transformers used in the initial stages of receivers in order to improve the selectivity of such circuits.

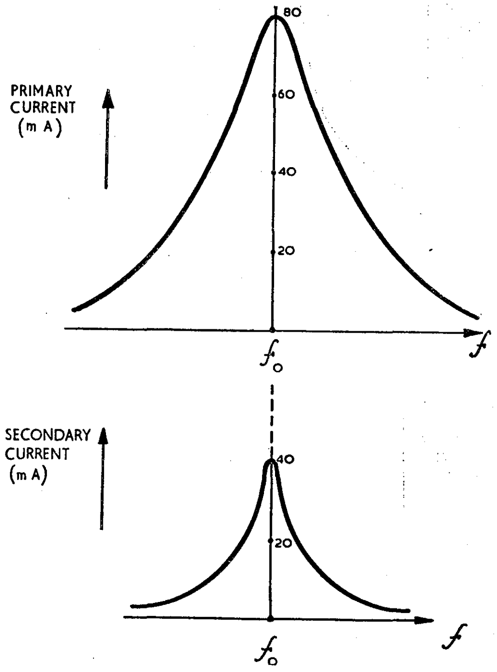


Fig. 5—LOOSE COUPLING

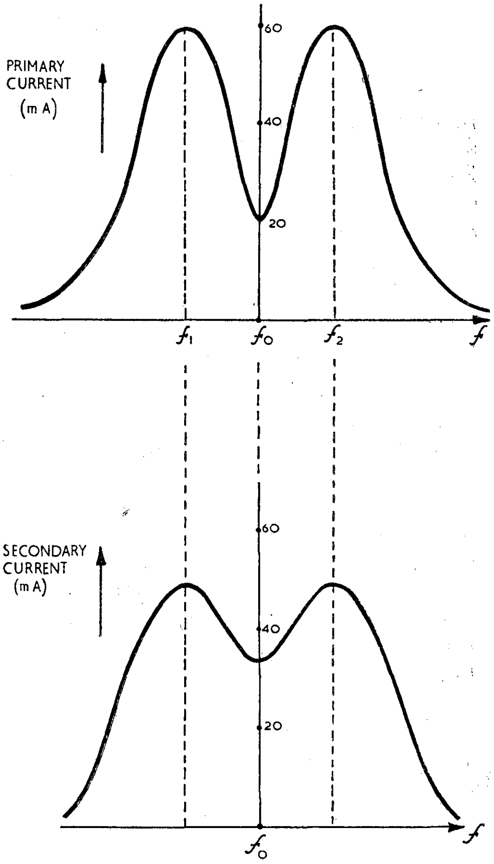


Fig. 6—TIGHT COUPLING

12. **Tight coupling.** This is the term used to denote a high value of k, with appreciable interaction between the circuits. The variations in primary and secondary currents

(*b*) *Critical k.* This condition is shown by the curve marked (c) in Fig. 9. Two peaks are now evident in the primary curve, while the secondary curve still shows a single maximum, the greatest possible.

(*c*) *Large k.* The primary and secondary response curves for conditions where k is greater than the critical value are shown by curves (d) and (e) of Fig. 9. Double peaks appear in both primary and secondary curves, though the 'dip' between them is much more marked in the primary for reasons given in Para. 12(*d*).

Frequencies of Peaks

15. Where the value of k is greater than the critical value, double-humping occurs in both primary and secondary response curves. The peaks occur at a *virtual* resonant frequency when the primary reactance is zero. The positions of the two virtual resonant peaks can be found from the equation:—

$$f = \frac{f_0}{\sqrt{1 \pm k}} \qquad \cdot\cdot \qquad \cdot\cdot \quad (3)$$

where f_0 is the normal resonant frequency, f is the frequency of maximum primary response, and k is the coupling coefficient. The frequencies at which peaks occur in the primary circuit are, therefore:—

$$f_1 = \frac{f_0}{\sqrt{1 + k}} \text{ (below resonance)}$$

$$f_2 = \frac{f_0}{\sqrt{1 - k}} \text{ (above resonance)}$$

Bandwidth

16. If the bandwidth of a coupled circuit is considered as the spacing between the frequencies at which the peaks occur in the primary circuit:—

$$\text{Bandwidth} = f_2 - f_1$$

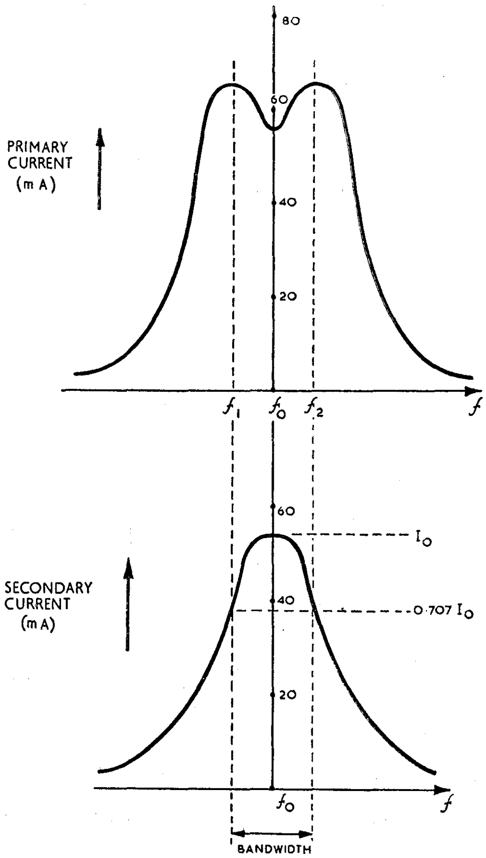
$$= \frac{f_0}{\sqrt{1 - k}} - \frac{f_0}{\sqrt{1 + k}}$$

$$\therefore \frac{\text{Bandwidth}}{f_0} = \frac{1}{\sqrt{1 - k}} - \frac{1}{\sqrt{1 + k}}$$

And if k is small:—

$$\frac{\text{Bandwidth}}{f_0} = k$$

$$\therefore \text{Bandwidth} = kf_0 \qquad \cdot\cdot \qquad \cdot\cdot \quad (4)$$



HALF POWER BANDWIDTH = $f_2 - f_1$ AT CRITICAL COUPLING.

Fig. 10—k, Q, AND BANDWITH

At critical coupling for identical circuits:—

$$k = \frac{1}{Q}$$

$$\therefore \text{Bandwidth} = \frac{f_0}{Q} \qquad \cdot\cdot \qquad \cdot\cdot \quad (5)$$

This is the half-power bandwidth as defined in Sect. 5, Chap. 2 and is as shown in Fig. 10.

17. **Wide-band r.f. transformers.** For coupled circuits, having identical primary and secondary circuits with Q values of 100, the critical value of k is 0·01. From equation (4), it is seen that the bandwidth is then 1 per cent of the resonant frequency. In some cases, bandwidths of up to 10 per cent of the resonant frequency are necessary. For example, a circuit resonant at 50 Mc/s but giving full response at frequencies from 47·5 Mc/s to 52·5 Mc/s

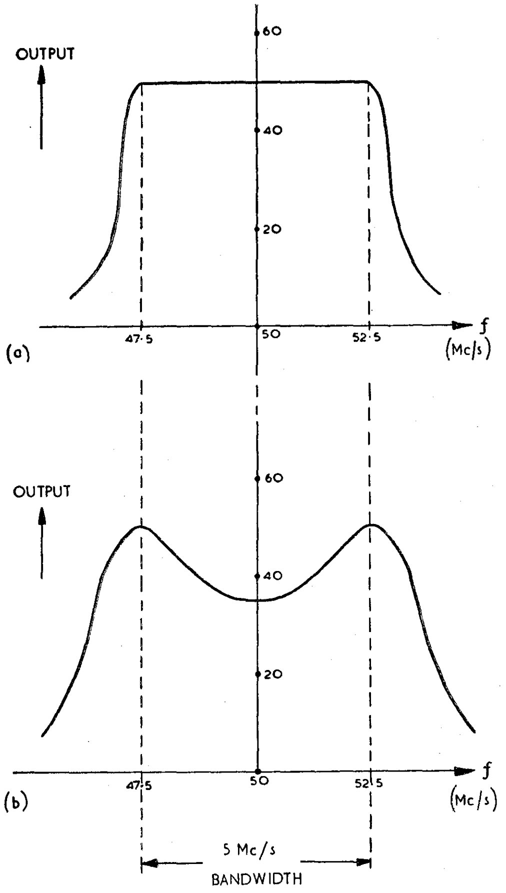


Fig. 11—WIDEBAND R.F. TRANSFORMER

(a bandwidth of 5 Mc/s) may be required as shown in Fig. 11(a). By increasing the value of k beyond the critical value it is possible to produce a circuit having a bandwidth of 5 Mc/s, but the response falls off rapidly between these points (Fig. 11(b)). The use of damping resistors across the primary and secondary will produce a flat top, but the overall response is considerably reduced as shown in Fig. 12(a). The method of obtaining a response of the kind illustrated in Fig. 11(a) is to combine the response curves of several pairs of coupled circuits. Each pair is coupled tightly enough to give two resonant peaks about 1 Mc/s apart. Slight damping may also be included to give a reasonably flat-topped response. If five such circuits are tuned to 48, 49, 50, 51 and 52 Mc/s respectively, the overall response curve will be reasonably flat over the bandwidth required, as shown in Fig. 12(b).

R.F. Power Transformers

18. The r.f. transformers considered up to the present have been concerned with the transfer of power at a *low level* over a desired band of frequencies. Such transformers are termed r.f. *voltage* transformers since the main consideration is the voltage developed across the output terminals. R.F. *power* transformers are normally used to transfer r.f. power at a *high level* from one part of a circuit to another. A typical use is in the transfer of power from a transmitter to an aerial system, as shown in Fig. 13. Maximum power is delivered to the aerial system when the impedance of this load is correctly 'matched' to the output impedance of the transmitter. Correct matching is achieved by:—

(*a*) Varying the coupling (Fig. 13(a)).

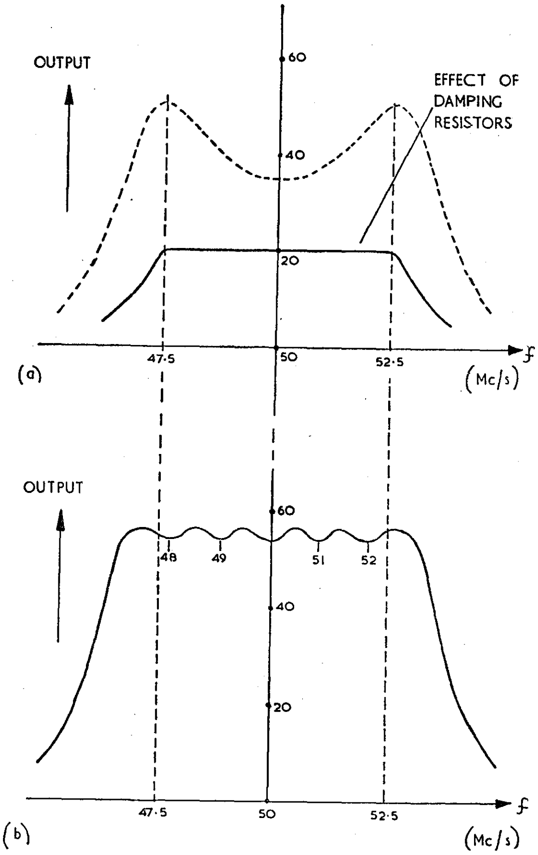


Fig. 12—METHODS FOR INCREASING THE BANDWIDTH

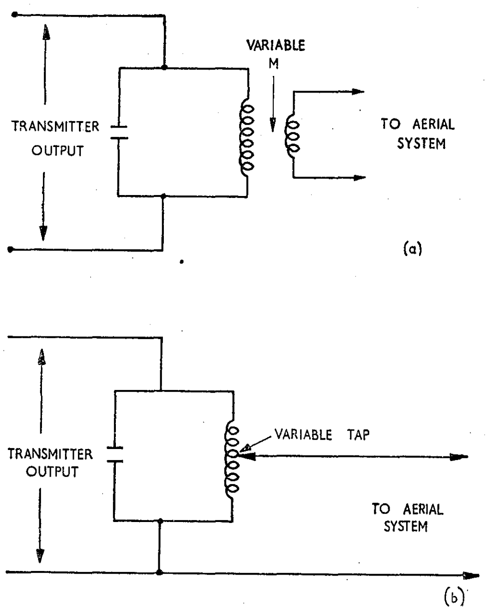
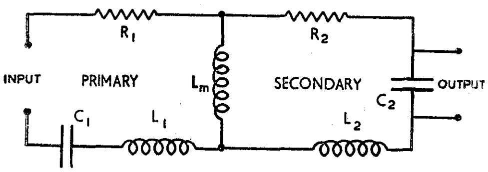


Fig. 13—R.F. POWER TRANSFORMERS

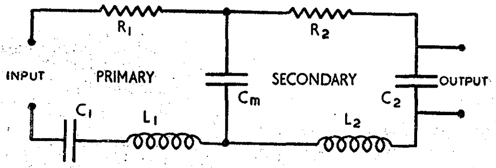
(*b*) Using an auto-transformer as in Fig. 13(b). (See Chap. 2 on auto-transformers).

Other Forms of Coupling

19. Energy can be transferred from one circuit to another by a variety of coupling methods in addition to the mutual inductive coupling method just considered. One method is to use an impedance which provides a current path *common* to both circuits (Fig. 14). The common impedance is *in parallel* with the secondary and the voltage developed across this impedance by the primary current acts as the applied e.m.f. to the secondary. Another form of coupling which is often encountered is 'top-end capacitive coupling' (Fig. 15). The two circuits are joined by a coupling capacitance that is *in series* with the secondary across the primary. A portion of the primary voltage depending on the value of the coupling capacitor, is then developed across the secondary.



(a) COMMON INDUCTIVE COUPLING



(b) COMMON CAPACITIVE COUPLING

Fig. 14—COMMON IMPEDANCE COUPLING

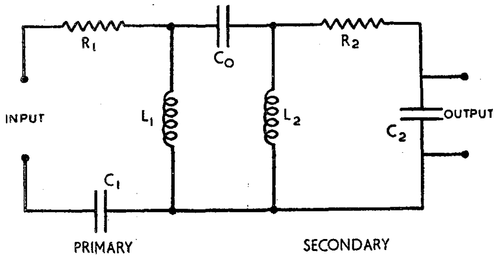


Fig. 15—TOP-END CAPACITIVE COUPLING

20. In top-end capacitive coupling, the primary and secondary inductance values can be varied by iron dust cores to alter the resonant frequency of the coupled circuits, and the two coils are normally placed at right angles to each other to avoid mutual inductive coupling (Fig. 16). The main practical difference between inductive and capacitive coupling is that in the latter the coupling is almost wholly capacitive and

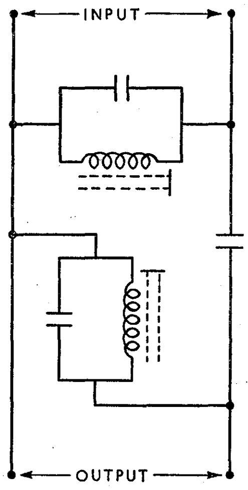


Fig. 16—PRACTICAL ARRANGEMENT OF TOP-END CAPACITIVELY COUPLED CIRCUITS

can be adjusted within fine limits, whereas in the former, unless an electric screen is provided between primary and secondary, both inductive and capacitive coupling are present and adjustment is difficult.

21. Other forms of coupling employ a combination of the methods described, but the behaviour of all coupled circuits follows the same general character as that discussed for mutual inductive coupling.

Screening

22. It is normal to confine the electric and magnetic fields produced by a r.f. transformer by means of a screening can, in order to prevent mutual interference between such fields at other parts of the circuit. The screening can, which is normally fixed to the chassis, gives combined electric and magnetic shielding. The electric lines of force terminate on the earthed can so that no electric field from the r.f. transformer exists outside the can. At the same time, eddy currents are induced in the can by the alternating magnetic field. These eddy currents produce a subsidiary magnetic field which is in such a direction as to neutralize the main field outside the can, giving magnetic screening. The dimensions of the screening can must be such as to alter the constants of the r.f. transformer as little as possible.

SECTION 7

CHAPTER 2

IRON-CORED TRANSFORMERS

Paragraph

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PART 1, SECTION 7, CHAPTER 2

IRON-CORED TRANSFORMERS

Introduction

1. A transformer consists essentially of an arrangement in which two coils are magnetically coupled to one another. If a *varying* current is passed through one coil, known as the primary, the changing magnetic flux linking with the second coil, called the secondary, induces a voltage therein. Thus, the mutually coupled circuits considered in Chap. 1 are transformers. At power and audio frequencies *iron cores* are generally used and the iron-cored transformer is then a special case of a mutually coupled circuit. The principles of iron-cored and air-cored transformers are the same, but it is convenient to treat them in a different manner for reasons which will become obvious in later paragraphs.

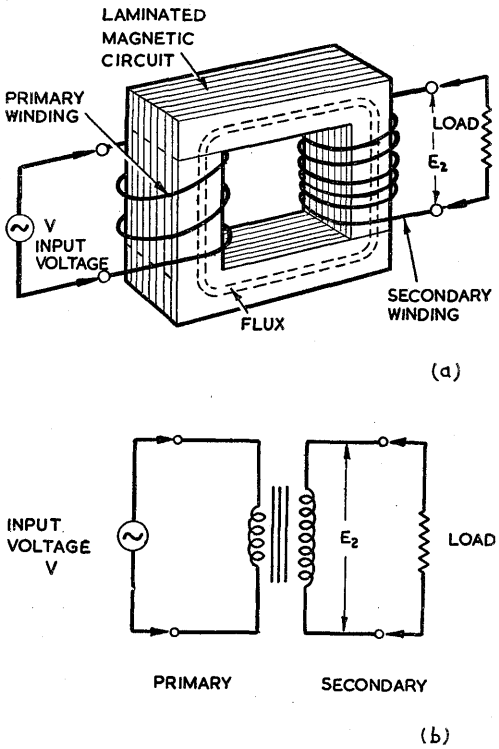


Fig. 1—IRON-CORED TRANSFORMER

2. An iron-cored transformer consists of two insulated coils wound separately over a closed magnetic circuit of low reluctance (Fig. 1). An alternating current generator is connected across the primary and the load is connected to the secondary. The alternating voltage applied to the primary establishes a current in this winding and an alternating magnetic flux is set up in the core. Most of the flux links with the secondary winding to induce a voltage in the secondary and this is available at the secondary terminals. If these terminals are closed by a load circuit, a secondary current is established and energy is expended in the load. The power transmitted to the load in the secondary must in the first place, *be drawn from the generator* by the primary circuit.

3. The effects of using iron are considerable:—

(*a*) The reluctance of the magnetic circuit is low and its permeability high, so that the magnetic flux density and the inductance of each coil are greatly increased.

(*b*) Nearly all the flux from the primary links with the secondary because of the low reluctance of the iron path in relation to that of air. Thus, the coupling coefficient k is very nearly equal to unity and for practical purposes can be considered so. Therefore, the mutual inductance M is large ($M = k\sqrt{L_1 L_2}$).

(*c*) Hysteresis and eddy current losses are introduced.

Magnetising Current

4. Fig. 2(a) shows the circuit of an iron-cored transformer in which the primary circuit resistance is so small as to be neglected, the primary inductance is L_1 and the secondary is open-circuited (off load). The primary circuit is not affected by the presence of the secondary since no current can be established in the secondary on open circuit. Thus, the current in the primary is dependent only on the applied voltage V and on the primary impedance. This current is termed the *magnetising current* I_M and is given by:—

$$I_M = \frac{V}{X_{L1}}$$

$$\therefore\ I_M = \frac{V}{\omega L_1} \quad \ldots \quad \ldots \quad (1)$$

The primary impedance is here considered to be a pure inductance so that I_M *lags* V by 90°. The flux Φ is proportional to the magnetising current and is *in phase* with it, so that Φ lags V by 90°. The phase relationship between V, I_M and Φ is as shown in Fig. 2(b).

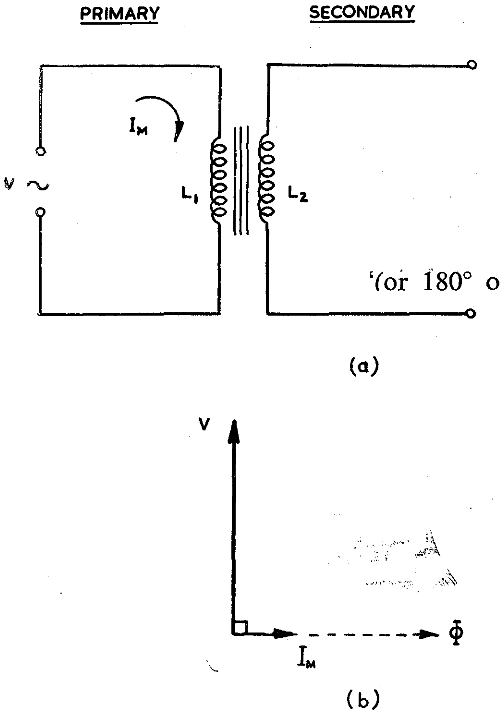


Fig. 2—MAGNETISING CURRENT

5. The primary inductance of a transformer is normally high because of the presence of the iron core. The magnetising current is, therefore, normally small. If however, the frequency of the applied voltage is *lower* than that for which the transformer is designed, the magnetising current $I_M = \frac{V}{\omega L_1}$ may become excessive and may cause damage to the primary winding. A transformer can be used at a frequency *higher* than that for which it is designed without any such danger.

Transformation Ratio

6. Fig. 3(a) shows a transformer with N_1 turns on the primary winding and N_2 on the secondary. The transformer is off load. When an alternating voltage V is applied to the primary a magnetising current I_M is established. The flux resulting from this current is changing at the rate $\frac{d\varnothing}{dt}$ webers per second, and since this flux is the same for primary and secondary windings (assuming k = 1), the primary induced voltage is,

$$E_1 = -N_1 \frac{d\varnothing}{dt} \text{ (volts)},$$

and the secondary induced voltage is,

$$E_2 = -N_2 \frac{d\varnothing}{dt} \text{ (volts)}.$$

The two voltages E and E [illegible] '(or 180° out of phase depending on the winding directions) since the same flux is cutting both windings.

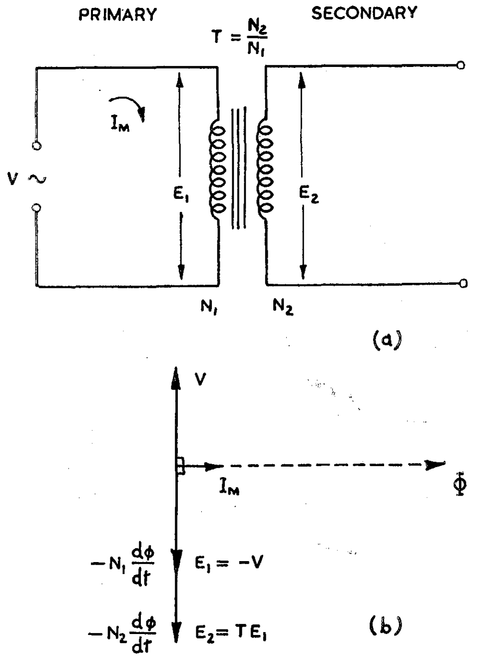


Fig. 3—TRANSFORMATION RATIO

The *magnitude* of the primary and secondary induced voltages depends, however, on the primary turns N_1 and the secondary turns N_2 respectively. Further, E_1 and E_2 are 180° out of phase with the applied voltage V, since E_1 is the *back e.m.f.* in the primary and is equal and opposite to V. The vector diagram showing these relationships is given in Fig. 3(b). Now:—

$$\frac{E_2}{E_1} = \frac{-N_2\, d\varnothing/dt}{-N_1\, d\varnothing/dt}$$

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

This relationship gives the transformation ratio T:—

$$T = \frac{N_2}{N_1} = \frac{E_2}{E_1} \quad .. \quad .. \quad (2)$$

7. The derivation of the name 'transformer' will now be clear; by choosing a suitable ratio for $\frac{N_2}{N_1}$ an alternating voltage can be transformed to any other required voltage of the same frequency. The transformation can be either *step-up* or *step-down.* A step-up transformer is one where T is greater than 1; a step-down transformer has T less than 1. A step-up ratio of 2 : 1 means that the secondary winding has twice as many turns as the primary and T = 2. A step-down ratio of 2 : 1 means that the secondary winding has half the turns of the primary and T = ½. An r.m.s. voltage of 230 V applied to each transformer in turn would give outputs of 460 V and 115 V r.m.s. respectively.

Transformer With More Than Two Windings

8. A transformer with three secondary windings is illustrated in Fig. 4. The same theory applies as for transformers with one primary and one secondary winding. Thus, the secondary voltages are proportional to the transformation ratio between the respective secondary winding and the primary.

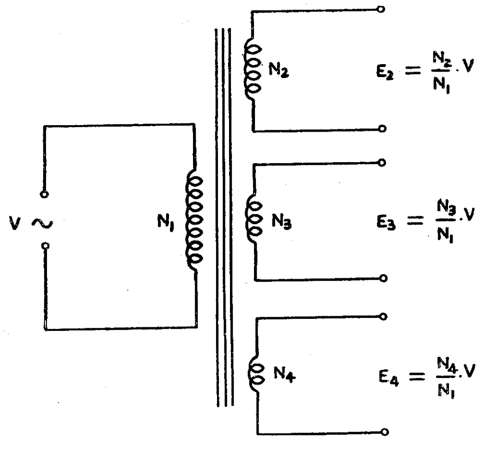


Fig. 4—TRANSFORMER WITH THREE SECONDARY WINDINGS

Hence:—

$$E_2 = T_2 V = \frac{N_2}{N_1} . V$$

$$E_3 = T_3 V = \frac{N_3}{N_1} . V$$

$$E_4 = T_4 V = \frac{N_4}{N_1} . V$$

A typical transformer of this type has V = 230 V, E_2 = 400 V, E_3 = 200 V, and E_4 = 6 V (all r.m.s. values).

Primary No-load Current

9. So far, the only current considered has been the magnetising current I_M. This lags 90° on the applied voltage when the primary circuit is assumed to be a pure inductance.

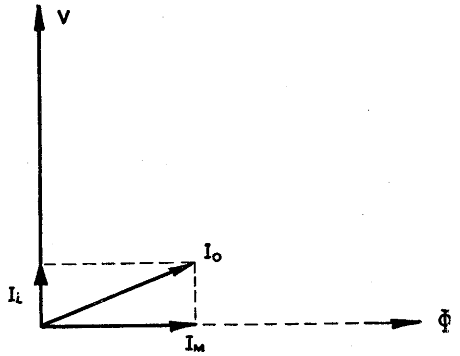


Fig. 5—PRIMARY NO-LOAD CURRENT

However, even when there is no secondary load current, power is being drawn from the supply because of hysteresis and eddy current losses in the iron core. The total primary no-load current is therefore made up of two components:—

(*a*) A magnetising current I_M which establishes magnetic flux in the core and which *lags* 90° on V.

(*b*) A loss component I_i representing hysteresis, eddy current and other power losses, which is *in phase* with V.

The primary no-load current I_o is then the vector resultant of I_M and I_i and lags in phase by *less* than 90° on V. Fig. 5 shows the vector diagram.

Transformer on Load

10. If the applied voltage V is constant, then the flux Φ is constant whatever load is connected to the secondary. Assuming no resistance in the transformer, the applied

voltage must be exactly equal and opposite to the back e.m.f. developed across the primary. Hence, if V is constant the back e.m.f. must be constant. But the back e.m.f. is proportional to the rate of change of flux, so that the flux must be such that its rate of change is a sine wave of constant amplitude. Thus, *the flux itself must be a sine wave of constant amplitude irrespective of the load on the secondary.*

11. When a load is connected to the transformer, a current $I_2 = \frac{E_2}{Z_2}$ is established in the secondary. This current will set up a magnetic flux of its own and since, from Para. 10, the magnetising flux must remain *constant* if the applied voltage is constant, some action must take place to nullify the flux due to I_2. In fact, a primary current I_1 flows. If I_1 is to have the opposite effect to I_2 it must be 180° out of phase. Its magnitude is determined by the fact that the two effects are to be equal. Now, the secondary flux Φ_2 is proportional to $I_2 N_2$. The flux produced by I_1 must equal this.

$$\therefore I_1 N_1 = I_2 N_2$$

$$\therefore \frac{I_1}{I_2} = \frac{N_2}{N_1} = T \quad .. \ (3)$$

$$\text{Thus, } T = \frac{E_2}{E_1} = \frac{I_1}{I_2} \quad .. \ (4)$$

From this it is seen that the *secondary* voltage E_2 is T times the primary voltage E_1, and the *primary* current I_1 is T times the secondary current I_2.

12. From equation (4) :—

$$\frac{E_2}{E_1} = \frac{I_1}{I_2}$$

$$\therefore E_2 I_2 = E_1 I_1$$

$$\therefore \text{Output power} = \text{Input power} \quad .. \ (5)$$

The perfect transformer introduces no loss and the efficiency is 100 per cent. In practice efficiencies of 99 per cent can be obtained.

Types of Load

13. **Pure resistance.** Fig. 6(a) shows a transformer with a purely resistive load connected to the secondary. The vector diagram is given in Fig. 6(b), and is constructed as follows:—

(*a*) The flux Φ is the reference vector since the flux is common to primary and secondary. The magnetising current I_M which produces this flux is in phase with Φ and lags 90° on the applied voltage V.

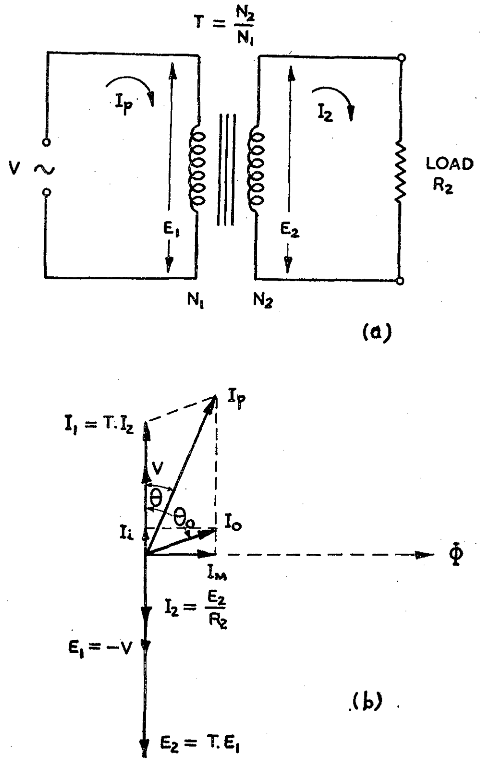


Fig. 6—TRANSFORMER WITH A RESISTIVE LOAD

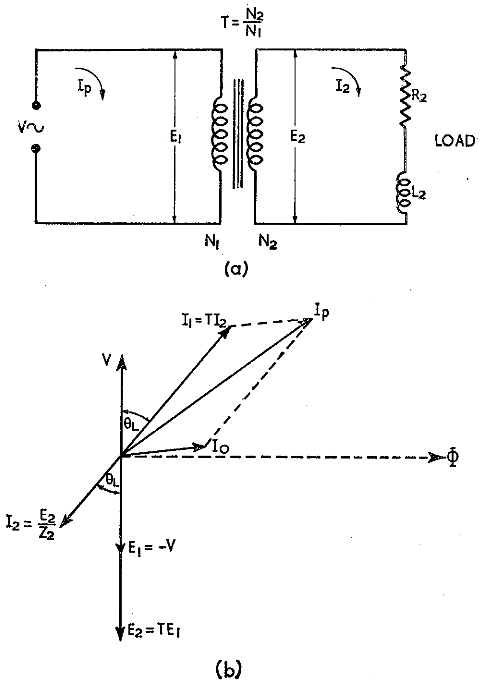
(*b*) A loss current I_i is established in phase with V, and the total no-load current I_o is the vector resultant of I_M and I_i as shown.

(*c*) The changing flux induces a back e.m.f. $E_1 = -V$ in the primary and an e.m.f. E_2 in the secondary. E_1 and E_2 are 180° out of phase with V, and $E_2 = T E_1 = -T V$.

(*d*) A secondary load current $I_2 = \frac{E_2}{R_2}$ is established *in phase* with E_2, since the load is a pure resistance.

(*e*) A primary current I_1 of magnitude $T I_2$ is established to produce a flux equal and opposite to that produced by I_2. I_1 is 180° out of phase with I_2.

(*f*) The total primary load current I_p is then the vector resultant of the no-load current I_o and the current I_1. I_p is seen to lag V by an angle θ which is less than the original angle θ . Thus, the phase difference between primary current and applied voltage is altered by the secondary load.



(a)

(b)

Fig. 7—TRANSFORMER WITH A RESISTIVE-INDUCTIVE LOAD

14. **Reactive load.** Fig. 7(a) shows a transformer with a resistive-inductive load connected to the secondary. The vector diagram is shown in Fig. 7(b) and is constructed in a manner similar to that given in Para. 13. However, the secondary current I_2 now *lags* E_2 by an angle θ_L and since I_1 is 180° out of phase with I_2, then I_1 lags V by the same angle θ_L. The resultant primary load current I_p is the vector resultant of I_o and I_1.

Impedance Transformation

15. Any impedance connected across the secondary will reflect an equivalent shunt impedance across the primary.

From equation (4):—

$$T = \frac{E_2}{E_1} = \frac{I_1}{I_2}$$

$$\therefore T^2 = \frac{E_2}{E_1} \times \frac{I_1}{I_2}$$

But $E_2 = I_2 Z_L$,

where Z_L = load impedance

And $E_1 = I_1 Z_p$,

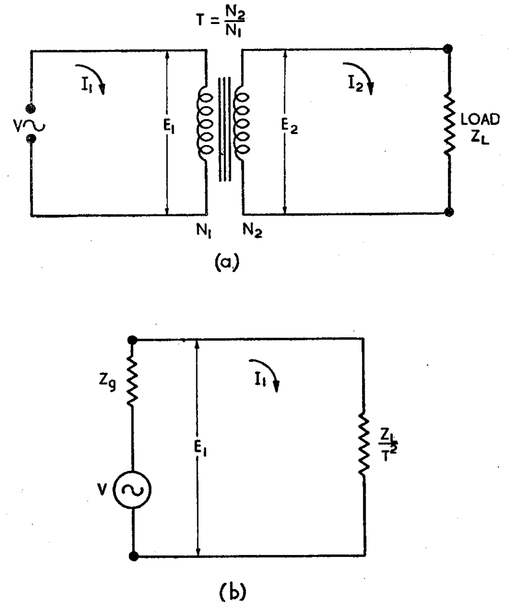
where Z_p = effective primary impedance.

$$\therefore T^2 = \frac{I_2 Z_L}{I_1 Z_p} \times \frac{I_1}{I_2}$$

$$= \frac{Z_L}{Z_p}$$

$$\therefore Z_p = \frac{Z_L}{T^2} \quad .. \quad .. \quad (6)$$

Thus, a secondary load resistance R_L is equivalent to a resistance $\frac{R_L}{T^2}$ shunting the primary winding and a secondary load reactance X_L is equivalent to a reactance $\frac{X_L}{T^2}$ shunting the primary. A transformer does not alter the *angle* of an equivalent primary impedance reflected from the secondary, so that an inductive reactance in the secondary appears as an equivalent inductive reactance in the primary. The equivalent circuit for Fig. 8(a) is illustrated in Fig. 8(b). The internal impedance of the supply is Z_g and the impedance of the primary winding is assumed to be so *large* in relation to the reflected impedance shunting it that it is neglected.



(a)

(b)

Fig. 8—REFLECTED IMPEDANCE

16. The maximum power transfer theorem states that the power taken by a load from a generator is greatest when the impedances of

the generator and the load are equal. In a.c. circuits where the two impedances are not equal, a transformer of suitable transformation ratio can be interposed between the generator and the load. The value of T must be such that the equivalent primary impedance reflected from the secondary load equals the internal impedance of the generator. Maximum transfer of power to the load is then obtained. This process is known as '*impedance matching*'. Fig. 9 illustrates a typical example. The load and generator are shown in Fig. 9(a). If direct connection were made to the load, the power developed in the 4 ohms resistor would be negligible. A transformer is, therefore, interposed. If the impedances are to match:—

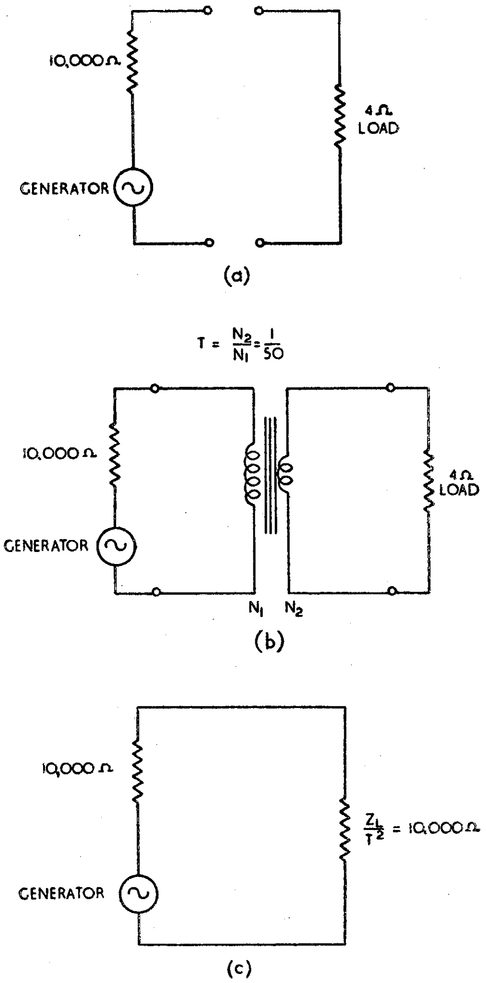


Fig. 9—IMPEDANCE MATCHING

$$T^2 = \frac{Z_L}{Z_p}$$

$$= \frac{4}{10{,}000}$$

$$= \frac{1}{2{,}500}$$

$$\therefore T = \frac{1}{50}$$

Thus a transformer having a *step-down* ratio of 50 : 1 will transform the 4 ohms in the secondary to 10,000 ohms in the primary, and maximum power will then be transferred to the load. The circuit arrangement is given in Fig. 9(b), and the equivalent circuit as seen from the generator in Fig. 9(c)

Transformer Losses

17. The various losses associated with a transformer are summarized below:—

(*a*) **Iron losses.**

(i) *Magnetising current.* In an 'ideal' transformer the primary inductance will offer an infinite impedance and no magnetising current will flow. In practice, however, a magnetising current *does* flow since the primary impedance is not infinite.

(ii) *Eddy current loss.* Resistive (heating) losses are caused by eddy currents circulating in the core of the transformer.

(iii) *Hysteresis loss.* Resistive (heating) losses occur in taking the core through its magnetisation cycle.

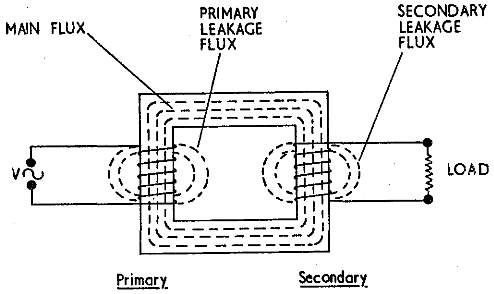


Fig. 10—LEAKAGE FLUX

(*b*) **Copper losses.** These are resistive (heating) losses due to the resistance of the windings, including the losses due to skin effect.

(*c*) **Flux leakage losses.** All the primary flux will not link with the secondary, and *vice versa.* This 'leakage flux' is shown in Fig. 10. The induced voltages are therefore smaller than those indicated by a coupling factor of unity.

(*d*) **Self-capacitance of windings.** Capacitive losses occur because of the capacitance between the turns in the windings. Such losses are important at audio frequencies.

18. Because of the factors above, the efficiency of a transformer, given by $\frac{\text{Output power}}{\text{Input power}} \times 100$ (per cent), is always less than 100%. For small transformers it is around 80 to 90 per cent, rising to 95 per cent for medium transformers and 99 per cent for large transformers.

Construction of Transformers

19. A transformer is constructed in such a way that the losses are kept to a minimum.

(*a*) *Magnetising current.* This is kept to a low value by using, where possible, a primary winding which has a high inductance and a low resistance.

(*b*) *Eddy current loss.* The core is built up of suitably shaped thin laminations, averaging about 0·012 inches thick, of silicon iron or other similar material. The surfaces of the laminations are oxidised to reduce eddy current losses to a minimum.

(*c*) *Hysteresis loss.* This is reduced by using a core material which has a low hysteresis loss, e.g., silicon iron, mumetal, or permalloy.

(*d*) *Copper losses.* The use of heavy gauge copper wire for the windings will reduce these losses.

(*e*) *Flux leakage losses.* The magnetic circuit is so constructed that it has a low value of reluctance, the laminations being arranged in such a way that air gaps are kept to a minimum.

(*f*) *Self-capacitance of windings.* Due attention is paid to the method of winding the turns.

20. Two forms of transformer magnetic circuit in common use are illustrated in Figs. 11 and 12. These are:—

(*a*) *Core type*, shown in Fig. 11 (a), in which parts of both primary and secondary windings are wound on opposite limbs to give a good flux linkage between the windings. The laminations are either L shaped or I and U shaped as shown in Fig. 11(b), and they are arranged in such a manner that all joints are staggered to give a low value of flux leakage (Fig. 11(c)). This type of magnetic circuit is used mainly for high voltage transformation at low power levels.

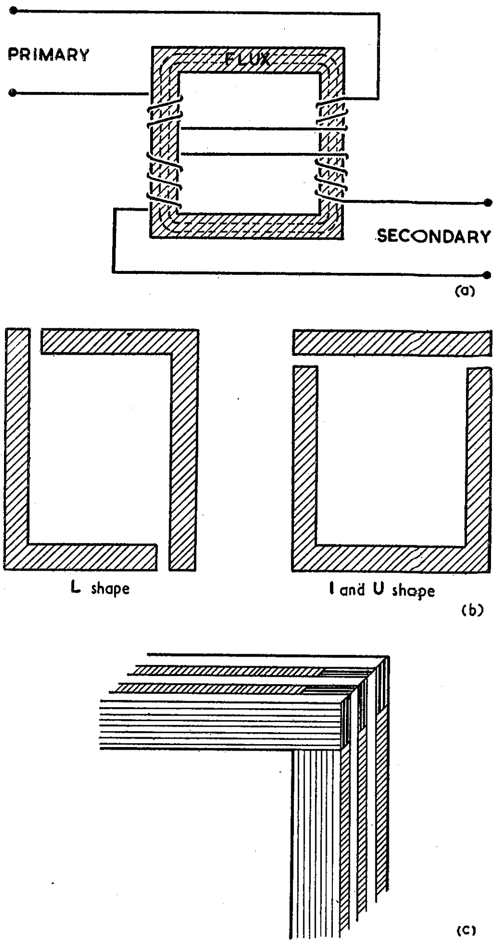


Fig. 11—CORE TYPE OF TRANSFORMER MAGNETIC CIRCUIT

(*b*) *Shell type*, shown in Fig. 12(a), in which all the windings are placed on the centre limb, the two outer limbs completing the magnetic circuit. To provide a constant value of reluctance at all points in the magnetic circuit, the cross-sectional area of the centre limb is made *twice* that of the remainder of the circuit. The

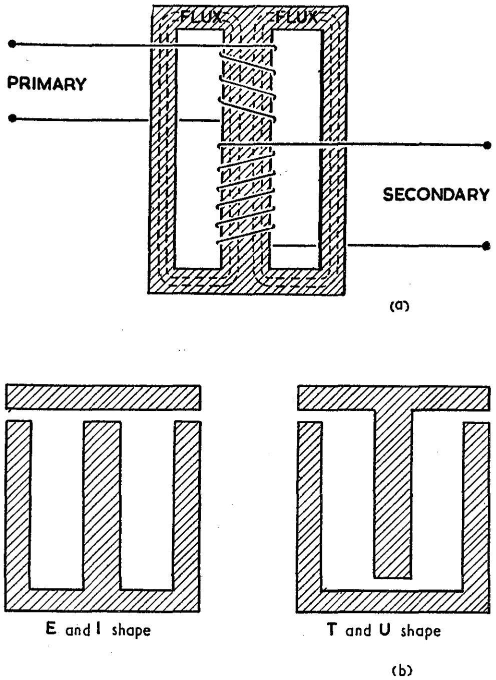


Fig. 12—SHELL TYPE OF TRANSFORMER MAGNETIC CIRCUIT

laminations are either E and I shaped, or T and U shaped as shown in Fig. 12(b), and the joints are staggered to give a low value of flux leakage. This type of magnetic circuit is in more common use and can be used at higher power levels than the core type.

21. The windings are usually of insulated copper wire wound on the limbs or limb of the core. The normal form of insulation for the wire is enamel, and the layers are interleaved with paper. Some thicker insulation is provided between the separate windings, and between the core and the first layer. In some transformers there are several secondary windings providing different output voltages, and the connecting leads to the separate windings are brought out to metal tags mounted on an insulated base. The circuit connections are made to these tags. Two methods of winding are in common use:—

(*a*) *Cylindrical.* The windings consist of concentric coils wound one inside the other and suitably insulated from each other as shown in Fig. 13(a).

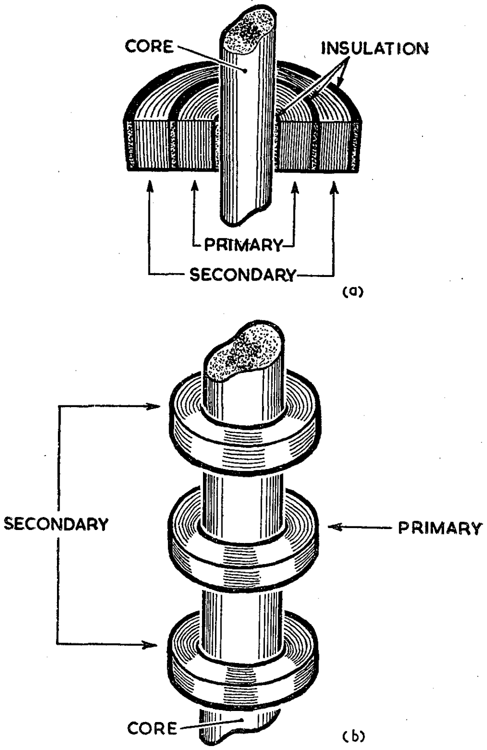


Fig. 13—METHODS OF WINDING

(*b*) *Sandwich.* These windings are separate ring coils assembled one on top of the other on the core as shown in Fig. 13(b).

22. Where self-capacitance between turns in a winding must be reduced to a minimum (e.g., in a.f. transformers) it is usual to use 'sectionalised windings' and also to insert an electric screen of copper foil between the windings. A sectionalised winding is one where the primary and the secondary windings are wound in small series-connected sections, similar to that of a pie-wound inductor, thereby reducing the self-capacitance. The sectionalised winding can be of cylindrical form (Fig. 14(a)) or sandwich form (Fig. 14(b)).

Power Transformers

23. One of the more important purposes of a transformer is the transmission of power. The transformer receives its supply from the a.c. mains or other source and transfers the power to the load at either a higher or a lower voltage level, depending on whether

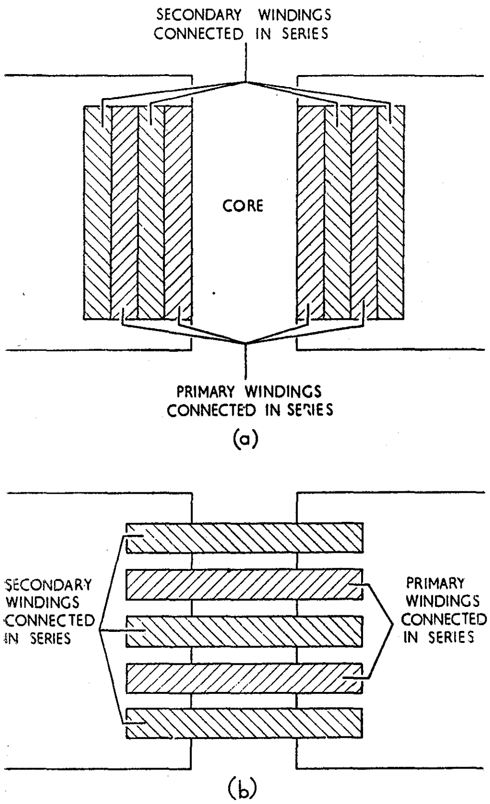


Fig. 14—SECTIONALISED WINDINGS

the transformer has a step-up or a step-down ratio. The input is normally 230V at 50 c/s and the primary winding has several tapping points in order to adjust the transformation ratio for any variation in input voltage (Fig. 15). Several secondary windings are usual. In certain installations, the frequency of the supply is higher than 50 c/s, being of the order of 400 c/s to 2,400 c/s. This simplifies design of the transformer and its associated circuits since the increased rate of change of flux will give higher induced voltages for the same number of turns. The number of turns in all the windings, and hence the size of the transformer, can therefore be considerably reduced.

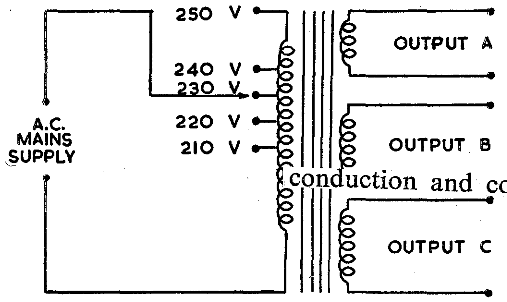


Fig. 15—MAINS TRANSFORMER WINDINGS

24. Although the percentage loss of power in transformers is very small, the amount wasted in the form of heat may be quite large when transforming large powers. It is then necessary to limit the temperature rise by cooling. In small transformers, natural air circulation is sufficient to ‘cool by conduction and convection”. the heat generated in the core and the windings. A power transformer of this type is shown in Fig. 16. In larger types, oil cooling is general and the transformer is immersed in a tank filled with insulating oil. The oil is carried by convection currents to the surface where it radiates its heat to the

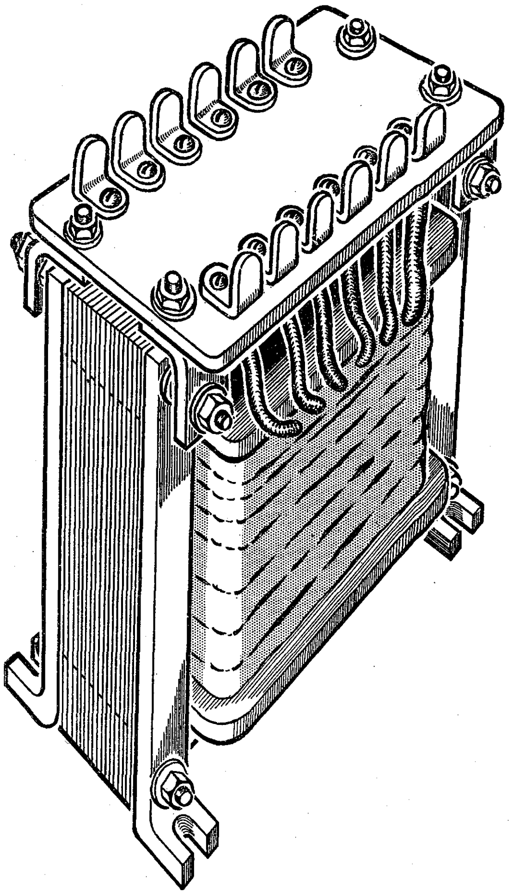


Fig. 16—AIR-COOLED POWER TRANSFORMER

atmosphere. The oil, apart from being a good cooling medium, assists in maintaining the insulation of the windings.

Audio Frequency (a.f.) Transformers

25. These transformers are used in radio equipments to provide voltage transformation and impedance matching. They must operate satisfactorily without undue distortion, over a fairly wide frequency range in the band 20 c/s to 20 kc/s.

26. **Distortion of waveform.** In a transformer, if the magnetising current I_M is sinusoidal so also is the magnetising force H. The flux Φ resulting from the magnetising force can be determined from a hysteresis loop for the core material. Fig. 17 shows the result for a typical material. The flux is seen to be distorted and lagging in phase on the magnetising force H because of the shape of the hysteresis loop. The curve for the rate of change of flux $\frac{d\varnothing}{dt}$ leads Φ by 90° and is considerably distorted because of the flattening at the peaks of the Φ curve. The secondary output voltage is proportional to $\frac{d\varnothing}{dt}$ and is similarly distorted, a pronounced third harmonic component being in fact produced. If the material is allowed to reach saturation the distortion is even more pronounced, and for this reason *air-gaps* are often included in transformer magnetic circuits. To reduce distortion, a material with a narrow hysteresis loop and a high saturation level is selected for the core material.

27. **Air-gaps.** Air-gaps in magnetic circuits are a virtual necessity for a.f. transformers in which the primary winding is carrying

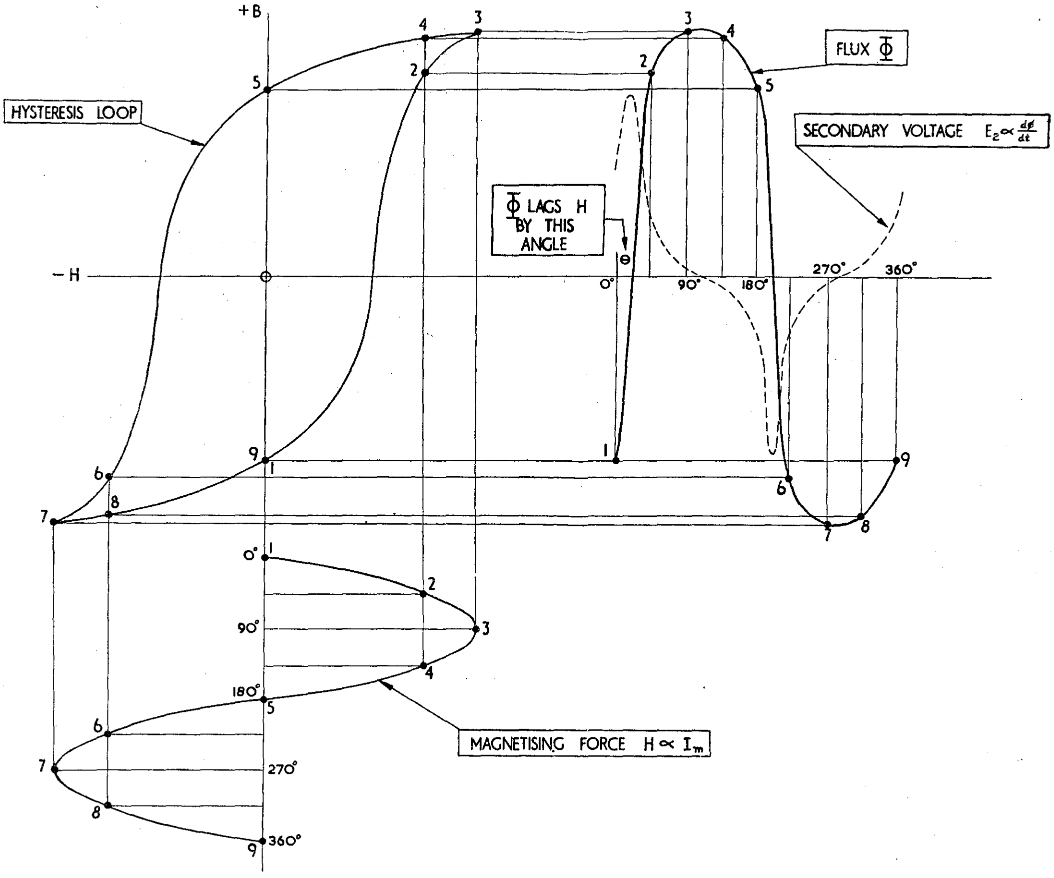
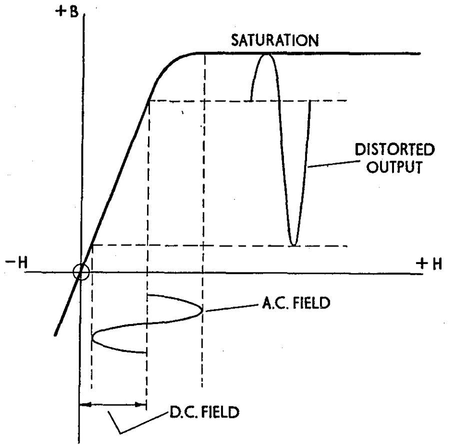
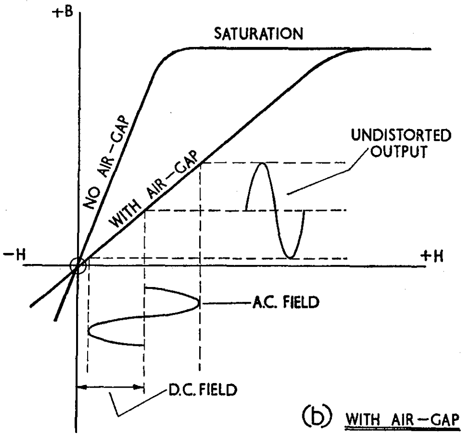


Fig. 17—DISTORTION OF WAVEFORM BY HYSTERESIS



(a) NO AIR-GAP



(b) WITH AIR-GAP

Fig. 18—EFFECT OF AIR-GAPS

d.c. as well as a.c. components. Unless this is done, severe distortion of the output waveform will result. Fig. 18(a) shows the effect of super-imposing an a.c. field on a d.c. field for a transformer which has no air-gap. With air-gaps in the magnetic circuit the slope of the hysteresis curve is reduced and the distortion reduced as shown in Fig. 18(b).

28. **Self-capacitance.** This is reduced in a.f. transformers by using sectionalised windings and by the insertion of an electric screen between the windings. This screen is earthed so that electric lines of force terminate on the screen, but it is so constructed that it has no effect on magnetic lines of force.

29. **Types of a.f. transformer.** Three types of a.f. transformer are common:—

(*a*) *Amplifier input transformers.* These are inserted in the input circuit of an amplifier for matching purposes and to step up the input voltage to a satisfactory level. The primary impedance must be high, and this is obtained by using a high permeability mumetal core and several thousand turns. Mumetal can be used in this case since the primary winding carries no direct current which would otherwise saturate the core and cause distortion. A step-up ratio is usual up to a limit of about 8 : 1.

(*b*) *Intervalve transformers.* These are inserted between two stages in an amplifier. Again the primary inductance must be large, but since the primary winding carries direct current in addition to the audio frequency current, a mumetal core cannot be used. To prevent magnetic saturation and resultant distortion, a silicon iron core *with an air-gap* is used. Because of the large number of primary turns, the step-up ratio is limited to about 4 : 1. A typical intervalve a.f. transformer is shown in Fig. 19.

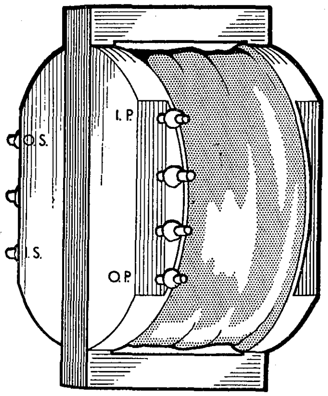


Fig. 19—TYPICAL A.F. TRANSFORMER

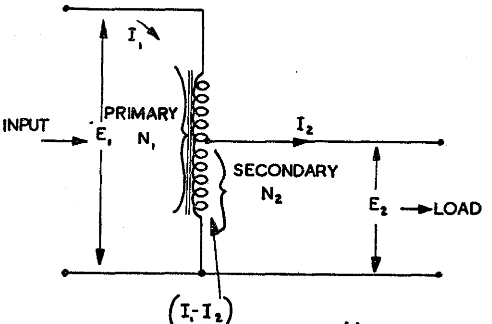
(*c*) *Amplifier output transformers.* These transformers transfer power from the amplifier to the load. They usually carry direct current in the primary so that a silicon iron core with an air gap is usual. The transformation ratio is usually *step-down* and depends on the impedance matching required.

Auto-transformers

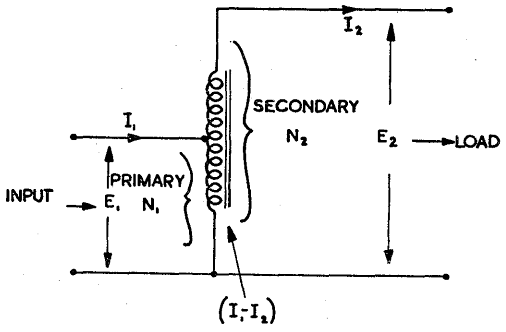
30. In the auto-transformer the primary is part of the secondary, or *vice versa* (Fig. 20).

On applying a voltage E_1 to the primary winding of N_1 turns, a current I_1 is established. The flux resulting from this current induces a voltage E_2 in the secondary winding of N_2 turns so that a current I_2 is established in the load. The auto-transformer can be either step-down as shown in Fig. 20(a), or step-up as shown in Fig. 20(b), depending on whether N_2 is less or greater than N_1. Thus, as with a normal transformer:—

$$\frac{E_2}{E_1} = \frac{N_2}{N_1} = \frac{I_1}{I_2} = T.$$



(a) STEP-DOWN



(a) STEP- UP

Fig. 20—AUTO-TRANSFORMERS

31. The main advantage of the auto-transformer is that copper and leakage losses are reduced, for the current in the secondary winding is the *difference* between I_1 and I_2 and this is obviously less than I_2. This advantage is greatest when N_1 and N_2 are nearly equal, for then I_1 and I_2 are also nearly equal, and the accompanying losses are small. Hence, auto-transformers are most useful when a *small* transformation ratio is required.

Variac

32. Fig. 21(a) illustrates the arrangement of a continuously variable auto-transformer known as a 'variac'. It consists of a layer toroidal winding round a ring-shaped core, the moving brush contact A sliding round a bared track on the toroid to alter the transformation ratio as required. In this way, a continuously adjustable output voltage is obtained. The construction is shown in Fig. 21(b).

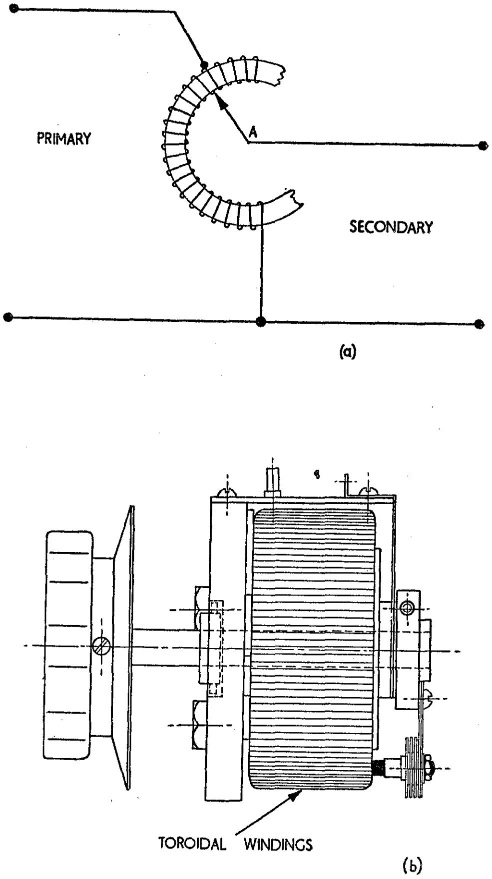


Fig. 21—VARIAC

Three-phase Transformers

33. Transformation of three-phase a.c. at one voltage to three-phase a.c. at another voltage can be effected either by means of three *separate* single-phase transformers, or by a *single* three-phase transformer. In the latter, the shell type iron core of *equal* limb area carries the primary and secondary

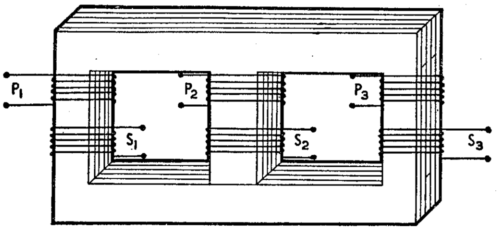


Fig. 22—THREE-PHASE TRANSFORMER

windings of each phase on each of the three limbs as shown in Fig. 22. In effect, each limb is used as a separate transformer for one phase.

34. All the variations possible with three-phase generator and load connections (see Sect. 5, Chap. 4) are available with the primary and secondary windings of a three-phase transformer. For example, the primary and secondary may both be star-connected (Fig 23(a)), in which case the line voltages are connected by the relation:—

$$\frac{V_2}{V_1} = \frac{N_2}{N_1} = T.$$

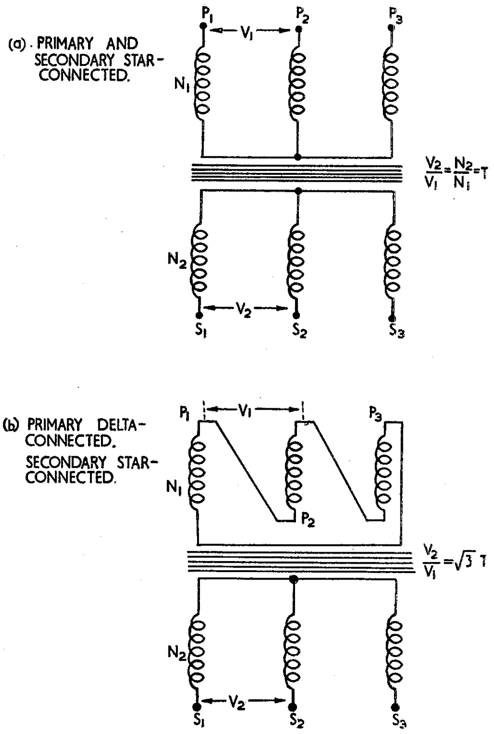


Fig. 23—CONNECTIONS TO A THREE-PHASE TRANSFORMER

Alternatively, the primary may be delta-connected and the secondary star-connected (Fig. 23(b)), when:—

$$\frac{V_2}{V_1} = \sqrt{3}\ T.$$

35. For the operation of certain radio equipments a *two-phase* power supply is required, and it is then necessary to convert from the three-phase mains supply to a two-phase supply. It is possible to do this by a suitable arrangement of transformers. A typical circuit, known as '*the Scott method*' is given in Fig. 24. Two transformers are used, with one winding on one transformer centre-tapped to one winding on the other, to give a star-connected arrangement. By applying the three-phase input to the terminals A, B and C a two-phase output is obtained across the terminals PQ and RS. Conversion from two-phase to three-phase can be obtained by reversing the procedure.

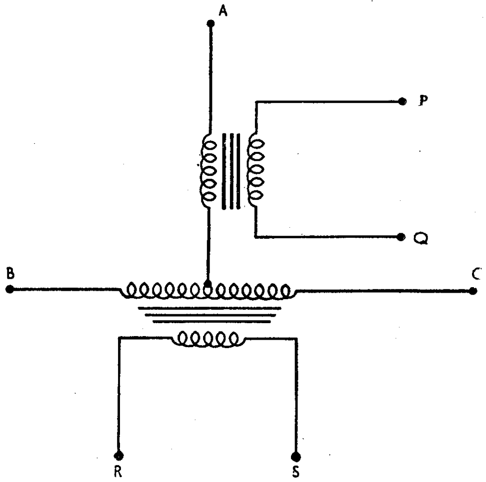


Fig. 24—SCOTT CONNECTIONS

Instrument Transformers

36. When measuring large currents in d.c. circuits it is usual to use shunts to limit the current through the meter to the f.s.d. value (see Sect. 6, Chap. 1). This is not convenient with a.c. instruments and it is, therefore, usual to pass the current to be measured through the primary of a 'current' transformer, the secondary being connected to the ammeter as shown in Fig. 25. The

step-up ratio *reduces* the secondary current in the ratio $\frac{N_2}{N_1}$ and by suitably adjusting this ratio the required f.s.d. current can be obtained. Measurement of high values of alternating voltage is normally obtained by using a *step-down* transformer.

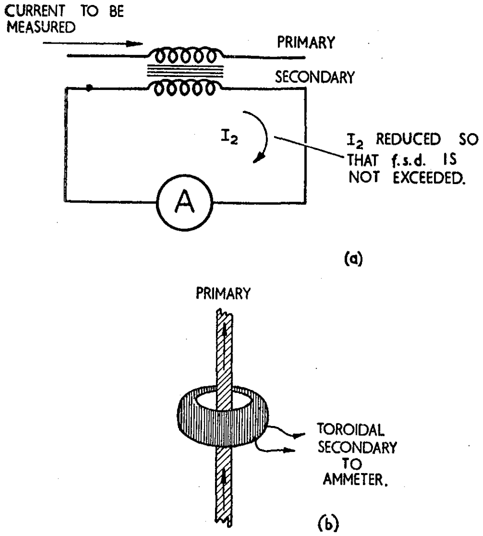


Fig. 25—'CURRENT' TRANSFORMER

Phase-shifting Transformers

37. Phase-shifting transformers serve the same purpose with respect to phase as do current and voltage transformers with respect to amplitude; they are used to alter the phase but not the amplitude of an alternating current. The phase-shifting transformer is similar in construction to an induction motor with a wound rotor (see Sect. 5, Chap. 4), except that the 'rotor' is set in any desired position but does not thereafter rotate. The stator windings are the primary of the transformer and the rotor is the secondary, the primary windings producing a rotating magnetic field. The primary flux threads the secondary, and whatever the position of the secondary the flux linked with it goes through one complete cycle for each rotation of the field. Since this is the same however the secondary is placed, the *magnitude* of the induced voltage does not depend on the position of the rotor. However, the *phase* of the secondary voltage depends on the position of the rotor with respect to the primary windings. In Fig. 26, if the secondary coil is in position 1 the induced voltage in it is zero when the rotating field is acting in the direction H_1, while if it is in position 2 this happens when the field is in the direction H_2. Rotating the secondary through angle θ thus causes a change of θ in the phase of the secondary voltage.

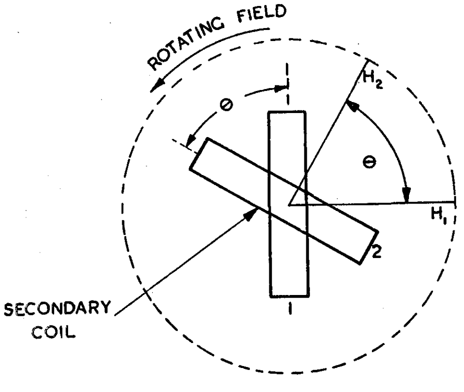


Fig. 26—PHASE SHIFTING TRANSFORMER

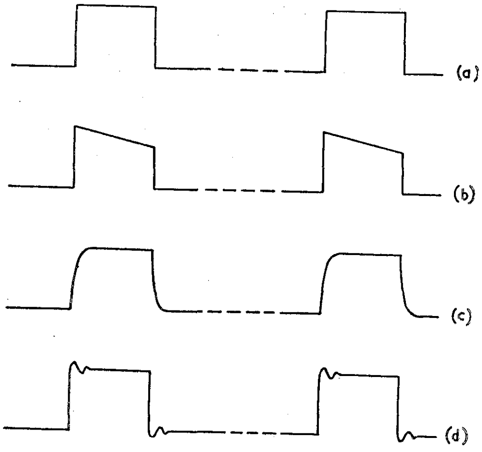


Fig. 27—DISTORTION IN A PULSE TRANSFORMER

Pulse Transformers

38. These are used to handle very short duration pulses of rectangular waveform as shown in Fig. 27(a). They must be so designed that distortion of the waveform by the transformer is kept to a minimum. Distortion may be due to several factors:—

(*a*) A low value of primary inductance will introduce distortion of the type shown in Fig. 27(b). The primary inductance is kept high and the distortion reduced by using a high permeability material for the core and a winding of many turns.

(*b*) Self-capacitance and resistive losses will introduce distortion of the type shown in Fig. 27(c). Self-capacitance is reduced by proper arrangement of the windings. Resistive losses are due mainly to eddy current and hysteresis losses and these are reduced by the use of very thin laminations and a material with a low hysteresis loss.

(*c*) Leakage inductance, together with the self-capacitance of the transformer, will introduce distortion of the type shown in Fig. 27(d). Such distortion is termed 'ringing' and is caused by resonance between the leakage inductance and the self-capacitance. Leakage inductance is reduced by coupling the primary and secondary windings as closely as voltage insulation requirements will permit, and by employing a magnetic core of special high permeability material.

Constant Voltage Transformers

39. This type of transformer is used to deliver a *stabilized* voltage to its load despite variations in the input voltage applied to it. Two forms of constant voltage transformer are commonly encountered:—

(*a*) Motor controlled tapped transformer.

(*b*) Saturated core transformer.

40. **Motor controlled tapped transformer.** This arrangement is normally used to compensate for large input voltage changes in high power circuits. An electric motor, fitted with a governor, is run from the same supply as that used for the input to the transformer. Due to the action of the governor the motor runs at approximately constant speed despite variations in the supply voltage. The movement of the governor in maintaining constant speed is used to operate a switch connected to 'taps' on the transformer windings. The transformation ratio is thereby automatically adjusted in such a way that the secondary output voltage is virtually unaffected by variations in the supply voltage.

41. **Saturated core transformer.** This type of transformer is constructed as shown in Fig. 28. It consists of a tuned secondary winding S *loosely* coupled to the primary winding P (by virtue of the magnetic shunt leakage path L) and a compensating winding C *tightly* coupled to the primary. When a low voltage is applied to the primary most of the flux threads the secondary, producing a secondary voltage proportional to the transformation ratio; little flux takes the shunt leakage path L since the air gap produces a relatively high value of reluctance. As the applied voltage increases, the circulating current in the tuned circuit increases and section B of the core begins to saturate so that the reluctance of this part of the core increases. The reluctance of the leakage path L is now low in relation to that of section B of the core and most of any further flux increase due to an increase in primary voltage takes the leakage path. The secondary voltage still rises very slowly, but this is compensated for by the small antiphase voltage acting in series with it from winding C. The constant output voltage may be taken from the whole or part of the secondary.

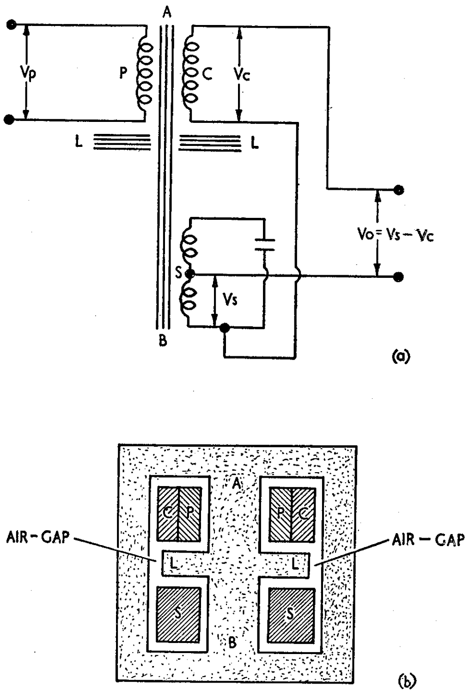


Fig. 28—SATURATED CORE TRANSFORMER

42. Because of the deliberate saturation of the core, distortion of the input waveform is inevitable. The resonant secondary circuit minimizes this effect, but where a pure a.c. waveform is required the secondary output voltage is passed through a suitable 'filter'. Fig. 29 shows how the secondary output voltage varies with input voltage for several load conditions.

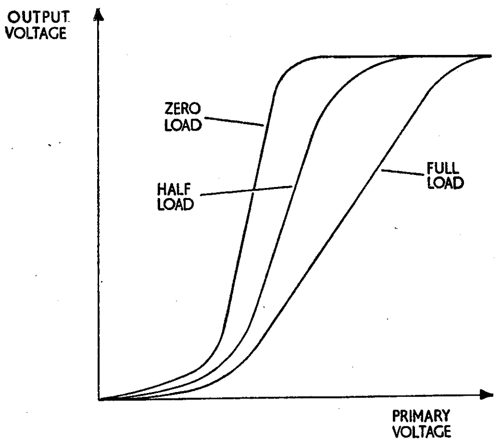


Fig. 29—VARIATION OF OUTPUT VOLTAGE WITH INPUT VOLTAGE

SECTION 7

CHAPTER 3

TRANSDUCTORS

Paragraph

PART 1, SECTION 7, CHAPTER 3

TRANSDUCTORS

Introduction

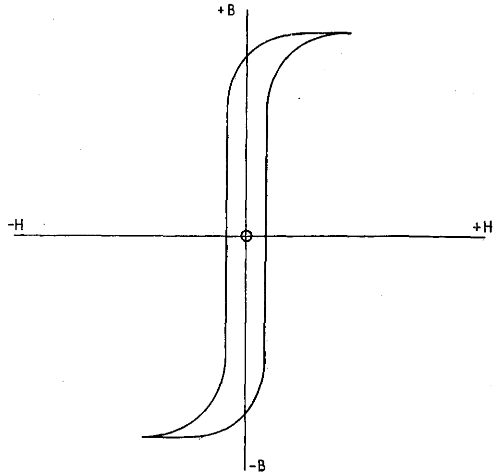
1. The transductor, or saturable reactor, is the main circuit element in 'magnetic amplifiers'. The magnetic amplifier will be considered in detail in Sect. 10, but since the transductor consists of two or more coils wound on a core of magnetic material, it has many of the features of an iron-cored transformer and it is convenient to discuss the basic principles of transductors in this Section.

Saturable Reactor

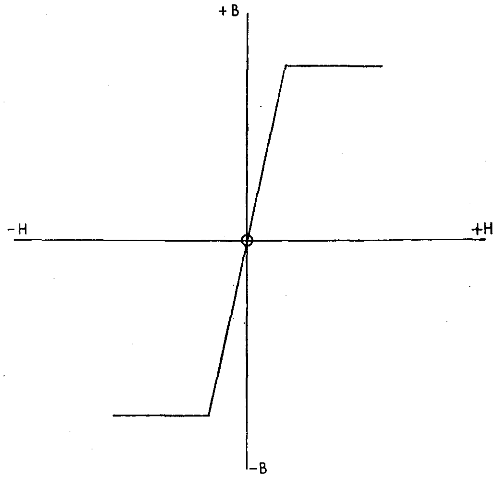
2. It was shown in Sect. 2, Chap. 2 that the inductance of a coil wound on a core of magnetic material is proportional to the absolute permeability of the core; that is, $L = N^2 \frac{\mu a}{l}$. However, the permeability of a magnetic circuit is *not* a constant (see Sect. 2, Chap. 1). Provided the core is not saturated, its permeability is high and so also is the inductance. If the core is saturated, the permeability falls off rapidly and the inductance falls to a low value.

3. If a coil, of negligible resistance, is wound on a core of magnetic material and connected to a source of alternating voltage V, the current established in the coil is $I = \frac{V}{\omega L}$. Thus, when the core is unsaturated and L is a large value, the current will be small; when the core is in the saturated condition so that L falls to a low value, the current will be large.

4. In most applications this change in inductance is undesirable, and core materials which gradually approach saturation are generally used in transformers and in electric machines. Transductors, however, operate *by virtue* of this saturation effect and the core materials used in transductors are those which saturate sharply for a small change in magnetising force, e.g., mumetal. The hysteresis loop for mumetal is given in Fig. 1(a), the ideal curve for a transductor material being as shown in Fig. 1(b).



(a) MUMETAL



(b) IDEAL

Fig. 1—HYSTERESIS LOOPS FOR TRANSDUCTOR MATERIALS

A.C. Applied to a Saturable Reactor

5. Fig. 2 shows an iron-cored inductance L connected in series with a resistance R across an a.c. supply of variable voltage V. The hysteresis loop for the core material may be

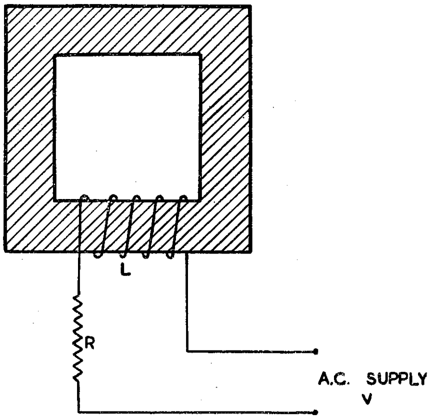


Fig. 2—A.C. APPLIED TO A SATURABLE REACTOR

assumed to have the ideal form of Fig. 1(b) so that the core will saturate at a low value of m.m.f.

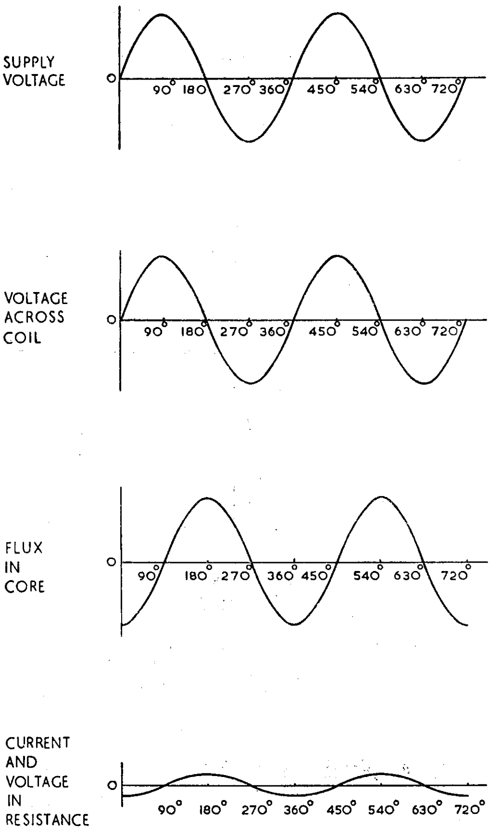


Fig. 3—AMPLITUDE INSUFFICIENT TO CAUSE SATURATION

6. Fig. 3 illustrates the conditions in the circuit when a low value of alternating voltage, insufficient to saturate the core, is applied. In the unsaturated state, the inductance L is high and the resistance R can be neglected in comparison. The impedance of the circuit is then high and inductive, so that the current established is small and lags on the applied voltage by 90°. The current is, in fact, merely the magnetising current necessary to produce the flux in the core. The flux itself is undistorted and lags the applied voltage by 90°.

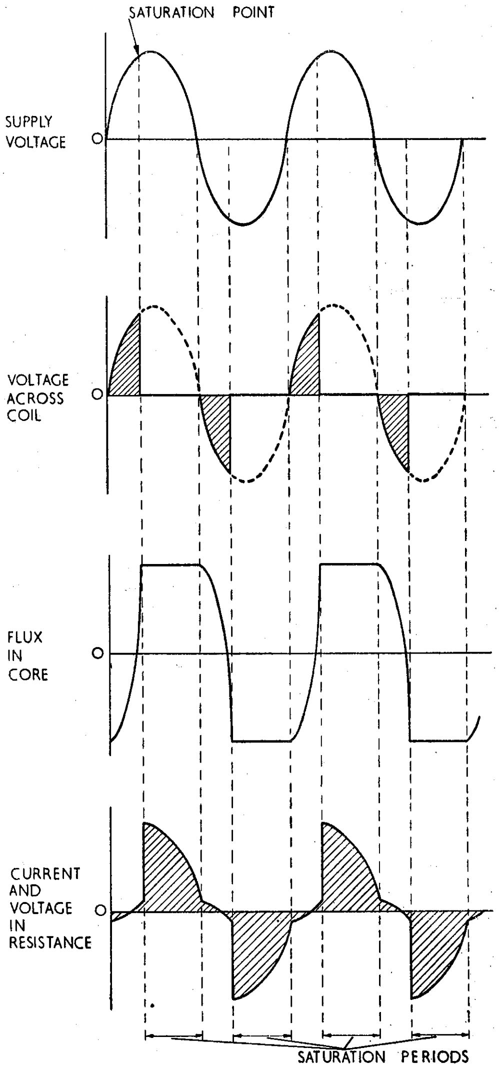


Fig. 4—EFFECT OF SATURATION OF THE CORE

7. Fig. 4 shows the effect of increasing the amplitude of the applied voltage to an extent sufficient to saturate the core. There is little change in the current in the circuit until the voltage reaches the point at which the core saturates. Saturation occurs sharply (because of the shape of the hysteresis loop for the core) and the core remains saturated for a short period on each half cycle. During that period the flux cannot change and, since the back e.m.f. is zero, the inductance may be considered to be short-circuited. From Kirchhoff's second law, the sum of the voltages across the coil and across the resistance must at every instant equal the applied voltage. Since the voltage across the coil has fallen to zero, the instantaneous voltage of the supply is suddenly applied to the resistance R and current flows in the resistance as a series of pulses. The duration of each pulse is equal to the saturation period and this, in turn, is dependent on the amplitude of the applied voltage. As the supply voltage is further increased, the core is held saturated for a longer period in each half cycle and the pulses of current are of longer duration.

8. Summarizing, it is seen that when the core is not saturated, the inductance of the coil is high and the major portion of the applied voltage is developed across the coil. During the saturation period, the inductance of the coil is low and major portion of the applied voltage is developed across the resistance.

D.C. Control of Saturation

9. While the saturation period can be varied by adjusting the amplitude of the supply voltage, it is much more convenient to provide control by an external circuit. Fig. 5(a) shows a saturable reactor with an

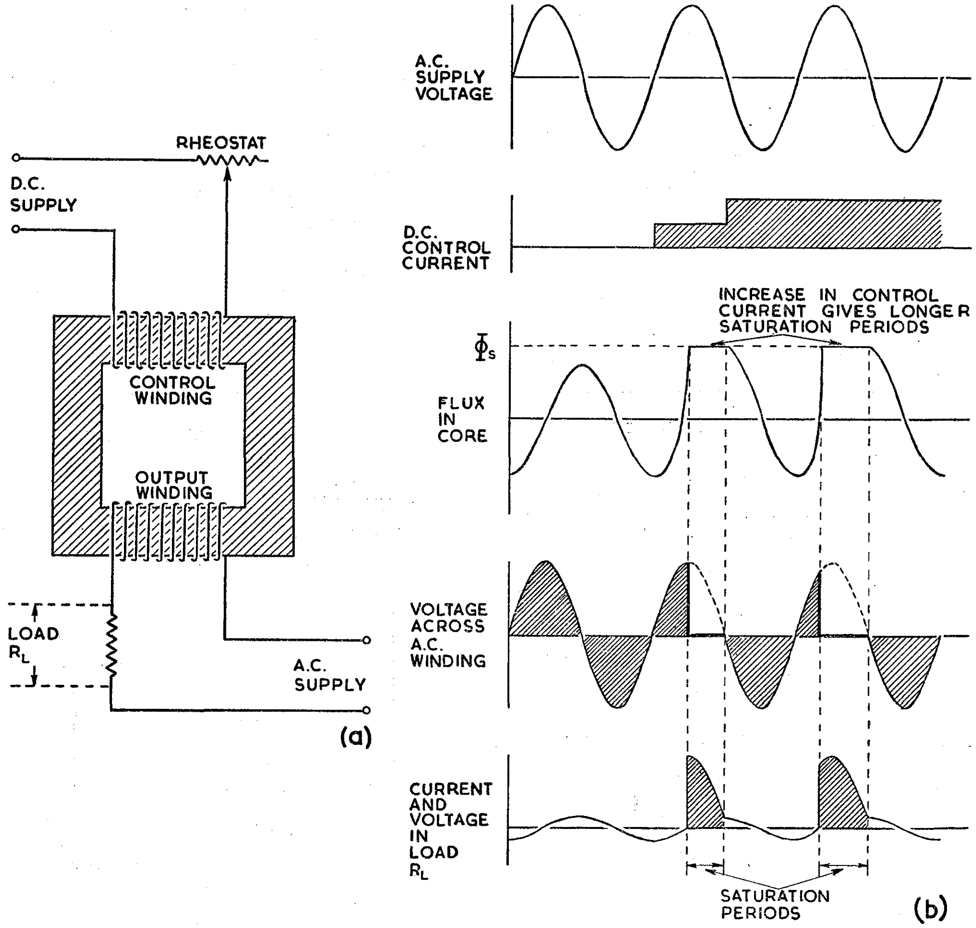


Fig. 5—EFFECT OF CONTROL WINDING

additional winding, known as the *control winding*, which is connected to a source of d.c. supply via a rheostat. The amplitude of the a.c. supply voltage is fixed. With zero current in the control winding, the flux in the core almost, but not quite, reaches saturation point. A direct current in the control winding produces additional flux and, since this additional flux is unidirectional it causes the core to saturate on *every other* half cycle as shown in Fig. 5(b).

10. Variation of the control current causes the periods of saturation to be correspondingly increased or decreased. Thus, the pulses of current through the load resistance R_L, and the voltage developed across this resistance, are similarly varied. Quite small changes in the control current will cause large changes in the alternating current in the load resistance R_L. When a varying control signal is applied to the control winding an amplified signal appears in the a.c. circuit. This is the basis of the magnetic amplifier.

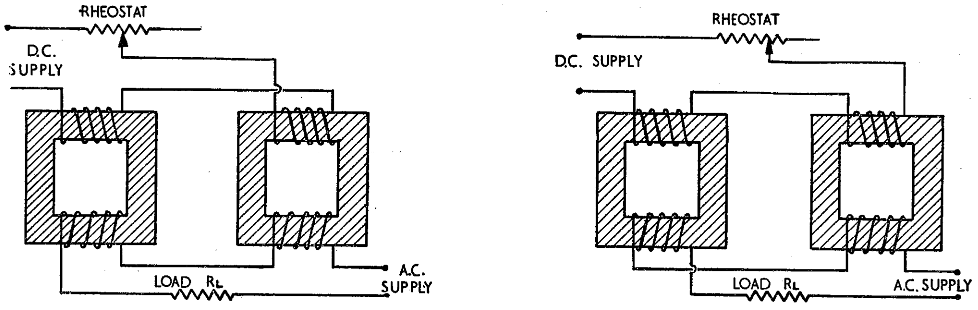
Practical Form of Transductor

11. The simple transductor described in Paras. 9 and 10 suffers from two disadvantages:—

(*a*) The control current in the d.c. winding affects only one half cycle of the a.c. supply.

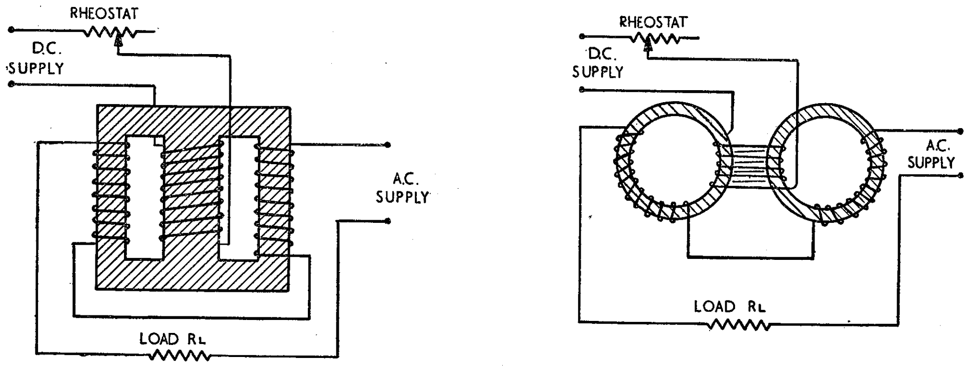
(*b*) The output and control windings are coupled by the same magnetic circuit and the alternating voltages induced in the control circuit by transformer action from the output winding may be sufficiently large to prevent effective control.

12. The first defect is remedied by providing *two* magnetic circuits which will saturate on alternate half cycles of the a.c. supply. The second defect is remedied by arranging the windings and cores so that the transformer effect is neutralized. Some typical arrangements are shown in Fig. 6.



(a) CONTROL WINDINGS IN OPPOSITION

(b) OUTPUT WINDINGS IN OPPOSITION



(c) THREE-LIMB TRANSDUCTOR

(d) TWO TOROIDS WITH COMMON CONTROL WINDING

Fig. 6—PRACTICAL FORMS OF TRANSDUCTOR

13. The transductor may be made in two identical portions as shown in (a) and (b) of Fig. 6. Voltages at the supply frequency induced in the control windings will be cancelled if the two halves of either the control winding or the output winding are wound in opposition. Alternatively, the magnetic circuits can be grouped so that the fluxes due to the alternating current in the output winding are in opposition through the control winding and no alternating voltage will be induced in that winding. This is the arrangement shown in (c) and (d) of Fig. 6. Although the output windings are shown connected in series in Fig. 6, they may be connected in parallel and still achieve the desired result.

Operation of Transductor

14. Fig. 7 illustrates an elementary transductor which consists of two identical inductors A and B enclosed by a common control winding. The output windings are series-connected and have the same number of turns so that the load current I_L produces equal fluxes Φ_A and Φ_B in cores A and B respectively. The common control winding is connected to a d.c. supply via a rheostat by means of which the control current may be varied. The output windings are connected in series with a load resistance R_L across an a.c. supply of voltage E. The amplitude of E is such that with no control current I_C in the control winding, the cores will at peak, be nearly but not quite saturated. In this unsaturated state the output windings have a very high impedance and the load current I_L will be very small.

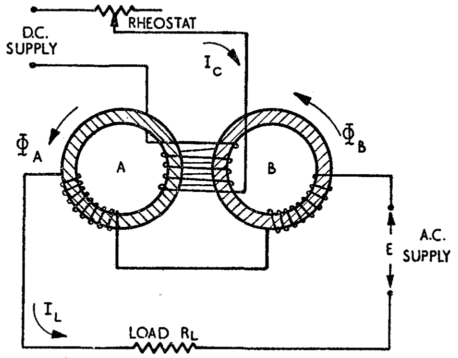


Fig. 7—CONNECTIONS TO AN ELEMENTARY TRANSDUCTOR

15. As the cores are so near to saturation, even a very small current I_C in the control winding will produce sufficient additional flux to cause both cores to saturate. However both cores do *not* saturate simultaneously since the flux produced by the direct current

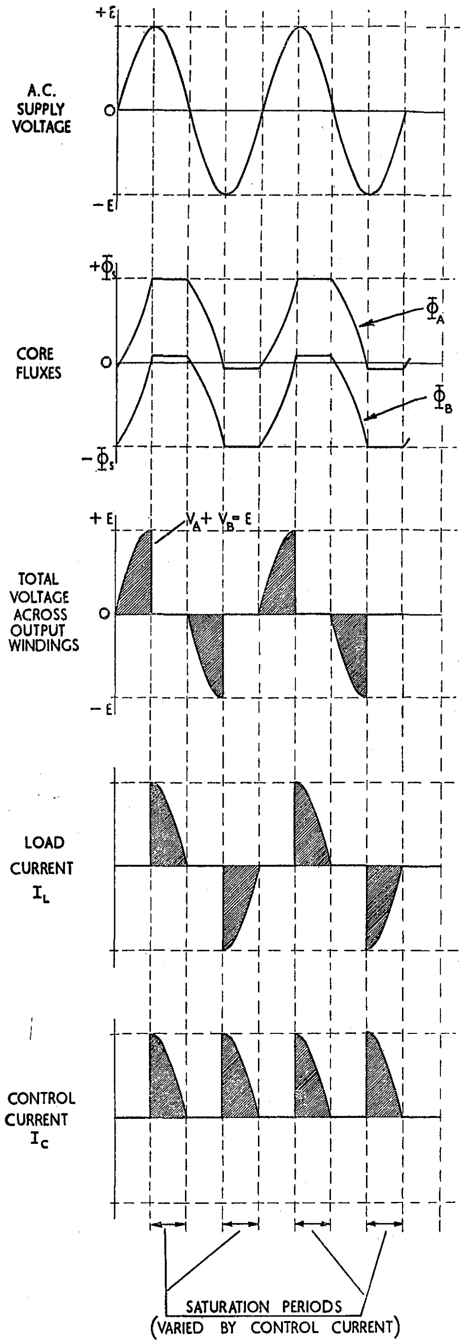


Fig. 8—OPERATION OF A SIMPLE TRANSDUCTOR

I_c is unidirectional, and will alternately add to and oppose the alternating fluxes in each core. That is, on one half cycle of the alternating voltage, the unidirectional flux will be in the same direction as the flux Φ_A in core A and, at the same time, it will be opposing the flux Φ_B in core B. On the next half cycle of the a.c. supply, the relative directions of the alternating fluxes will be changed, and the unidirectional control flux will now add to the flux Φ_B and oppose the flux Φ_A. With current I_c in the control winding, the fluxes will, therefore, be displaced as shown in Fig. 8, so that core A will be saturated on one half cycle of the a.c. supply and core B on the next.

16. In the unsaturated state the impedance of a transductor is high and, for the purpose of illustration, it may be assumed that there is zero load current unless the cores are saturated. Since the resistance of the load resistor R_L is very small compared with the impedance of the unsaturated transductor it may be ignored, and the voltage across each output winding will be equal to half the alternating voltage supply; that is, $V_A + V_B = E$. Assuming that the transductor is already in operation and that the hysteresis loop for each core is of the ideal form of Fig. 1(b), then the voltage, current and flux waveforms will be as shown in Fig. 8.

17. When flux Φ_A reaches the saturation point Φ_S there can be no further change of flux and the back e.m.f. falls to zero. Voltage V_A across inductor A falls sharply to zero and the winding behaves as though short-circuited. Inductor B however, is *not* saturated and behaves as a transformer together with the control winding for the period in which inductor A is saturated. Because of transformer action, the low impedance of the control circuit is reflected into the output winding B, making the voltage V_B very small and holding Φ_B constant at the particular value it has reached at that instant. With the voltages across the output windings A and B suddenly falling to zero, the full instantaneous supply voltage is applied to the load R_L and load current I_L flows. Since inductor B is behaving as a transformer, then to hold the flux Φ_B constant, a pulse of current must be established in the *control* circuit in order to produce a flux in opposition to that produced by the load current pulse.

18. A similar sequence of events occurs on the next half cycle of the a.c. supply. Inductor B saturates and inductor A behaves as a transformer giving the same results as before. Since the flux Φ_A is relatively in opposition to the flux Φ_B, the pulse of control current will be of the *same* polarity as before.

19. When the control current I_c is zero, the cores are unsaturated and the output current I_L is zero. As the control current is increased progressively in either direction, the cores saturate earlier and the output current increases. The control current may be increased until the a.c. supply voltage is applied to the load throughout the *complete* cycle. Thereafter, further increase in the control current will not result in greater output. A graph showing the relationship between the output current and the control current is given in Fig. 9. Amplification is obtained since a small variation in the control current will give a large variation in the output current.

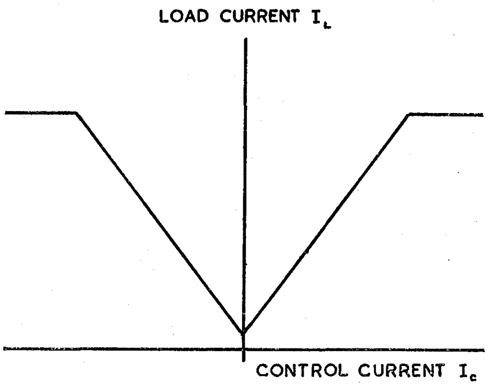
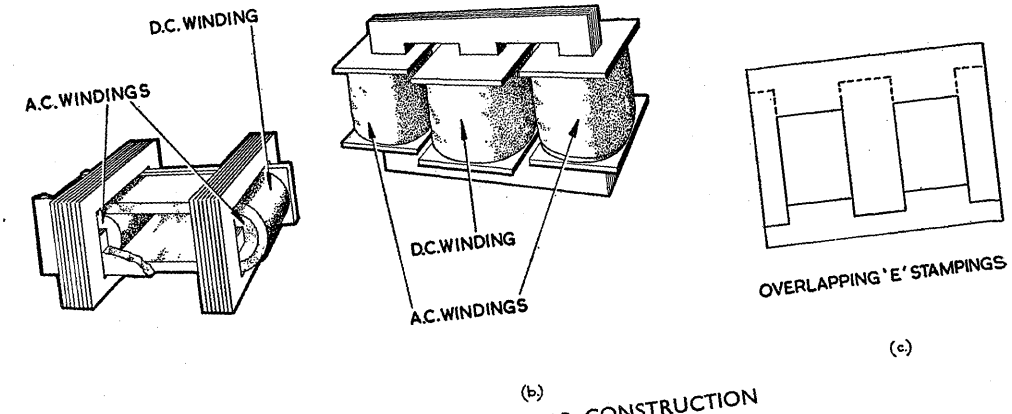


Fig. 9—CHARACTERISTICS OF A SIMPLE TRANSDUCTOR

Construction of Transductors

20. The simple form of transductor consisting of two separate identical cores, as illustrated in (a) and (b) of Fig. 6, is seldom used in practice because, although no alternating voltage is present in the complete d.c. winding, equal and opposite voltages are induced in each half of the control winding. These induced voltages necessitate adequate insulation which occupies considerable winding space. This disadvantage is avoided by using d.c. windings which enclose both magnetic circuits, as illustrated



(a) (b) (c)

Fig. 10—TYPICAL TRANSDUCTOR CONSTRUCTION

in (c) and (d) of Fig. 6. Thus, during the period when neither magnetic circuit is saturated, no alternating voltage is induced in the control winding. Two typical constructional arrangements which are in common use are shown in (a) and (b) of Fig. 10.

21. The cores of the transductors are constructed of laminations in much the same way as normal transformer cores. Special precautions are taken to reduce air gaps to a minimum and it is normal to use special overlapping E type stampings to achieve this, as shown in Fig. 10(c).

22. Even when the greatest care is taken there will always be some residual air gap when interleaved laminations are used. Since a much greater m.m.f. is required to provide a given flux across even a small air gap than to provide that same flux in a continuous core, this residual air gap will have the undesirable effect of reducing the slope of the hysteresis loop. The only type of core which has no air gap is that constructed of ring stampings. Although a special winding technique is involved, the toroidal construction shown in Fig. 11 is the ideal arrangement.

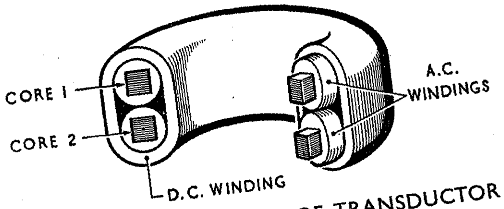


Fig. 11—TOROIDAL TYPE OF TRANSDUCTOR

Advantages of Transductors

23. (*a*) Its robust construction makes the transductor very reliable and ensures long life with no maintenance.

(*b*) The transductor is immediately available from the moment of switching on, i.e., no heating time is required, as it is for valves.

(*c*) The transductor has a high efficiency, of the order of 75 per cent.

(*d*) Power gains of the order of 10^7 can be obtained. The transductor shown in Fig. 12 gives an output of 70 watts for an input of 35 microwatts. This is a power gain of 2×10^6.

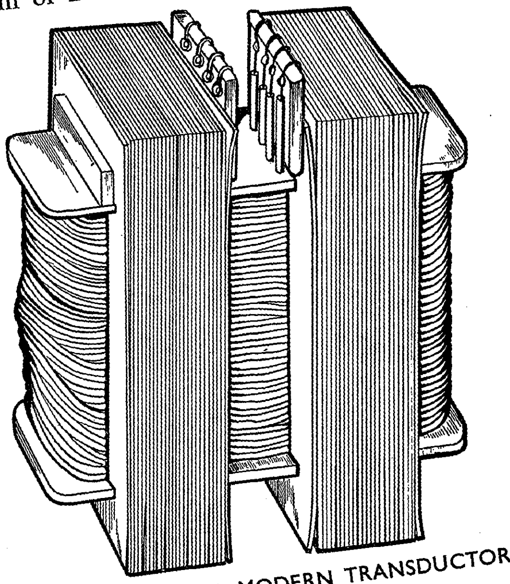


Fig. 12—A MODERN TRANSDUCTOR

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Disadvantages of Transductors

24. (*a*) The value of the output current depends on the value of the control current and since the inductance of the control circuit causes the control current to lag behind the applied signal voltage, a change in signal voltage does not produce an instantaneous change of output current. In the transductor, the delayed response to an applied signal corresponds to the time constant of the control circuit and this may vary from a few milliseconds to about 1 minute.

(*b*) For a short time constant, the supply frequency should be high. However, the latter is limited to a few kc/s since magnetic materials able to give satisfactory results at higher frequencies are not normally available. Because of this the transductor is more suited to the amplification of signals which change slowly, i.e., it forms a useful alternative to the d.c. valve amplifier.