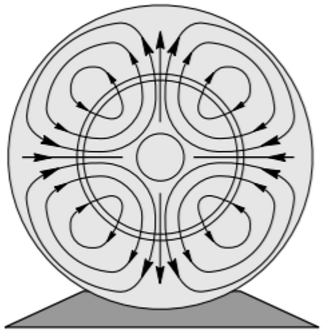
Induction Machine

Wednesday, February 6, 2019 5:13 PM

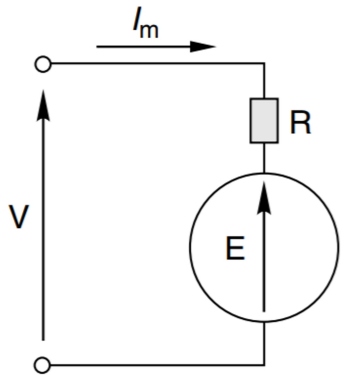
1. 'Electric Motors and Drives' de Austin Hughes
2. Advanced electric drive vehicles de Ali Emadi

Chapter 5

The term ‘4-pole’ reflects the fact that the flux leaves the stator from two N poles, and returns at two S poles. Note, however, that there are no physical features of the stator iron mark it out as being 4-pole, rather than say 2-pole or 6-pole. As we will see, it is the layout and interconnection of the stator coils that sets the pole number.



When we apply voltage to the stator, a variable magnetic field is generated, this magnetic field induces in the same stator BackEMF

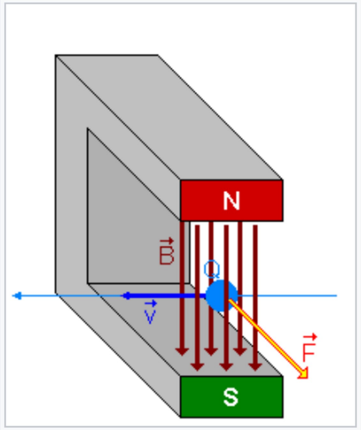


We find in practice that the term Im*R (which represents the volt drop due to winding resistance) is usually very much less than the applied voltage V. In other words most of the applied voltage is accounted for by the opposing e.m.f., E.

But we have already seen that the e.m.f. is proportional to Bm and to f.
So:
$Bm = k \cdot V/f$

When the stator is connected to a 3-phase supply, a sinusoidally distributed, radially directed rotating magnetic flux density wave is set up in the air-gap. The speed of rotation of the field is directly proportional to the frequency of the supply, and inversely proportional to the pole number of the winding. The magnitude of the flux wave is proportional to the applied voltage, and inversely proportional to the frequency. When the rotor circuits are ignored (i.e. under no-load conditions), the real power drawn from the mains is small, but the magnetising current itself can be quite large, giving rise to a significant reactive power demand from the mains (This would be during startup I think, maybe not).

Force in the rotor is produced by Lorentz force:



Fuerza sobre una partícula cargada.

$$\mathbf{f} = \int_L I \cdot d\mathbf{l} \times \mathbf{B}$$

When a mechanical load is applied to the shaft, the rotor slows down, the slip increases, rotor currents are induced and their MMF results in a modest (but vitally important) reduction in the air-gap flux wave. This in turn causes a reduction in the e.m.f. induced in the stator windings and therefore an increase in the stator current drawn form the supply. This is a stable process.

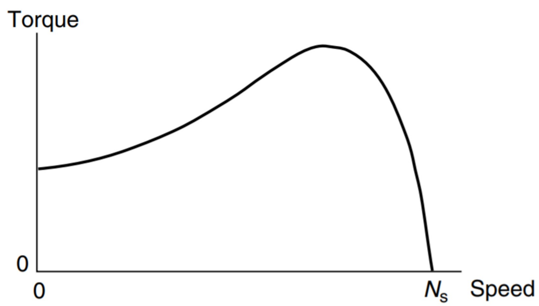


Figure 5.19 *Typical complete torque–speed characteristic for cage induction motor*

We argued that as the slip increased, and the rotor did more mechanical work, the stator current increased. Since the extra current is associated with the supply of real (i.e. mechanical output) power (as distinct from the original magnetising current which was seen to be reactive), this additional 'work' component of current is more or less in phase with the supply voltage, as shown in the phasor diagrams (Figure 5.20). Angulo entre stator voltage y I_load_component ~= 0 .
This approximation is only valid at low slips.

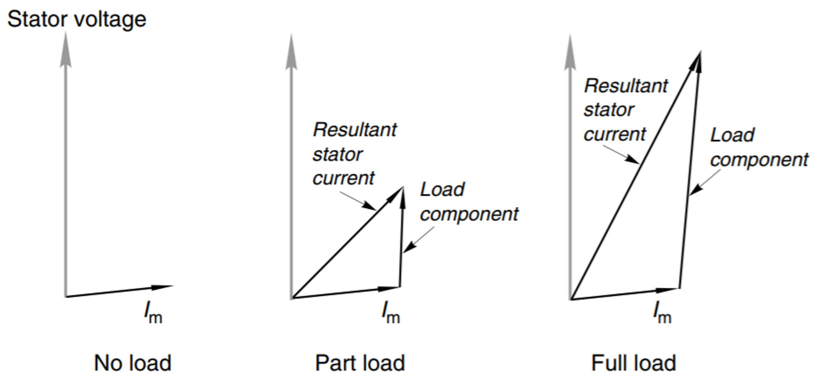


Figure 5.20 *Phasor diagrams showing stator current at no-load, part-load and full-load. The resultant current in each case is the sum of the no-load (magnetising) current and the load component*

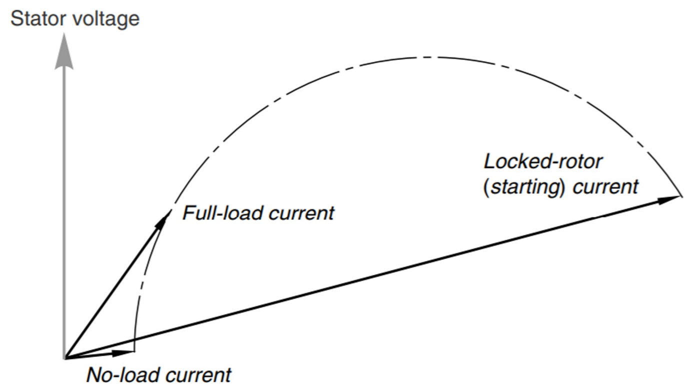
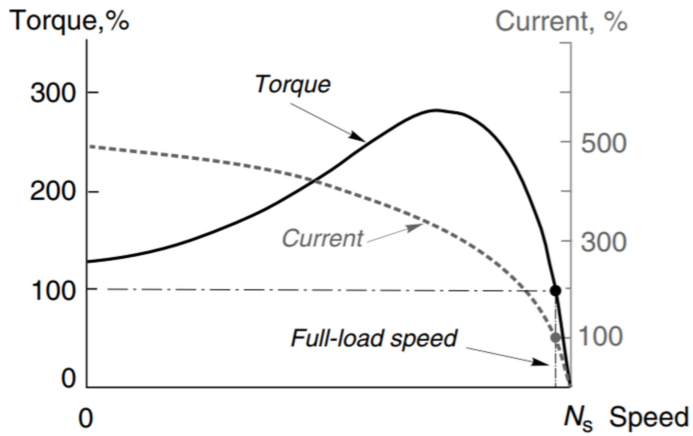


Figure 5.21 *Phasor diagram showing the locus of stator current over the full range of speeds from no-load (full speed) down to the locked-rotor (starting) condition*



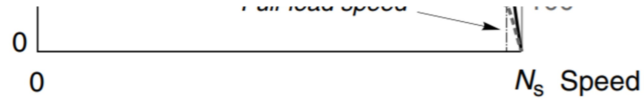


Figure 5.22 *Typical torque–speed and current–speed curves for a cage induction motor. The torque and current axes are scaled so that 100% represents the continuously rated (full-load) value*

Chapter 6: OPERATING CHARACTERISTICS OF INDUCTION MOTORS

Different types of loads

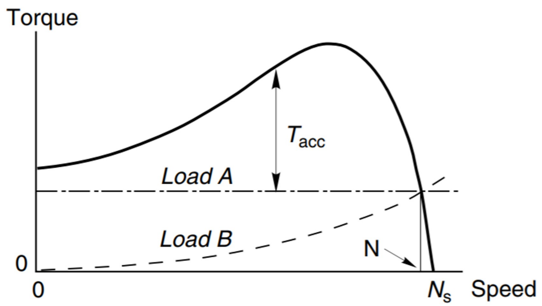


Figure 6.5 *Typical torque–speed curve showing two different loads which have the same steady running speed* (N)

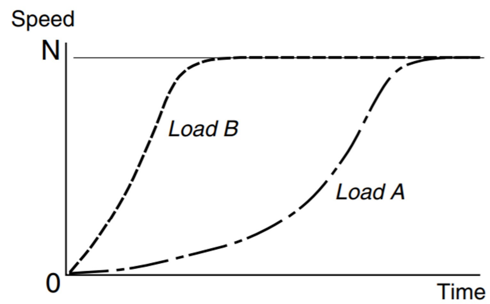


Figure 6.6 *Speed–time curves during run-up, for motor and loads shown in Figure 6.5*

Motor crawling:

Users should not be too alarmed as in most cases the motor will ride through the harmonic during acceleration, but in extreme cases a motor might, for example, stabilise on the seventh harmonic, and ‘crawl’ at about 214 rev/min, rather than running up to 4-pole speed (1500 rev/ min at 50 Hz), as shown by the dot in Figure 6.7.

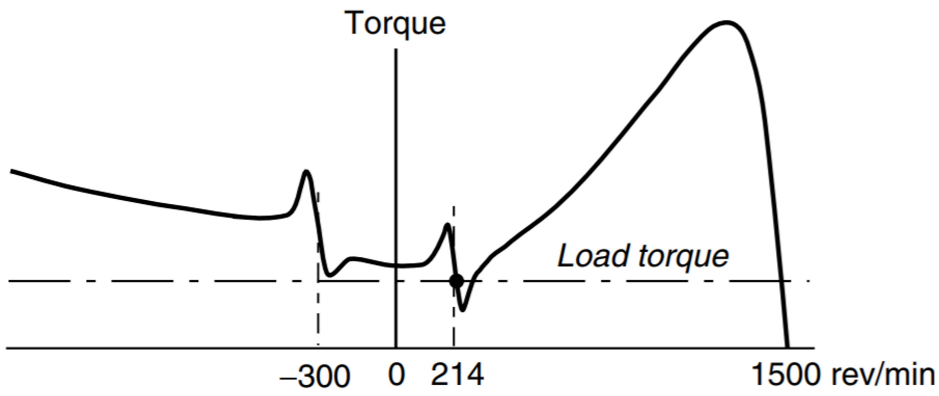


Figure 6.7 *Torque–speed curve showing the effect of space harmonics, and illustrating the possibility of a motor ‘crawling’ on the seventh harmonic*

To minimise the undesirable effects of space harmonics the rotor bars in the majority of induction motors are not parallel to the axis of rotation, but instead they are skewed (typically by around one or two slot pitches) along the rotor length. This has very little effect as far as the fundamental field is concerned, but can greatly reduce the response of the rotor to harmonic fields.

Rotor efficiency

$$\eta_r = \frac{\text{Mechanical output power}}{\text{Rated power input to rotor}} = (1 - \mathrm{s})$$

This result is very important, and shows us immediately why operating at small values of slip is desirable. With a slip of 5% (or 0.05), for example, 95% of the air-gap power is put to good use. But if the motor was run at half the synchronous speed (s = 0.5), 50% of the airgap power would be wasted as heat in the rotor.

We can also see that the overall efficiency of the motor must always be significantly less than (1 - s), because in addition to the rotor copper losses there are stator copper losses, iron losses and windage (air resistance) and friction losses. This fact is sometimes forgotten, leading to conflicting claims such as ‘full-load slip = 5%, overall efficiency = 96%’, which is clearly impossible.

TORQUE–SPEED CURVES – INFLUENCE OF ROTOR PARAMETERS

- For small values of slip, i.e. in the normal running region, the lower we make the rotor resistance the steeper the slope of the torque–speed curve becomes, as shown in Figure 6.9.

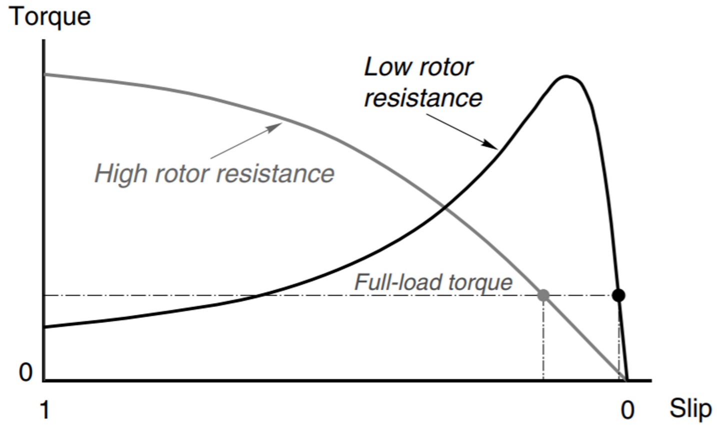


Figure 6.9 *Influence of rotor resistance on torque–speed curve of cage motor. The full-load running speeds are indicated by the vertical dotted lines*

We can see that at the rated torque (shown by the horizontal dotted line in Figure 6.9) the full-load slip of the low-resistance cage is much lower than that of the high-resistance cage. But we saw earlier that the rotor efficiency is equal to (1 - s), where s is the slip. So, we conclude that the low-resistance rotor not only gives better speed holding, but is also much more efficient. There is of course a limit to how low we can make the resistance: copper allows us to achieve a lower resistance than aluminium, but we can’t do anything better than fill the slots with solid copper bars.

As we might expect there are drawbacks with a low-resistance rotor. The starting torque is reduced (see Figure 6.9), and worse still the starting current is increased. The lower starting torque may prove insufficient to accelerate the load, while increased starting current may lead to unacceptable volt drops in the supply. Altering the rotor resistance has little or no effect on the value of the peak (pullout) torque, but the slip at which the peak torque occurs is directly proportional to the rotor resistance.

To sum up, a high-rotor resistance is desirable when starting and at low speeds, while a low resistance is preferred under normal running conditions. To get the best of both worlds, we need to be able to alter the resistance from a high value at starting to a lower value at full speed. Obviously we cannot change the actual resistance of the cage once it has been manufactured, but it is possible to achieve the desired effect with either a ‘double cage’ or a ‘deep bar’ rotor. Manufacturers normally offer a range of designs, which reflect these trade-offs, and the user then selects the one which best meets his particular requirements.

INFLUENCE OF SUPPLY VOLTAGE ON TORQUE–SPEED CURVE

Consider the torque–speed curves for a cage motor shown in Figure 6.14. The curves (which have been expanded to focus attention on the low-slip region) are drawn for full voltage (100%), and for a modestly reduced voltage of 90%. With full voltage and fullload torque, the motor will run at point X, with a slip of say 5%. Since this is the normal full-load condition, the rotor and stator currents will be at their rated values. Now suppose that the voltage falls to 90%. The load torque is assumed to be constant so the new operating point will be at Y. Since the air-gap flux density is now only 0.9 of its rated value, the rotor current will have to be about 1.1 times rated value to develop the same torque, so the rotor e.m.f. is required to increase by 10%. But the flux density has fallen by 10%, so an increase in slip of 20% is called for. The new slip is therefore 6%. The drop in speed from 95% of synchronous to 94% may well not be noticed, and the motor will apparently continue to operate quite happily. But the rotor current is now 10% above its rated value, so the rotor heating will be 21% more than is allowable for continuous running.

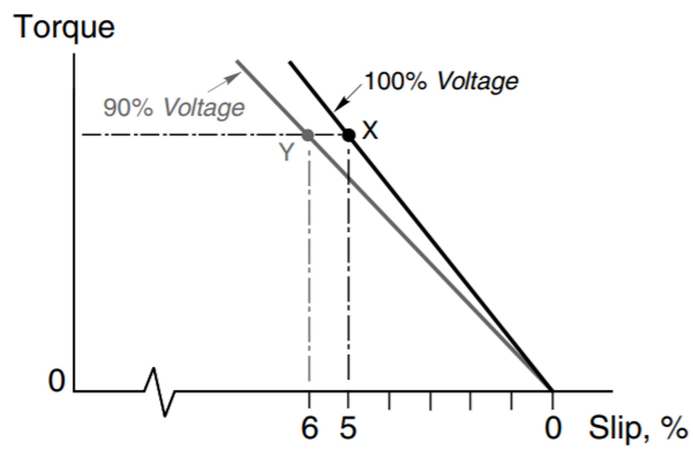


Figure 6.14 *Influence of stator supply voltage on torque–speed curves*

GENERATING AND BRAKING

We can see from Figure 6.15 that the decisive factor as far as the direction of the torque is concerned is the slip, rather than the speed. When the slip is positive the torque is positive, and vice versa. The torque therefore always acts so as to urge the rotor to run at zero slip, i.e. at the synchronous speed. If the rotor is tempted to run faster than the field it will be slowed down, whilst if it is running below synchronous speed it will be urged to accelerate forwards.

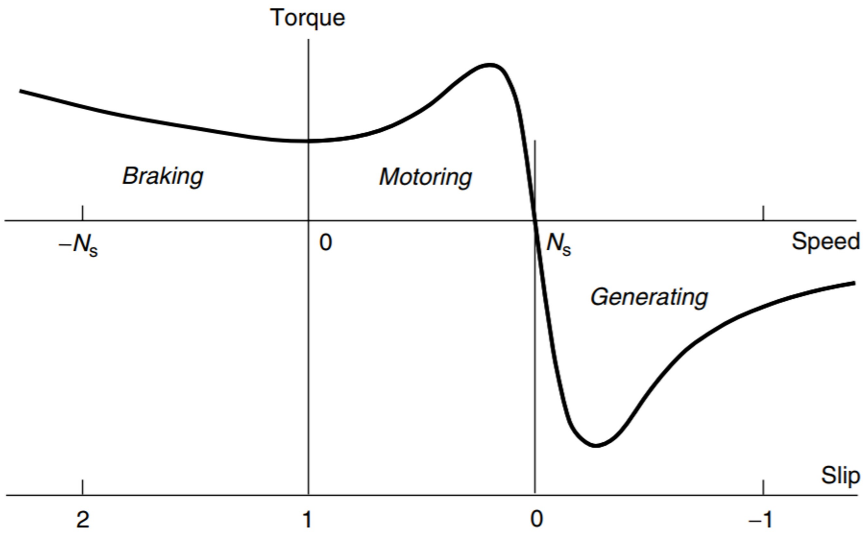


Figure 6.15 *Torque–speed curve over motoring region (slip between 0 and 1), braking region (slip greater than 1) and generating region (negative slip)*

Consider a cage motor driving a simple hoist through a reduction gearbox, and suppose that the hook (unloaded) is to be lowered. Because of the static friction in the system, the hook will not descend on its own, even after the brake is lifted, so on pressing the 'down' button the brake is lifted and power is applied to the motor so that it rotates in the lowering direction. The motor quickly reaches full speed and the hook descends. As more and more rope winds off the drum, a point is reached where the lowering torque exerted by the hook and rope is greater than the running friction, and a restraining torque is then needed to prevent a runaway. The necessary stabilising torque is automatically provided by the motor acting as a generator as soon as the synchronous speed is exceeded, as shown in Figure 6.15. The speed will therefore be held at just above the synchronous speed, provided of course that the peak generating torque (see Figure 6.15) is not exceeded.

When the rotor rotates over synchronous frequency, the slip becomes negative. In this case, the induced torque also becomes negative and the induction machine operates as a generator. However, since the machine itself cannot run over synchronous speed, it has to be accelerated by a prime mover and, this way, it draws power from it. When working as a generator, induction machine provides active power to the electrical source. However, due to a single source of excitation and, hence, lack of a field excitation on the rotor (this was the case in synchronous machines), it still needs to draw the reactive magnetizing current (Im). (From Ali Emadi - Advanced electric drive vehicles, p179 pdf)

Injection braking

This is the most widely used method of electrical braking. When the 'stop' button is pressed, the 3-

phase supply is interrupted, and a d.c. current is fed into the stator via two of its terminals. The d.c. supply is usually obtained from a rectifier fed via a low-voltage high-current transformer. We saw earlier that the speed of rotation of the air-gap field is directly proportional to the supply frequency, so it should be clear that since d.c. is effectively zero frequency, the air-gap field will be stationary. We also saw that the rotor always tries to run at the same speed as the field. So, if the field is stationary, and the rotor is not, a braking torque will be exerted.

***By using power electronic inverters, stator frequency of an induction motor can be changed and as shown in Figure 5.70, this translates the torque/speed curve along the speed axis. At speeds lower than the base speed, if only the frequency is reduced, this increases the lux, causes the motor to operate in the nonlinear region and, hence, leads to an increase in the magnetization current. To keep the lux constant, the stator voltage should be linearly decreased with the stator frequency. If voltage-to-frequency (V/f) ratio of an induction machine is kept constant, the breakdown torque remains constant and the torque-speed curve is shifted along the speed axis. ---> The pullout torque can be set at whatever speed?!

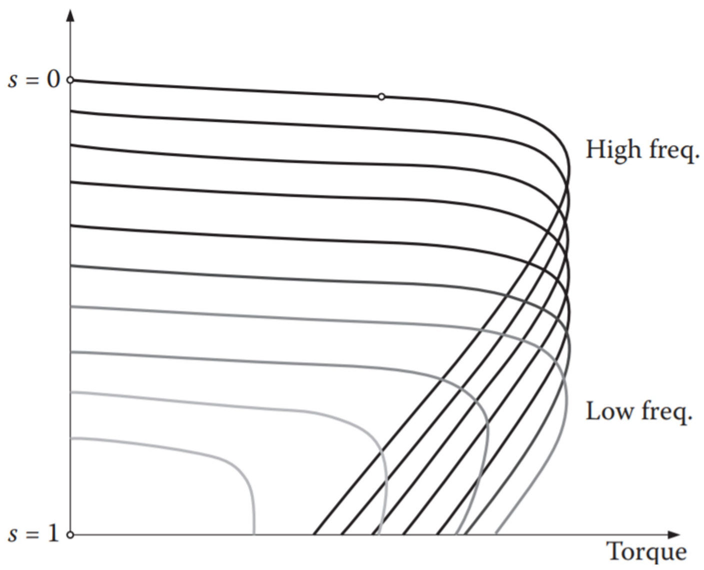


FIGURE 5.70 Typical torque-speed characteristics of an induction motor for variable-frequency operation.

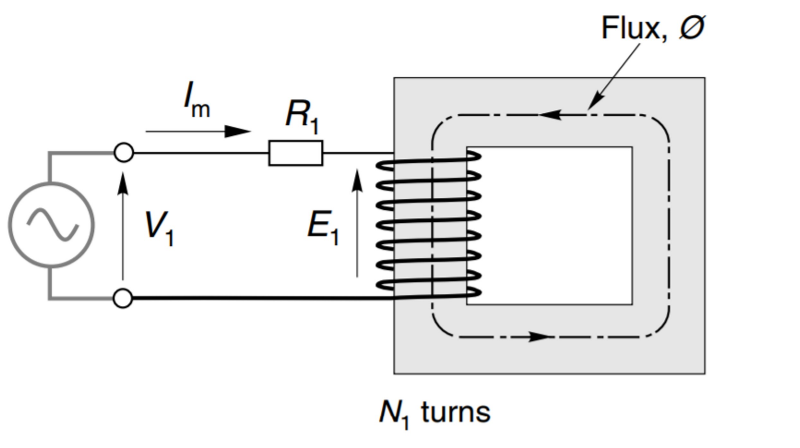
Chapter 7: INDUCTION MOTOR EQUIVALENT CIRCUIT

Transformer equivalent

The transformer equivalent is created in order to have an easier to understand equivalent to the IM.

Analysis without secondary winding

The analysis is started by the transformer without secondary winding (or with the secondary open circuited), if it were an IM, it would be an IM without rotor.



$$V_1 = E_1 + I_m R_1$$

$$E_1 = N_1 \frac{d\phi}{dt}$$

$$\frac{d\phi}{dt} = \frac{E_1}{N_1} = \frac{V_1}{N_1}$$

As mentioned above, we normally aim to keep the peak flux constant in order to fully utilise the magnetic circuit, and this means that changes to voltage or frequency must be done so that the ratio of voltage to frequency is maintained.

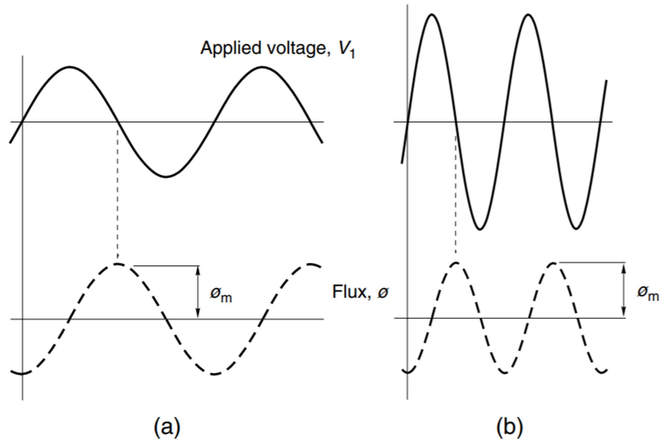


Figure 7.5 *Flux and voltage waveforms for ideal transformer operating with sinusoidal primary voltage. (The voltage and frequency in* (b) *are doubled compared with* (a), *but the peak flux remains the same.)*

For example, suppose we have a transformer core with a cross-sectional area 5 cm × 5 cm, and we decide we want to use it as a 240 V, 50 Hz mains transformer. How many turns will be required on the primary winding?

We can assume that, as discussed in Chapter 1, the flux density in the core will have to be limited to say 1.4 T to avoid saturation. Hence the peak flux in the core is given by

$$\phi_m = B_m \times A = 1.4 \times 0.05 \times 0.05 = 3.5 \times 10^{-3}\ \text{Wb} = 3.5\,\text{m Wb}$$

The peak voltage ($\hat{V}$) is the r.m.s (240) multiplied by $\sqrt{2}$; the frequency (f) is 50, so we can substitute these together with ϕ_m in equation (7.5) to obtain the number of turns of the primary winding as

$$N_1 = \frac{240\sqrt{2}}{2\pi \times 50 \times 3.5 \times 10^{-3}} = 308.7\ \text{turns}$$

We cannot have a fraction of a turn, so we choose 309 turns for the primary winding. If we used fewer turns the flux would be too high, and if we used more, the core would be under-utilised.

Viewed from the supply, the ideal transformer at no-load looks like an open circuit, as it draws no current. We will see later that a real transformer at no-load draws a small current, lagging the applied voltage by almost 908, and that from the supply viewpoint it therefore has a high inductive reactance, known for obvious reasons as the 'magnetising reactance'. An ideal transformer is thus seen to have an infinite magnetising reactance.

Transformation of impedances from primary to secondary

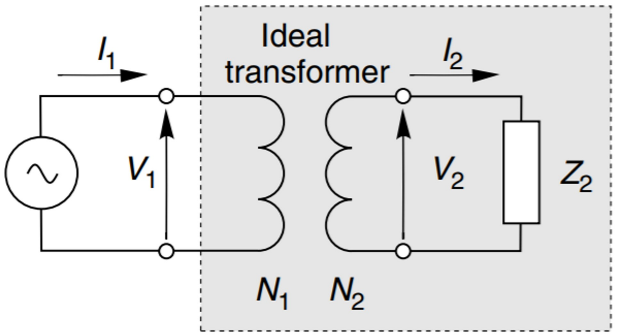
Assuming efficiency = 1 the next equation is easily derived:

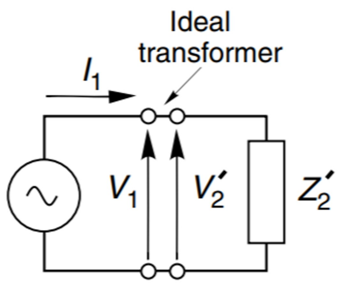
$$\frac{V_1}{V_2} = \frac{N_1}{N_2} = \frac{I_2}{I_1}$$

(It makes a lot of logical sense: The ideal transformer effectively 'scales' the voltages by the turns ratio and the currents by the inverse turns ratio)
Now from that equation the next set of relationships can be stablished:

$$\frac{V_1}{I_1} = \frac{(N_1/N_2)V_2}{(N_2/N_1)I_2} = \left(\frac{N_1}{N_2}\right)^2 \frac{V_2}{I_2} = \left(\frac{N_1}{N_2}\right)^2 Z_2 = Z_2'$$

Which can be used for transforming impedances from primary <-->secondary





Z2' is known as the referred impedance

The real transformer

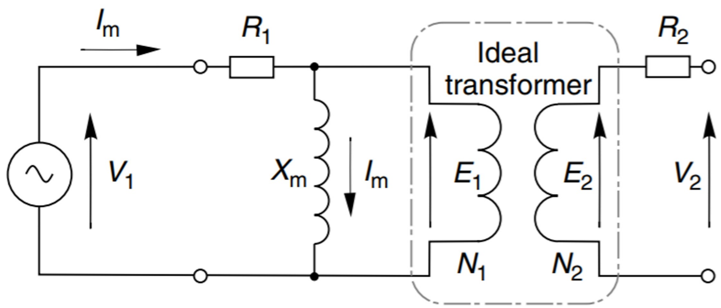


Figure 7.7 *Equivalent circuit of real transformer under no-load conditions, allowing for presence of magnetising current and winding resistances*

Magnetising reactance

The transformer in Figure 7.7 is shown without a secondary load i.e. it is under no-load conditions. As previously explained the current drawn from the supply in this condition is known as the magnetising current, and is therefore denoted by Im. In circuit theory terms, the magnetising reactance is simply the reactance due to the self-inductance of the primary of the transformer.

It turns out that when a transformer is operated at its rated voltage, the term Im*R1 is very much less than V1, so we can make the approximation V1 ~= E1, we can continue to say that the flux will be determined by the applied voltage, and it will not depend on the reluctance of the magnetic circuit (according to equation in the beginning). However, the magnetic circuit of the real transformer clearly has some reluctance, so it is to be expected that it will require a current to provide the MMF needed to set up the flux in the core. If the reluctance is R and the peak flux is fm, the magnetic Ohm's law gives:

$$\mathrm{MMF} = N_1 I_m = \mathcal{R}\,\phi_m, \quad \therefore\ I_m = \frac{\mathcal{R}\,\phi_m}{N_1}$$

The magnetising reactance X_m is given by $X_m = \dfrac{V_1}{I_m}$

In most transformers the reluctance of the core is small, and as a result the magnetising current (Im) is much smaller than the normal full-load (rated) primary current. The magnetising reactance is therefore high, and we will see that it can be neglected for many purposes. However, it was pointed out in Section 7.2 that the induction motor (our ultimate goal!) resembles a transformer with two air-gaps in its magnetic circuit. If we were to simulate the motor magnetic circuit by making a couple of saw-cuts across our transformer core, the reluctance of the magnetic circuit would clearly be increased because, relative to iron, air is a very poor magnetic medium. Indeed, unless the saw-blade was exceptionally thin we would probably find that the reluctance of the two air-gaps greatly exceeded the reluctance of the iron core.

We might have thought that making saw-cuts and thereby substantially increasing the reluctance would reduce the flux in the core, but we need to recall that provided the term Im*R1 is very much less than V1, the flux is determined by the applied voltage, and the magnetising current adjusts to suit, as given by equation (7.11). So when we make the sawcuts and greatly increase the reluctance, there is a compensating large increase in the magnetising current(Magnetising reactance value drops), but the flux remains virtually the same.

Eddy current losses

The final refinement of the no-load model takes account of the fact that the pulsating flux induces eddy current and hysteresis losses in the core:

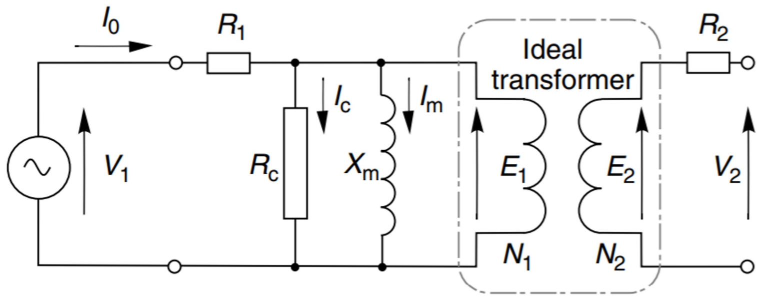


Figure 7.8 *Development of no-load equivalent circuit to include presence of losses in iron core*

The no-load current (I0) consists of the reactive magnetising component (Im) lagging the applied voltage by 90 degrees, and the core-loss (power) component (Ic) that is in phase with the applied voltage. The magnetising component is usually much greater than the core-loss component, so the real transformer looks predominantly reactive under open-circuit (no-load) conditions.

Leakage reactance (X1)

In the ideal transformer it was assumed that all the flux produced by the primary winding linked the secondary, but in practice some of the primary flux will exist outside the core and will not link with the secondary. This leakage flux, which is proportional to the primary current, will induce a voltage in the primary winding whenever the primary current changes, and it can therefore be represented by a ‘primary leakage inductance’ (l1) in series with the primary winding of the ideal transformer.

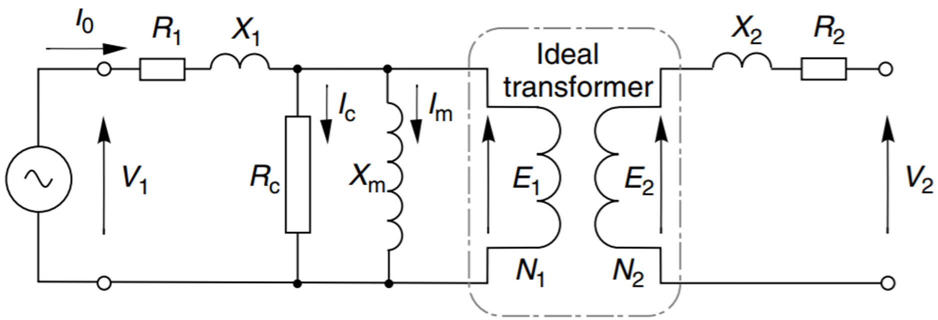


Figure 7.9 *Development of no-load equivalent circuit to include leakage reactances*

Because the leakage reactances are in series, their presence is felt when the currents are high, i.e. under normal full-load conditions, when the volt-drops across X1 and X2 may not be negligible, and especially under extreme conditions (e.g. when the secondary is short-circuited) when they provide a vitally important current-limiting function.

Real transformer on load – exact equivalent circuit

We saw that an impedance Z2 connected to the secondary could be modelled by a referred impedance Z2' connected at the primary, where Z2' = (N1/N2)^2 * Z2. We can use the same approach to refer not only the secondary load impedance Z2 but also the secondary resistance and leakage reactance.

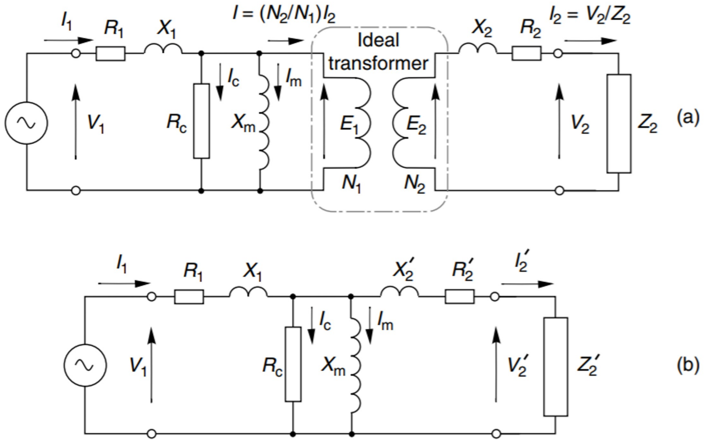


Figure 7.10 *'Exact' equivalent circuit of real transformer supplying load impedance* Z_2. *(The circuit in* (a) *includes the actual secondary parameters, while in* (b) *the parameters have been 'referred' to the primary side.)*

The first circuit is seldom used as it is easier to refer the secondary impedances to the primary side.
When circuit calculations have been done using this circuit the actual secondary voltage and current are obtained from their referred counterparts using the equations:

$$V_2 = V_2' \times \frac{N_2}{N_1} \quad \text{and} \quad I_2 = I_2' \times \frac{N_1}{N_2}$$

The principal effect of the non-zero series elements of the real transformer is that because of the volt-drops across them, the secondary voltage V2' will be less than it would be if the transformer were ideal. And the principal ffect of the parallel elements is that the real transformer draws a (magnetising) current and consumes (a little) power even when the secondary is open-circuited.

Simplified transformer circuit

Under normal conditions, the volt-drop across R1 and X1 will be a small fraction of the applied voltage V1. Hence the voltage across the magnetising branch (Xm in parallel with Rc) is almost equal to V1: so by moving the magnetising branch to the left-hand side, the magnetising current and the core-loss current will be almost unchanged. Moving the magnetising branch brings massive simplification in terms of circuit calculations. Firstly because the current and power in the magnetising branch are independent of the load current, and secondly because primary and referred secondary impedances now carry the same current (I2') so it is easy to calculate V2' from V1 or vice-versa. Further simplification results if we ignore the resistance (Rc) that models iron losses in the magnetic circuit, and we combine the primary and secondary resistances and leakage reactances to yield total resistance (RT) and total leakage reactance (XT) as shown in Figure 7.11(b). The iron losses will have to be considered in case that the efficiency of the machine is desired to be analysed.

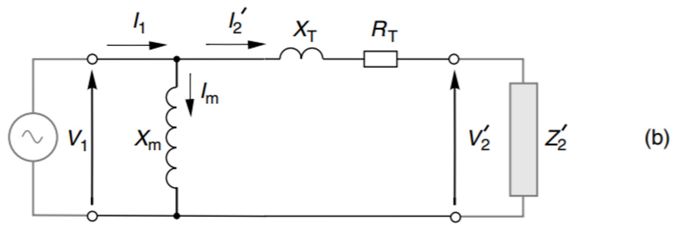


Figure 7.11 *'Approximate' equivalent circuit of real transformer*

As a transformer designer we then face a dilemma: if we do what seems obvious and aim to reduce the series impedance to improve the steady-state performance, we find that the short-circuit current increases and there is the problem of exceeding the permissible levels dictated by the supply authority. Clearly the so-called ideal transformer is not what we want after all, and we are obliged to settle for a compromise. We deliberately engineer sufficient impedance (principally via the control of leakage reactance) to ensure that under abnormal (short-circuit) conditions the system does not self-destruct! In Chapters 5 and 6, we mentioned something very similar in relation to the induction motor: we saw that when the induction motor is started direct-on-line it draws a heavy current, and limiting the current requires compromises in the motor design. We will see shortly that this behaviour is exactly like that of a short-circuited transformer

DEVELOPMENT OF THE INDUCTION MOTOR EQUIVALENT CIRCUIT

Stationary conditions

We can represent the motor at rest (the so-called 'locked rotor' condition) simply by setting Z2' = 0 in Figure 7.11. Given the applied voltage we can calculate the current and power that will be drawn from the supply, and if we know the effective turns ratio we can also calculate the rotor current and the power being supplied to the rotor.

Modelling the electromechanical energy conversion process

A very important observation in relation to what we are now seeking to do is that we recognised earlier that although the rotor currents were at slip frequency, their effect (i.e. their MMF) was always reflected back at the stator windings at the supply frequency. This suggests that it must be possible to represent what takes place at slip-frequency on the rotor by referring the action to the primary (fixed-frequency) side, using a transformer-type model; and it turns out that we can indeed model the entire energy-conversion process in a very simple way. All that is required is to replace the referred rotor resistance (R2') with a fictitious slip-dependent resistance (R2'/s) in the short-circuited secondary of our transformer equivalent circuit.

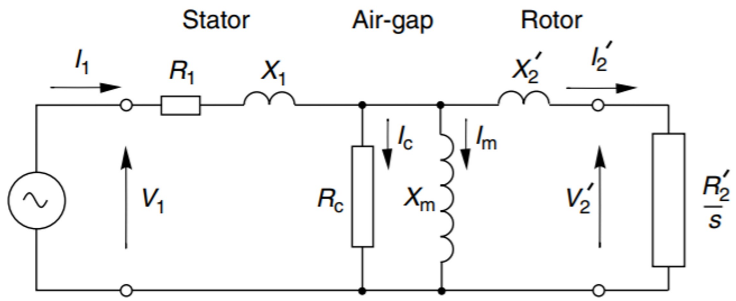


Figure 7.12 *'Exact' per-phase equivalent circuit of induction motor. The secondary (rotor) parameters have been referred to the primary (stator) side*

PROPERTIES OF INDUCTION MOTORS

Considering rotor copper losses (R2'), for the motor to be a good electromechanical energy converter, most of the power entering the circuit on the left must appear in the electromechanical load resistance. This is equivalent to saying that for good performance, the output voltage V2' must be as near as possible to the input voltage, V1.

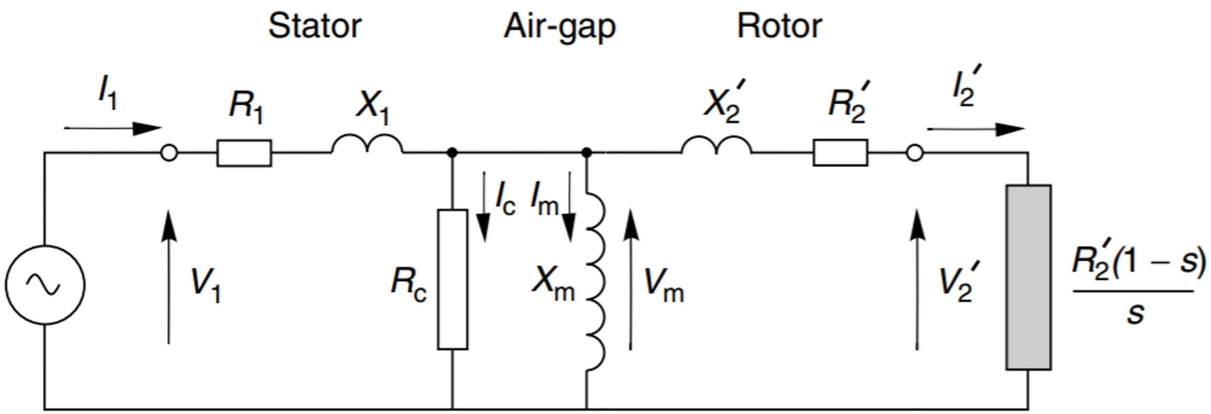


Figure 7.13 *Exact equivalent circuit with effective rotor resistance (R_2'/s) split into R_2' and $R_2'((1-s)/s)$. The power dissipated in R_2' represents the rotor copper loss per-phase, while the power in the shaded resistance $R_2'((1-s)/s)$ corresponds to the mechanical output power per-phase, when the slip is s*

Power balance

The power balance for the rotor can be derived as follows: Power into rotor (P2) = Power lost as heat in rotor conductors + Mechanical output power

$$P_2 = (I_2')^2 R_2' + (I_2')^2 R_2' \left(\frac{1-s}{s}\right) = (I_2')^2 \frac{R_2'}{s}$$

$$\text{Power lost in rotor heating} = sP_2$$

$$\text{Mechanical output power} = (1-s)P_2$$

$$\text{Rotor efficiency } (P_{\text{mech}}/P_2) = (1-s) \times 100\%$$

These relationships show that of the power delivered across the air-gap fraction s is inevitably lost as heat, leaving the fraction (1 s) as useful mechanical output. Hence an induction motor can only operate effciently at low values of slip.

Torque

We know that mechanical power is torque times speed, and that when the slip is s the speed is (1 s) vs, where vs is the synchronous speed. Hence from the power equations above we obtain:

$$\text{Torque} = \frac{\text{Mechanical power}}{\text{Speed}} = \frac{(1-s)P_2}{(1-s)\omega_s} = \frac{P_2}{\omega_s}$$

All of the relationships derived in this section (Torque and Power) are universally applicable and do not involve any approximations.

In the book there's an example of analysing a motor starting at page 284 'PERFORMANCE PREDICTION – EXAMPLE' **and might be good for get comfortable with the motor.**