5 Magnetic Circuits and Permanent Magnets

Having specified the main magnetic and electric dimensions of our motor design, it is necessary to determine the motor equivalent-circuit parameters so that its performance could be predicted and checked to see if it met the customer specification. Magnetic saturation is an important issue in that it may ruin our performance. In the example, we tried to keep away from it using sensible choices of the flux density. If we want a Permanent Magnet motor we also need to know how to set the level of B to match our design.

5.1 Magnetic Circuits

5.1.1 Magnetic Reluctance

Ampere’s Law:

\[ \int Hdl = NI = \sum \phi \frac{dL}{\mu A} = \phi \sum \frac{L_m}{\mu_m A_m} \text{ for } m \text{ sections} \]

The reluctance \( R_m \) of block of magnetic material of length \( L_m \), and cross-sectional area \( A \) is then defined:

\[ R_m = \frac{L}{\mu A} \quad (5.1) \]

where \( \mu \) is the magnetic permeability of the material. By analogy with electrical circuits

\( NI \) is then equivalent to Voltage and flux \( \phi \) is equivalent to current.

We can now create a magnetic equivalent circuit that is analogous to an electrical circuit. Resistances in the magnetic circuit are called magnetic reluctances and are defined in a similar way to resistances in an electrical circuit (look at the dimensions).
5.1.2 Magnetising Current

Now the magnetic circuit of the machine can be divided into five components through which the magnetic flux passes:

1. stator core;
2. stator teeth;
3. Airgap;
4. rotor teeth; and
5. rotor core.

The magnetic flux passes through these elements in series.

![Diagram of magnetic circuit](image)

**Fig. 5.1**

Parallel branches of reluctance model the splitting of flux flowing in the stator and rotor cores. Therefore, the total magnetic reluctance is

\[ R_T = \]

In this case, the amp-turns driving the magnetic flux is the effective stator turns times the magnetising current. Therefore, by estimating the reluctance of each part of the machine, we can find the magnetising current. The value of \( \mu \) will depend on the magnetic saturation of each part of the machine. Since we know the flux density levels throughout the machine we can calculate \( \mu \) from the B-H characteristic of the electrical steel we are using.

\[ \mu = \frac{dB}{dH} \]

Take the slope at the rms value of B for a good estimate.
Fig. 5.2  4% Silicon Iron Motor Laminations

The total magnetic reluctance is therefore

\[ R_T = \]

The magnetic flux flowing around the magnetic circuit can be obtained from
the specific magnetic loading

\[ \phi = \frac{\pi d}{2p} B = \]

The effective magnetising amp turns can be shown by Fourier analysis of the
mmf waveforms to be

\[ NI = \frac{12}{\pi^2 p} N p h k_w I_{m\text{rms}} = \phi R_T \]

The magnetising current is therefore

\[ I_{m\text{(rms)}} = \]

Magnetic equivalent circuits are a simple way of accounting for magnetic
saturation. If we were to assume infinitely permeable steel and ignore
saturation, the magnetic circuit shown in Figure 3.4 would reduce to only
the air-gap reluctance \( R_g \).
The magnetising current for infinitely permeable steel would be

Clearly magnetic saturation has a significant influence on the magnetising current. This approach also allows the designer to see quickly if he has over-saturated any particular part of the machine (e.g. the teeth) and adjust the machine dimensions if necessary.

5.1.3 Iron Losses

Iron losses are estimated for each part of the machine, and curves are used to give the losses. Note that the curves are a specific loss, so the teeth with a high magnetic flux density or ‘polarisation’ have high losses, in our example, but have a small volume.

Fig. 5.3 Linton and Hirst Lamination data (semiprocessed Electrical Steel)
5.2 Permanent Magnet Materials

Over the past thirty years, the performance of permanent magnet materials has risen dramatically with the introduction of ‘Rare Earth magnets’ such as Samarium-Cobalt and Neodymium-Iron-Boron. From a simplistic point of view, magnets can be used in electrical machines as a replacement for conventional dc field winding. The advantage being that they require no external supply or removal of copper-losses from the winding. However until relatively recently, the most commonly used magnet materials were ferrite compounds and they were only capable of producing low flux density levels (e.g. 0.1 Tesla). In terms of specific magnetic loading this meant that for a given required output, an electric motor employing ferrite magnets had to be physically larger (S = Speed x Volume x B x J) than its counterpart utilising a conventional dc coil winding. However, rare earth magnet compounds are capable of sustaining flux density levels up to approximately 1 Tesla and as a result there is a lot of development work on permanent-magnet machines, up to and above 1000 kW (wind generators).

5.2.1 Material Properties

Figure 5.3 illustrates the B-H characteristic of a permanent-magnet. Initially the material is unmagnetised and will subsequently be subjected to a high magnetic field which drives the magnet along OA. The magnetising field is then removed and the magnet relaxes along AC. This point is known as the remanent flux-density of the material. If the magnet is then inserted into a magnetic circuit the operating point will move into the 2nd quadrant -point D -the exact position of which will be determined by the load line of the circuit. If external amp turns are applied in the opposite direction, the operating point will move further down the curve to R for example. If the external amp-turns are now removed the material 'recoils' along RS. Applying the external excitation again simply moves the operating point back to R again.

This loop is known as a minor loop which is normally thin enough to be modelled by a straight line with a slope $\mu_{rec} \mu_0$ where $\mu_{rec}$ is known as the recoil permeability.

In terms of permanent-magnet motor design, the load line is determined from the magnetic circuit, as above and the external excitation amp-turns are provided by a winding, if any. Clearly if the external excitation pushes the operating point past the knee of the curve, point P, then an irreversible loss of flux-density results. This point is known as the intrinsic coercivity $H_{ci}$. In reality machine designers would usually constrain the secondary maximum current to prevent operating past the point where B = 0 or the magnet coercivity $H_c$. 
The position of the load line is determined by the magnetic circuit of the machine, with the ‘secondary excitation’ winding open-circuit.

The product of the magnet’s B and H is termed the energy product. This quantity (which is not the actual energy stored in the magnet) is a measure of how hard the magnet is working to provide flux against the demagnetising effect of the external magnetic circuit. For the material shown above, the energy products are rectangular hyperbolas and clearly for the linear demagnetisation curve shown in Fig. 1 the position of \((BH)_{\text{MAX}}\) occurs at the mid-point, \(B=B_r/2\). It is common practice to design the magnetic circuit so that the load point corresponds to \(B=B_r/2\).

One usually assumes that the possibility of a short-circuit in the motor winding does not result in excessive currents leading to demagnetisation.
5.2.2 Magnet Types

There are four main types of permanent magnet in use today.

a) Alnico/Alcomax

b) ceramic Ferrites

c) Rare earth (e.g. Sm-Co)

d) Nd-Fe-B (neodymium-iron-boron)

Clearly we are mostly interested in the Left Hand Quadrant - the Demagnetisation curve, Fig. 5.4.

![Demagnetisation curves for popular magnet types.](image)

The rare earth magnet types are very attractive, with excellent straight line characteristics. There are issues with shaping them to suit a design - hence the polymer bonded types. Careful consideration of their maximum operating temperature is needed, but for machines the temperatures are usually acceptable. Cost is an interesting question, with the patents on rare earth magnets running out!

Alcomax is tricky to deal with in electrical machines and this type is now only used in various simple magnetic holding applications, particularly where its high temperature stability is needed\(^1\).

\(^1\) http://www.magnetsales.co.uk/magnets/castalcomax.html
5.2.3 Load Line Example

Ampere’s Law:

Flux is conserved:

The approximate magnet characteristic is

Solving for the operating point:

If we assume $A_g = A_m$, and that we wish to operate at the maximum energy product ($B_m = B_r/2$) then to minimise volume of magnet we must minimise the airgap length.

We are treating the magnet as a coil carrying a current, with a reluctance.
5.2.4 Choice of Magnet type

The most convenient method of comparing the magnetic performance of different types and grades of permanent magnet is to consider their maximum energy product (BHmax). This is the point where the magnet will provide most energy for the minimum volume, so:-

<table>
<thead>
<tr>
<th>Material</th>
<th>kJ/m³</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ferrite (anisotropic)</td>
<td>26</td>
</tr>
<tr>
<td>Samarium Cobalt (2:17)</td>
<td>208</td>
</tr>
<tr>
<td>Neodymium-Iron-Boron (N38H)</td>
<td>306</td>
</tr>
</tbody>
</table>

It is frequently necessary to know what flux density will be on the pole face of a magnet, for example to give us $\bar{B}$. The value will depend on the magnetic circuit. The following table shows typical pole face flux densities for the four grades when working at approximately their BH\text{max} points.

<table>
<thead>
<tr>
<th>Material</th>
<th>mT</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ferrite (anisotropic)</td>
<td>100</td>
</tr>
<tr>
<td>Samarium Cobalt (2:17)</td>
<td>350</td>
</tr>
<tr>
<td>Neodymium-Iron-Boron (N38H)</td>
<td>450</td>
</tr>
</tbody>
</table>

In our design methods then we will have to decide on permanent magnet material at the beginning of the process and then design the magnetic circuit fairly early on as part of the design process.

Shape, tolerances and quantity will influence the cost of individual magnets. These can be factored as follows:

<table>
<thead>
<tr>
<th>Material</th>
<th>Factor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ferrite (anisotropic)</td>
<td>Low(x1)</td>
</tr>
<tr>
<td>Samarium Cobalt 2:17)</td>
<td>Very High(x20)</td>
</tr>
<tr>
<td>Neodymium-Iron-Boron (N38H)</td>
<td>High (x10)</td>
</tr>
</tbody>
</table>

With current lamination types, and motor designs there is little to be gained from finding more powerful magnets.

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2 http://www.magnetsales.co.uk/application_guide/magnetapplicationguide.html
5.3 A Superconducting Motor?

Superconducting motors can develop the same torque and horsepower within a motor frame that is nearly a third the size of a comparably rated conventional motor, Whitcomb said. Losses can be half those of a conventional motor of equal power. The main factor leading to an HTS motor's smaller size for a given horsepower output is the magnetic-field strength that superconducting magnets create. Iron teeth, used to enhance magnetic flux in conventional rotors and stators, are not needed by superconducting motors, he said. "This means that the current densities in the active regions are not limited by iron saturation. The stator only requires the use of back iron acting primarily as a shield to keep magnetic flux inside the machine. A resulting HTS air core configuration, with higher flux density, is significantly smaller and lighter than conventional ac synchronous motors," Whitcomb said. Gamble, at American Superconductor, added, "Our estimates are that the active length of a superconducting motor will be on the order of one-quarter of that for a permanent magnet alternative.

The induction machine dwarfs a superconductor motor. The smaller machine, approximately 2.3m x 1.65m full scale, owes its svelte figure to no iron tooth loss and higher air gap flux density.

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http://www.memagazine.org/backissues/april00/features/hunt/hunt.html
6 Induction Motor Drives I

Three-phase cage-induction motor drives are very important in today's industry. In many applications requiring variable speed, it is desirable to use induction motors because they are very robust and relatively cheap to manufacture compared to other types of motor. In this lecture, we identify the main issues to be considered when using induction motors in drives.

6.1 Induction motor principles

6.1.1 The equivalent circuit of the induction motor.

![Per-phase equivalent circuit](image)

The familiar equivalent circuit is applicable to the steady state only.

6.1.2 Slip

The key feature of voltages induced in the rotor winding or bars depends on the rotor rotating at a different rate to the flux wave.

The flux induces voltages on the conductors in the rotor slots (think of the flux cutting method). The voltage pattern mirrors the ‘local’ magnitude of the flux, and the magnitude of the voltage also depends on the slip.
The slip in the induction motor is a measure of the difference between the speed of the rotating magnetic field and the speed of the rotor.

\[ s = \] (6.1)

From this definition we may write

\[ \text{(6.2)} \]

and so the rotor speed may be written as

\[ \text{(6.2)} \]

6.1.3 Physics of torque production using space vectors

The physical currents \((I)\), mmfs \((NI)\) and fluxes \((B)\) may be represented conveniently by ‘space phasors’ or ‘vectors’. For example the stator mmf is given by the stator currents as follows:

\[ i_a e^{j\theta_e} + i_b e^{j\theta_e+j2\pi/3} + i_c e^{j\theta_e+j4\pi/3} \]

Where \(\theta_e\) is the electrical space angle (or angle around stator in a 2 pole machine \(\theta_e = \theta / p\)) and the three sinusoidally distributed windings are at 0, 120, 240 degrees.

The space vector form of this is

\[ \vec{i_a} + \vec{i_b} + \vec{i_c} \]
For balanced 3 phase ac supplies, combining the 0, 120, 240 degree timing of the phase waveforms and the 0, 120, 240 degrees orientation of the three windings results in the single sinusoidally distributed stator \(NI\) which rotates and is always proportional to \(3/2\) times the peak magnitude of the current in any one phase (as long as the supplies are balanced).

Mathematically, a single rotating vector may represent the stator NI. Thus, a *rotating stator current vector* can be defined as the sum of the phase current vectors, which are fixed (and the magnitude adjusted):

\[
\vec{i}_s = \frac{2}{3}(\vec{i}_a + \vec{i}_b + \vec{i}_c)
\]

This produces the stator NI which interacts with the Rotor NI in the magnetic circuit of the machine to give us the single flux wave.

Similarly, vectors for the rotor NI, H and currents, the stator, rotor and resultant fluxes and the stator and rotor voltages may be described, and a transformer-like full transient model created. The maths as usually written is beyond our scope, but the idea of one vector variable coming from three components is invaluable and the steady state equivalent circuit then gives us genuine insight into the full transient behaviour.
6.1.4 Torque

Consider then the vector of flux. With slip, this produces a sinusoidal distribution of rotor voltages and currents in a predominantly resistive rotor circuit.

Mathematically, torque is given by the Vector Cross Product of the flux vector and either of the rotor or stator current vectors. In the steady state, the cross product is dealt with simply by considering the in-phase component of the rotor current vector. i.e.

\[ T \propto \vec{\psi} \times \vec{i}_r = \]

where \( \cos(\phi) \) is the power factor of the rotor circuit.
\[
\cos(\phi) = \frac{R_2}{\sqrt{(R_2)^2 + (s\omega L_2)^2}}
\]

The cross product of vectors is always valid and implies that the induction motor is capable of high performance - if we can control it!

6.1.5 Calculation of Torque

The steady-state electromagnetic torque developed by the induction motor may be found from the power balance.

Noting that there is also a torque balance and everything on the stator side happens in synchronism, the equation above becomes

Thus

The equivalent circuit is derived in this way.

\[
P_{gap} = \frac{V_2}{\sqrt{\left(\frac{R_2}{s}\right)^2 + (\omega L_2)^2}} \frac{R_2 s}{s} \frac{\left(\frac{R_2}{s}\right)^2 + (\omega L_2)^2}{\sqrt{\left(\frac{R_2}{s}\right)^2 + (\omega L_2)^2}}
\]

So we need only calculate the power dissipated in the component, \( R_2/s \), and divide by the synchronous speed.

Other ways of calculating:

\[
P_{gap} = 3V_2 I_2 \cos \phi = 3V_2 \frac{V_2}{\sqrt{\left(\frac{R_2}{s}\right)^2 + (\omega L_2)^2}} \frac{R_2 s}{s} \frac{\left(\frac{R_2}{s}\right)^2 + (\omega L_2)^2}{\sqrt{\left(\frac{R_2}{s}\right)^2 + (\omega L_2)^2}}
\]

Note the way \( s \) has moved!

\[
T =
\]
6.1.6 Torque-speed characteristic

For fixed stator voltage and fixed frequency, the standard industrial induction motor has the characteristic shown.

Note the quadrants and the shape of the characteristic. It is crucial to realise that the motor operates stably

This is similar to the dc motor characteristic!
6.2 Industrial Type Motors

Let's think about a typical induction motor, e.g., 415V delta connected 7.5 kW 4-pole motor:

\[ \begin{align*}
R_1 &= X_1 = \\
R_2 &= X_2 = \\
X_m &= \\
\text{Speed at rated torque} &= \text{slip} = \\
\text{At Full Load} &\quad R_2/s = 
\end{align*} \]

Check to see if at rated operation the series stator terms may be neglected.

The torque equation then becomes

To obtain the slope of the torque speed curve, where it operates, differentiate with respect to s.

To find the asymptotic slope, we want the value at s = 0.
Consider what this means on the torque speed curve:

\[ \omega L_2 \ll R_2/s. \]  

(6.5)

At a constant ratio of supply voltage and frequency, the slope is constant.

Variable frequency induction motor drives aim to keep in the linear part of the torque speed curve, so we can safely assume that \( \omega L_2 \ll R_2/s \).

**Steady-state operation, in the linear negative slope portion of the characteristic only, means the induction motor can be considered as having linear behaviour as long as a constant ratio of voltage to frequency is maintained. Industrial motors usually have a very linear characteristic, with a small slip, up to about 1.5X their full load torque.**
7 Induction Motor Drives II

Induction motors appear in a wide variety of drive systems, from the cheap and simple right through to very sophisticated systems. The cheapest drives are open loop, with a triac phase controller providing a variable voltage supply to the motor. At the other end is the flux-vector controlled drive, for closed loop servo applications. In this lecture we start with the basic methods and then focus on the variable voltage variable frequency (VVVF) open loop drive.

Consider again the slip equation:

Thus the speed may be varied by altering the supply frequency or the slip.

7.1 Speed control by changing the slip

Recall that only the linear part of the torque-speed curve is used. It's slope was given by (6.5)

To change the speed for a given torque we can adjust the voltage or the rotor resistance. The speed is a function of the load and at no load the motor still runs synchronously.

Clearly, by changing the rotor resistance, the slope of the stable portion is changed. Historically, added rotor resistance was used widely in induction motor speed control. This is quite possible in wound rotor machines which have slip rings for the rotor connections. It is not possible with cage rotor motors.
7.1.1 Speed control by varying the voltage

Here, the difficulty is the loss of torque in general, so this is very specific to fan and pump loads.

7.2 Speed control using a VVVF supply

It is quite clear that the most efficient and robust speed control method is based on adjusting the supply frequency.

The synchronous speed \( \omega/p \) is then changed.

7.2.1 Conditioning the system

In Variable Voltage Variable Frequency induction motor drives, the underlying assumption is that the motor operates in a steady state manner. In practice this is ensured by a ramp limiter on any of the input control variables, such that no rapid changes are allowed.

Thus the steady state analysis based on the idea of slip remains valid and no extremes of the torque speed curve are reached. The low pass filter on the demand input may also be represented as an integrator.
7.2.2 V/f control

In section 6.2 we saw that the voltage should be kept on ratio with the frequency, and we know we will then have constant flux. This can be seen in considering the equivalent circuit:

As the frequency is varied, the magnitudes of the reactive components vary.

In the conventional industrial IM, the leakage inductances are just 3 or 4 percent of the magnetising inductance: At no load, \( R_2/s \) is very large so we only need consider

At high frequencies, and reasonably large motors,

Then

the rated magnetising current can be obtained by controlling the supply voltage in proportion to the supply frequency, This is known as *constant V/f control.*
At low frequencies

7.2.3 Analysis of a VVVF IM drive

It follows then, that the performance of the IM with constant \( V/f \) applied may be understood by considering the equivalent circuit above, neglecting the small series stator terms, \( R_1 \) and \( L_1 \).

\[
V_1 = \text{(7.1)}
\]

The torque is given by equation 6.4

Applying equation 7.1 to equation 6.4, and rearranging

\[
\frac{T}{P} = 3k^2 \frac{s \omega R_2}{R_2^2 + s^2 \omega^2 L_2^2} \quad \text{(7.2)}
\]

Now the torque only depends on \( s \omega \), the slip frequency, i.e. the actual frequency seen by the rotor bars or winding

Zero torque is at \( s \omega = 0 \) (i.e. \( s=0 \)), i.e. when the rotor is rotating at the supply frequency. Mapping the new torque function to the speed axis gives an infinite set of curves similar to that at rated voltage and frequency.

Fig. 7.3
Note that it is NOT convenient to plot \( s \) on the set of torque speed curves. For a particular load torque, the value of \( s \omega \) is constant, so at different frequencies the value of \( s \) will be different.

7.2.4 Simplified analysis

Note that on the stable part of the torque-speed curve at rated torque and flux,

Since we are definitely only using the stable part, we can further simplify the induction machine model and equation 7.2:

\[
\frac{T}{p} = 3k^2 \frac{s \omega}{R_2} = 3p \frac{k^2}{R_2} (\omega_s - \omega_r) \tag{7.3}
\]

Hence

\[
\frac{dT}{ds\omega} = 3pk^2 \frac{1}{R_2} = -\frac{1}{p} \frac{dT}{d\omega_r}
\]

Compare this to equation 6.5. Plotting:

The intercept on the speed axis is at \( \omega_s = \omega/p \). The torque equation can be written in various forms while retaining the constant V/f relationship. Compare to the armature voltage controlled dc motor (with constant field flux).

\[
T_{dc} = \frac{(k\phi)^2}{R_a} (\omega_b - \omega)
\]
### 7.3 VVVF Drive with pwm inverter

The basic three phase switched mode dc-ac inverter bridge is supplied by a dc source.

![Fig. 7.5](image)

The dc supply may be batteries or an ac-dc rectifier with capacitor smoothing.

#### 7.3.1 Inverter Modulation methods (*Not in Exam*)

A variety of *modulation strategies* are possible. An attractive strategy for induction motor drives is *Space Vector Modulation*.

![Fig. 7.6](image)

A variable voltage variable frequency ac waveform is produced fairly easily by chopping up a dc voltage. The three phase concept is almost universal.
7.4 DEMONSTRATION

The inverter drive is as shown in the lecture. It clearly demonstrates open-loop speed control through variation of the supply frequency.

The inverter drive demonstrates the ramp limiter

It also demonstrates gear changing in the modulation where the switching frequency stays in a small range around a set frequency.
8 Induction Motor Drives III

So far we have covered the IM operation in a drive. Clearly the both the IM and the power electronic converter impose limits on the operation. This lecture the operating limits of the drive are explained for speeds up to the rated speed of the motor and beyond and some typical schemes will be described.

8.1 Operating limits

8.1.1 V/f below base speed

*Base speed* ($\omega_b$) is the speed (rads s$^{-1}$) of the motor as inferred from the information on the name-plate of the motor. With a typical motor, supplied at its rated (or base) voltage ($V_b$) at its ‘rated’ or base frequency, it is the maximum speed upto which rated flux can be maintained. Since at no load the motor operates at close to synchronous speed ($\omega_s$ rads s$^{-1}$).

Then

$$k_{\text{rated}} =$$

The inverter has a maximum output voltage, and this is usually closely matched to the IM, such that it can run up to base speed with full flux. So $V_{\text{max}} = V_b$.

8.1.2 Rotor current limit

The rotor current for a particular torque (and $s$) is given by

$$I_2 = \frac{V_1}{\sqrt{\frac{R_2^2}{s^2} + \omega^2 L_2^2}}$$

(ignoring series stator impedances as before).

With constant V/f, and rearranging

$$I_2 = \frac{s\omega k}{\sqrt{R_2^2 + s^2 \omega^2 L_2^2}}$$

(8.2)

A particular torque requires a unique value of $s\omega$, regardless of $\omega$, hence $I_2$ is constant for a given torque, independent of frequency.
Maximum rotor current depends on cooling and is therefore known at base speed, but not at other speeds, unless the motor is *through ventilated* - where an external blower blows air through the motor. This is common on large motors used in a VVVF drive. In a *totally enclosed* motor, the rotor cools via the stator, with a simple fan on the shaft as fixed speed motors often have. In these cases there are restrictions on the duty of the motor drive. The mass of the motor has a specific heat capacity and acts to smooth out temperature variations (Later lecture).

### 8.1.3 Stator current limit

This is imposed by the copper losses in the stator winding (the stator heating limit). The stator current is given by

\[
\bar{I}_1 = \bar{I}_2 + \bar{I}_m
\]

(8.3)

For operation at rated flux density, the magnetising current \( I_m \) is at its nominal, rated value (V/f constant) and is reactive. Eqn. 8.2 indicates that the rotor current is resistive and maximum stator current occurs when the rotor current is also at a maximum.

Thus the stator current maximum is defined along with the rotor current maximum. It does not add any further constraint.

### 8.1.4 Voltage boosting at low \( \omega \)

As previously mentioned, at a low frequency the flux will tend to be lower than desired due to the voltage lost across the stator resistance \( R_1 \), particularly at high torque.

The following might be considered

\[
V = k\omega_s + c
\]

But, heavily over fluxes the IM at low load torques, at low speeds. Clearly a feedback system would be better.
A simple solution is to simply fix the minimum voltage at low frequencies including at zero.

Hence the following curve is used:

![Fig. 8.1](image)

The voltage boost is often set as a commissioning parameter, for the maximum load to be started.

### 8.1.5 Operation above base speed - Field Weakening

It is usual to operate beyond the base frequency. Usually the rated voltage of the IM should not be exceeded and the inverter also has a voltage limit. Higher than base speed is therefore accommodated by allowing the IM to have a lower than rated flux, by allowing the V/f ratio to reduce, when above $\omega_b$. This is known as *field weakening*.

The Torque available is reduced:

1. Reduced flux @ maximum rotor current

2. Maximum possible torque @ reduced flux

Considering condition 1,

$$T = 3k^2 \frac{s \omega}{R_2}$$

Now $V_1$ is fixed at $V_b$. So

$$T = 3 \frac{V_b}{\omega} \frac{2}{R_2} s$$

Recall also

$$I_2 = \frac{sV_1}{\sqrt{R_2^2 + s^2 \omega^2 L_2^2}} = (8.4)$$
Substituting into Eqn. 8.4 the maximum voltage and current, a maximum value of $s$ will now apply (rather than a max value of $s\omega$).

$$s_{\text{MAX}} =$$

This corresponds to the original base value of $s$ Hence,

$$T =$$

The max torque rolls off with speed when field weakening.

This gives a constant power out, so field weakening is sometimes called ‘operation at constant power’.

The slope of the torque speed curve becomes shallower too!

$$\frac{dT}{d\omega_r} = -3\frac{V_b^2}{\omega^2} \frac{1}{R_2} \quad (8.5)$$

Condition 2 takes over when the value of $s\omega L_2$ approaches $R_2$ and our assumptions fail.

With the voltage fixed and the frequency increasing, it has a similar effect on the curves as fixed frequency, reduced voltage operation. Eventually we hit the peak of the torque speed curve.

The IM can't produce rated torque so $s < s_b$ ($\omega > \omega_b$). So only the voltage limit applies and the maximum torque is simply the peak torque. This portion is rarely used as it is hopelessly inefficient.
8.1.6 Torque speed curves with operating limits

These may now be sketched, with the limits applied:

Fig. 8.2

The limits are mirrored in each axis to cover all 4 quadrants if the power electronic converter allows.
8.2 Open Loop VVVF systems

8.2.1 Basic open loop drive

- Very simple, possibly single micro-chip solution
- Cheap
- No regenerative braking (rectifier front end)
- Simple protection trips on current and possibly voltage

Two feedback loops of the limit kind can be seen:

The measured stator current is compared with the preset limit. During motoring, if it is greater than the limit, the magnitude of the slip frequency is reduced slightly, by slightly reducing the applied frequency.

During braking, power flow through the inverter is reversed and so is negative. When using an uncontrolled diode rectifier, no energy may be returned to the supply. Consequently, the energy recovered from the motor tends to charge up the DC link capacitors. So the schemes always monitor the DC link voltage, and reduce the rate of deceleration until the link voltage stops increasing.
8.2.2 Open loop with slip compensation

Fig. 8.4

- Closed loop on current to estimate the load torque.
- Compensates for the loss of speed with Torque by increasing the applied frequency.
- Cheap current sensors available, with no long wires.
- Voltage and current limits as always.

This method is based on the simple model of the motor we developed in equation 7.3.

This scheme recognises that the induction motor (or asynchronous motor in North America) runs very near synchronism. The scheme works in both rated flux and field-weakening regions, motoring and braking, although the relationship between torque and \( s \omega \) is not constant in the field weakening region.

Open loop on speed schemes may be attractive in harsh environments such as oil rigs. For heavy braking, a *dynamic brake resistor* may be added to the dc link with a switch to bring it in if needed.
8.2.3 Closed loop slip frequency controlled

- Closed outer speed loop including the induction motor.
- Controller is typically PI and must have a limiter function.
- Tuning of the PI controller is important. The formal stability analysis is very difficult, so the PI parameters are set up as part of commissioning.
- The slip frequency limiter will apply the torque/current limit.
- Usually requires a dynamic braking resistor on the dc link.

The dynamic performance is poor as the controller needs to ramp limit the input, to avoid moving away from equation 7.3, but the speed is held under variable load conditions.

Fig. 8.5
8.3 Vector controlled

Reference was made in Lecture 5 to the *Space Vector* representation of the mmfs, fluxes and currents. These can then be used to formulate a model for the induction motor based entirely on vector quantities.

\[ T = \overline{\psi} \times \overline{I} \]

This is for instantaneous vector quantities. Therefore a control system to give the high performance possible with an induction motor will be based on the vector model. This needs realtime mathematics and control loops on the Im and I₂ (flux and torque producing current) vectors (beyond our scope).

Here we will simply look at the behaviour of a working model (in PSPICE). Fig. 8.6 shows an inertial load being accelerated. Note the nice step in torque with no overshoot and the way that the rotor current similarly ramp up and then assume the usual constant so we expect of the equation 7.3. Meanwhile the stator currents (measured through the voltage sources) have a constant magnitude but increasing frequency.

Fig 8.6

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