Chapter 3: Magnetic Materials

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Abstract – this chapter examines the magnetic properties and analysis of common soft and hard magnetic materials used in electric machines. It includes concepts such as magnetic circuit analysis, back-emf, iron losses, stacking factor and demagnetisation.

I. Introduction

There are two key materials used in electric machines: electrical conductors and magnetic materials. Electrical conductors were discussed in the previous chapter and this chapter introduces soft and hard magnetic materials and their properties and analysis. Fig. 1 compares their properties with air.

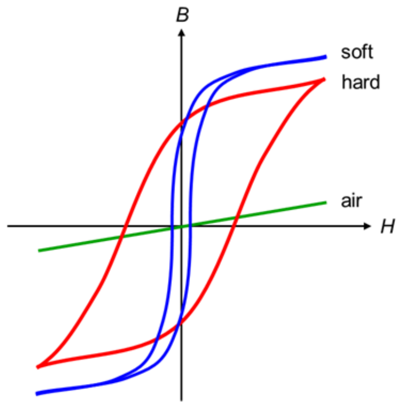

Fig. 1. Typical *BH* curves for soft and hard magnetic materials and air.

Soft magnetic materials, such as iron, are materials in which the magnetic flux density *B* can be easily changed by applying small values of magnetic field strength *H*. Hard magnetic materials, also known as permanent magnets, are materials in which the magnetic flux density *B* is less affected by the magnetic field strength *H*. The variation of magnetic flux density with magnetic field strength for a given material is called the BH curve.

II. Soft Magnetic Materials: Properties and Magnetic Analysis

A. Key Properties of Soft Magnetic Materials

The key requirements of soft magnetic materials in electric machines are to carry high values of AC magnetic flux with minimal amount of magnetomotive force (mmf) and power losses (called "iron" losses).

The magnetic flux Φ in Webers (Wb), in a magnetic conductor of cross-section area *A* which has a magnetic flux density *B* in Telsa (T) is given by,

$$\Phi = BA \quad [\text{Wb}] \qquad (1)$$

The maximum operating flux density in a soft magnetic material is called its saturation flux density, B_{sat}, and it is desirable that this be as large as possible. This allows an electric machine to be smaller for a given torque rating.

The magnetomotive force, mmf *NI* in ampere-turns (At) or amperes (A) in a magnetic conductor with a magnetic field strength *H* in A/m, permeability μ and average magnetic flux path length *l* is given by,

$$B = \mu H = \frac{\mu NI}{l} = \frac{\mu_R \mu_0 NI}{l} \quad [\text{Wb}] \qquad (2)$$

As was seen in the previous chapter, the permeability μ can be found from the relative permeability μ_R and the permeability of free space μ_0 ($4\pi \times 10^{-7}$H/m) using the equation $\mu = \mu_R\mu_0$. Re-arranging (2) gives the required mmf for a given flux density as,

$$NI = \frac{B \bullet l}{\mu_R \mu_0} \quad [\text{At}] \text{ or } [\text{A}] \qquad (3)$$

Thus to minimize the mmf *NI* it is desirable to have a magnetic material with a high permeability. This minimizes the current in electric machines required to produce the desired flux density in the airgap (called the magnetising current) and is particularly important for induction and reluctance machines.

Finally it is desirable that the magnetic material carry magnetic flux with low "iron" losses, which are a combination of hysteresis and eddy-current losses. These losses are a function of both the magnetic flux density *B* and the operating frequency *f* and have units of W/kg. These losses are discussed further in Section III. To minimise eddy-current losses, magnetic materials in electrical machines are normally constructed of sheets (laminations) which are electrically insulated from one another.

B. Common Soft Magnetic Materials

Table I compares the properties of five common soft magnetic materials used in electric machines.

TABLE I. Properties of Common Soft Magnetic Materials

Material	Saturation Flux Density @10kA/m	Max. Relative Permeability [1]	Iron Loss @ 1T W/kg	
			50Hz	400Hz
Silicon-iron - non-oriented - grain-oriented	 1.7 to 1.9T 1.8 to 2.0T	 >5,000 >10,000	 1 to 3 0.3 to 1	 20
Mild steel	1.4 to 1.9T	>1,000	-	-
Cobalt-iron	2.2 to 2.4T	>2,000	2	20
SMC [2]	1.2 to 1.6T	200 to 600	6 to 8	50 to 70
AMM [3]	1.6T	>100,000	0.04 to 0.06	1.5

Silicon iron is the most common and lowest cost type of magnetic material used in electric machines. It consists of iron with a small amount (up to 6%) silicon to increase its resistivity and hence reduce its iron losses. It has two main types: non-oriented material which has identical properties in both directions and grain-oriented material which has a preferred magnetic direction. Due to their rotating magnetic fields, electric machines normally use non-oriented materials however transformers can use the more expensive grain-oriented materials to obtain lower magnetic losses.

Mild steel may be used to carry non time-varying magnetic fields in electric machines such as in the rotors of permanent magnet machines.

Cobalt-iron offers significantly higher saturation flux density than silicon-iron but is considerably more expensive. It is mainly used in aerospace and other applications where small size and weight is more important than cost.

Soft magnetic composite (SMC) basically consists of iron powder and an insulating binder, which is compressed into a final core shape. It has lower permeability than other magnetic materials and higher losses than silicon-iron but has a unique capability to create 3D magnetic core shapes. Its low permeability means it is less suited to induction and reluctance machines, but can be used in permanent magnet machines.

Amorphous magnetic material (AMM) is made by rapidly cooling molten silicon-iron to form very thin strip which has very low iron losses. Its main application has been in power system distribution transformers.

III. Soft Magnetic Material Analysis

The analysis of soft magnetic materials will be demonstrated in this section using an example.

Consider the example magnetic core geometry shown in Fig. 2 below. It consists of two iron c-cores, separated by two airgaps of 0.5mm each. The iron cores have a combined mean magnetic flux path length of 0.5m, a uniform cross-sectional area of 5cm by 2cm and a relative permeability μ_R of 5,000. The coil consists of N = 100 turns.

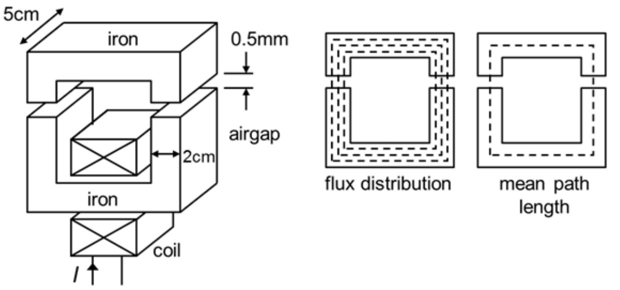

Fig. 2. Example magnetic circuit geometry used for analysis, also showing the idealised magnetic flux distribution and the concept of mean magnetic path length.

A. DC Magnetic Circuit Analysis

Given the core is operating at a flux density B of 1.5T, the required total mmf NI_T can be found as the sum of mmf drop across the core, NI_C and that across the two airgaps, NI_G using (3),

$$\begin{aligned} NI_T &= NI_C + NI_G = \frac{B_C l_C}{\mu_C} + \frac{B_G l_G}{\mu_G} \\ &= \frac{1.5\text{T}\bullet 0.5\text{m}}{5000\bullet 4\pi\times 10^{-7}\,\text{H/m}} + \frac{1.5\text{T}\bullet 2\times 0.0005\text{m}}{4\pi\times 10^{-7}\,\text{H/m}} \\ &= 119.4\text{At} + 1{,}194\text{At} = 1{,}313\text{At} \end{aligned} \tag{4}$$

where the magnetic flux density in the gap B_G is assumed to be the same as that in the iron B_C. It should be noted that the majority of the mmf drop occurs across the airgap as the larger mean magnetic path length of the iron (0.5m) is offset by its higher permeability. The resultant required coil current is 1,313At/100 = 13.13A.

A useful concept is the equivalent air length l_e, defined as the physical length divided by its relative permeability. For the above core, the equivalent air length of the iron is 0.5m/5000 = 0.1mm while that of the air is 1mm. The required mmf is proportional to the equivalent air length as shown in (4).

An alternative means for finding the total mmf is to use the concept of magnetic reluctance $\Re$ which is the magnetic equivalent to electrical resistance. For a magnetic conductor of mean length l and cross-section area A, the reluctance is given by,

$$\Re = \frac{l}{\mu_R \mu_0 A} \quad \text{[At/Wb] or [A/Wb]} \tag{5}$$

The total reluctance of the magnetic circuit, $\Re_T$ is given as the sum of the reluctance of the core $\Re_C$ and the airgap $\Re_G$. This is illustrated in Fig. 3 and can be calculated as,

$$\begin{aligned} \Re_T &= \Re_C + \Re_G = \frac{l_C}{\mu_R \mu_0 A_C} + \frac{l_G}{\mu_0 A_G} \\ &= \frac{0.5\text{m}}{5000\bullet 4\pi\times 10^{-7}\,\text{H/m}\bullet 0.02\text{m}\bullet 0.05\text{m}} \\ &\quad + \frac{2\bullet 0.0005\text{m}}{4\pi\times 10^{-7}\,\text{H/m}\bullet 0.02\text{m}\bullet 0.05\text{m}} \\ &= 79.58\times 10^3 + 795.8\times 10^3\,\text{At/Wb} \\ &= 875.4\times 10^3\,\text{At/Wb} \end{aligned} \tag{6}$$

The airgap has ten times the reluctance compared to the iron core as it has ten times the equivalent air length.

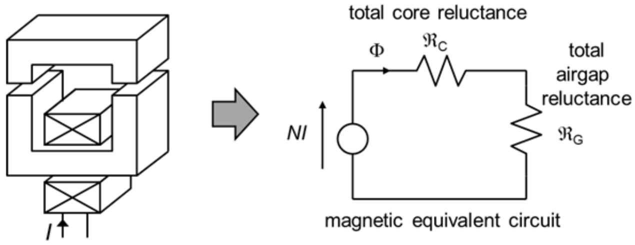

Fig. 3. Concept of magnetic equivalent circuit applied to example geometry.

The mmf required to produce 1.5T in the magnetic circuit can be found from the reluctance using the concept of flux Φ in Webers (Wb) which is the magnetic equivalent of current, where,

$$\Phi = BA = 1.5\text{T}\bullet(0.02\text{m}\times 0.05\text{m}) = 1.5\times 10^{-3}\,\text{Wb} \tag{7}$$

From which the mmf can be found using the magnetic equivalent of Ohm's law,

$$\begin{aligned} NI_T &= \Phi\Re_T = 1.5\times 10^{-3}\,\text{Wb}\bullet 875.4\times 10^3\,\text{At/Wb} \\ &= 1{,}313\text{At} \end{aligned} \tag{8}$$

which is the same as what was found using (4).

B. Flux-Linkage, Inductance, Energy and Force

The flux-linkage λ of a winding in volt-seconds (Vs) can be defined in terms of the flux Φ through the winding and the number of turns in the winding N,

$$\lambda \equiv N\Phi = 100\bullet 1.5\times 10^{-3}\,\text{Wb} = 0.15\text{Vs} \tag{9}$$

The inductance of a winding in Henrys (H) is related to the winding flux linkage λ and winding current I by,

$$\lambda = LI \quad \rightarrow \quad L = \frac{\lambda}{I} = \frac{0.15\text{Vs}}{13.13\text{A}} = 11.42\text{mH} \tag{10}$$

The inductance can also be found using the total magnetic reluctance of the circuit $\Re_T$ and the number of turns N by combining equations (8)-(10),

$$L = \frac{N^2}{\Re_T} = \frac{100^2}{875.4\times 10^3\,\text{At/Wb}} = 11.42\text{mH} \tag{11}$$

which is the same result as in (10).

The stored energy E in Joules (J) of a uniform magnetic field B in a material with a corresponding magnetic field strength H , relative permeability μ_R and volume V is given by,

$$E = \frac{1}{2} BH \times V = \frac{1}{2} \frac{B^2}{\mu_R \mu_0} \times V \quad \text{[J]} \tag{12}$$

From (12), the total stored energy E_T in the example geometry considered is the sum of the stored energy in the core E_C and the airgap E_G and is given by,

$$\begin{aligned} E_T &= E_C + E_G = \frac{1}{2} \frac{B_C^2}{\mu_R \mu_0} V_C + \frac{1}{2} \frac{B_G^2}{\mu_0} V_G \\ &= \frac{1}{2} \frac{(1.5\text{T})^2}{5000 \bullet 4\pi \times 10^{-7}\,\text{H/m}} 0.02\text{m} \bullet 0.05\text{m} \bullet 0.5\text{m} \\ &\quad + \frac{1}{2} \frac{(1.5\text{T})^2}{4\pi \times 10^{-7}\,\text{H/m}} 0.02\text{m} \bullet 0.05\text{m} \bullet 2 \bullet 0.0005\text{m} \\ &= 0.08953 + 0.8952 = 0.985\text{J} \end{aligned} \tag{13}$$

Note that the stored energy in the core and airgap has the same ratio as their equivalent air lengths.

In most electrical machines, particularly induction and reluctance machines, it is normal practice to attempt to reduce the airgaps in the machine to as small as mechanically possible to reduce the current in the winding to produce a given flux density (the magnetising current). However for some electromagnetic devices such as inductors, airgaps (or lower permeability magnetic materials) are often used deliberately to allow the device to store more energy.

The same value of total stored energy E as in (13) can also be found from the inductance and current using,

$$E = \frac{1}{2} LI^2 = 0.5 \bullet 0.01142\text{H} \bullet (13.13\text{A})^2 = 0.984\text{J} \tag{14}$$

The force of attraction F in Newtons (N), between the two magnetic cores in the example geometry can be found as the rate of change of stored energy in the airgap E_G with airgap length g. This is a function of the total airgap area A of both poles,

$$\begin{aligned} F &= \frac{dE_G}{dg} = \frac{1}{2} \frac{B_G^2}{\mu_0} A \\ &= 0.5 \frac{(1.5\text{T})^2}{4\pi \times 10^{-7}\,\text{H/m}} 2 \bullet 0.02\text{m} \bullet 0.05\text{m} \;=\; 1{,}790\text{N} \end{aligned} \tag{15}$$

The force of attraction is important for axial-flux machines where the axial forces can be large. It can also be important in radial-flux machines where asymmetries in the magnetic field due to the winding configuration or eccentricity can cause a net radial force on the rotor. This is called unbalanced magnetic pull (UMP).

C. Back-EMF Calculation

The above analysis has considered dc currents in the coil and the resultant constant magnetic fields. The analysis also applies to sinusoidally-varying quantities if magnetic saturation is ignored.

Thus for instance in the above magnetic circuit, a sinusoidally-varying current of peak value 13.13A will produce a sinusoidal magnetic flux density in the core of peak value 1.5T and a peak flux of 1.5mWb. At the peak value of current, the peak stored energy is 0.984J and the peak force of attraction between the cores is 1790N.

It is common in electrical machines to consider the peak value of a time-varying magnetic field rather than the rms or average value as magnetic saturation is a peak related effect rather than an average effect.

From Faraday's law, the induced voltage e in a coil is given by the rate of change of flux-linkage λ with time.

$$e(t) = \frac{d\lambda(t)}{dt} = N \frac{d\phi(t)}{dt} \tag{16}$$

From this it can be shown that the rms value of the induced voltage, E, for a sinusoidally-varying flux waveform of peak value Φ, and frequency f in a core with N turns is given by the so-called "back-EMF" equation,

$$\begin{aligned} E &= 4.44 N \Phi f \\ &= 4.44 \bullet 100 \bullet 1.5 \times 10^{-3}\,\text{Wb} \bullet 50\text{Hz} = 33.3\text{V} \end{aligned} \tag{17}$$

By rearranging (17), it can be shown that the peak flux is proportional to the ratio of the induced voltage E divided by the frequency f. As the terminal voltage V of the coil is generally comparable to the induced voltage E, thus by keeping the ratio of V/f constant, the flux in the machine can be kept roughly constant. This is principle is used for variable-speed control of AC machines below rated speed, thus if the operating frequency of the machine is halved, then the supply voltage should also be halved to keep the flux constant and avoid saturation.

The back-EMF equation also shows that electromagnetic devices such as induction machines and transformers operating with a constant frequency have a voltage limit. The flux in the machine is proportional to the applied voltage. Thus if the applied voltage is too large, then the flux in the machine becomes high which causes saturation in the core and can result in large (magnetising) currents in the winding and consequently overheating of the windings.

The back-EMF equation shows that applying a fixed AC voltage and frequency to a magnetic circuit effectively tries to maintain the peak flux in that circuit constant, irrespective of the reluctance of the circuit. On the other hand applying a fixed AC (or DC) current to magnetic circuit maintains a fixed mmf in the circuit, with the flux being determined by the reluctance of the circuit.

D. Non-Ideal Flux Distribution

The prior analysis assumed a simplistic model for the magnetic flux distribution in the circuit as shown in Fig. 4a. In practice, particularly around the airgaps, the flux does not restrict itself to straight lines but tends "bulge" out, this is called fringing flux, see Fig. 4b. Also, some flux from the coil does not link with the second magnetic core, this is called leakage flux, see Fig. 4c. These effects are best modelled using tools such as finite-element software to find the actual flux distribution in the machine.

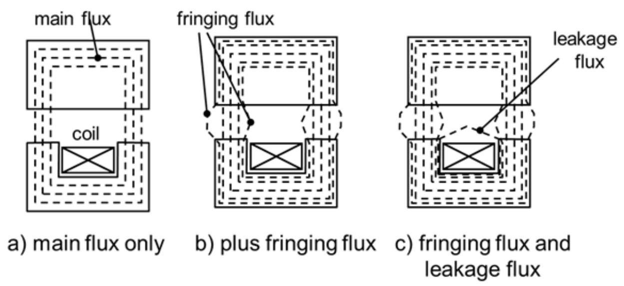

Fig. 4. Definition of fringing and leakage fluxes.

IV. Iron Losses

Iron loss occurs when soft magnetic material materials are exposed to changing magnetic fields. It typically consists of roughly half the total loss in electrical machines when they are operating under their rated condition.

A. General Hysteresis and Eddy-Current Loss Equation

There are two main types of iron losses: hysteresis losses P_h and eddy-current losses P_e, see Fig. 5. Hysteresis losses are associated with the energy lost in changing the magnetisation of the domains in the magnetic material, and are proportional to the frequency. Eddy-current losses are associated with the losses produced by electrical eddy-currents induced by changing magnetic fields in a conductive magnetic material. They are proportional to frequency squared.

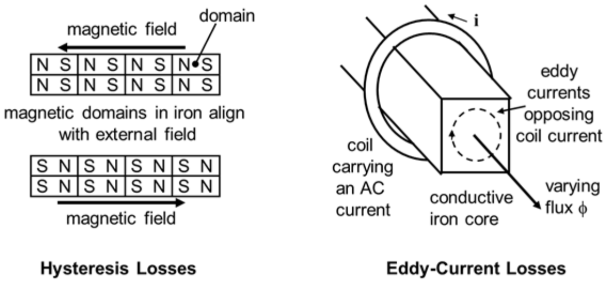

Fig. 5. Principles of hysteresis and eddy-current iron losses.

The iron loss P_{fe} in a magnetic material of volume V carrying a sinusoidally time-varying magnetic flux density of peak magnitude B and frequency f is given by a first approximation by,

$$P_{fe} = P_h + P_e = \left(k_h B^2 f + k_e B^2 f^2\right)V \quad [\mathrm{W}] \qquad (18)$$

where k_h and k_e are constants which depend on the type and thickness of the magnetic material. Note that both hysteresis and eddy-current losses are roughly proportional to the square of the flux density B.

Using the relationships shown in (18) between the loss types and flux density and frequency, knowing the loss at one flux density and frequency, it is possible to estimate the losses for another flux density and frequency. For example, if it is known at 50Hz and 1.5T, that the iron loss in a core is 10W and consists of 6W hysteresis loss and 4W eddy-current loss, then the same core at a frequency of 100Hz and 1T will have a loss of, approximation by,

$$\begin{aligned} P_{fe} &= 6\mathrm{W} \bullet \left(\frac{1\mathrm{T}}{1.5\mathrm{T}}\right)^2 \bullet \left(\frac{100\mathrm{Hz}}{50\mathrm{Hz}}\right) \\ &+ 4\mathrm{W} \bullet \left(\frac{1\mathrm{T}}{1.5\mathrm{T}}\right)^2 \bullet \left(\frac{100\mathrm{Hz}}{50\mathrm{Hz}}\right)^2 \\ &= 5.33\mathrm{W} + 7.11\mathrm{W} = 12.44\mathrm{W} \end{aligned} \qquad (19)$$

This shows that eddy-current loss becomes more important at higher frequencies as it has a frequency squared relationship. An alternative means to obtain the new value of iron loss is to solve for the constants k_h and k_e in (18).

B. Reducing Hysteresis Loss

Hysteresis losses can be reduced by selection of the material type and the thermal processing of the material. Hysteresis losses are significantly affected by mechanical stresses produced by the punching, stacking and winding processes. Thus the hysteresis loss in a finished motor is usually significantly higher than that predicted using the manufacturer's data.

After any mechanical operation, such as rolling of the lamination sheets or punching of the laminations from the sheet, the mechanical residual stresses in the material can be reduced by heating the material to a temperature of several hundred degrees Celsius. This is called annealing.

C. Reducing Eddy-Current Loss

Eddy-current loss can be reduced by dividing solid magnetic materials with thin sheets called laminations which are electrically insulated from one another. This is illustrated in Fig. 6. Note for eddy-current losses to be low in laminations, it is important that the magnetic flux flows in the plane of the laminations and not perpendicular to the lamination plane. Thus laminations are generally only applied to machines which have a 2D flux path. For machines with 3D geometries, materials such as soft magnetic composites are sometimes preferred.

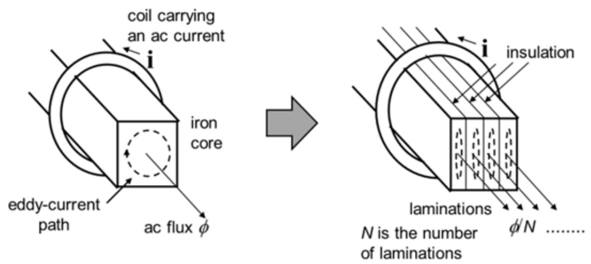

Fig. 6. Eddy-current losses can be reduced by using sheets of magnetic material which are electrically insulated from one another.

Under assumptions of a uniform magnetic field and neglecting skin-effect, the eddy-current loss in Watts/kg of thin laminations of thickness t, resistivity ρ and density d can be shown to be given by,

$$P_e = \frac{\pi^2 t^2 B^2 f^2}{6\rho d} = \frac{\pi^2 t^2}{6\rho d} B^2 f^2 = k_e B^2 f^2 \quad [\mathrm{W/kg}] \quad (20)$$

For example consider 0.47mm thick M-36 steel lamination steel with a resistivity of 0.43μΩm and density of 7.70gm/cm^3, its estimated eddy-current loss at 1.5T and 60Hz is,

$$\begin{aligned} P_e &= \frac{\pi^2 t^2 B^2 f^2}{6\rho d} = \frac{\pi^2 (0.00047\mathrm{m})^2 (1.5\mathrm{T})^2 (60\mathrm{Hz})^2}{6 \bullet 0.43 \times 10^{-6}\,\Omega\mathrm{m} \bullet 7700\mathrm{kg/m}^3} \\ &= 0.889\ \mathrm{W/kg} \end{aligned} \qquad (21)$$

The eddy-current equation is helpful as it shows some of the key relationships for eddy-current loss, that is, it is proportional to the square of the lamination thickness divided by the resistivity of the material. Thus halving the thickness of laminations reduces the eddy-current losses by a factor of four, and doubling the resistivity reduces it by a factor of two.

The two main disadvantages of using thinner laminations are firstly, that it increases the cost of constructing the motor as both the sheet material is more expensive and more laminations need to be punched and stacked for the same stack length. Secondly, because the laminations need to be electrically insulated from one another, there needs to be a thin insulating coating between them. Assuming a fixed thickness for this layer, the thinner the laminations, the less iron there is in the machine to carry magnetic flux. The lamination stacking factor is defined as the fraction of iron per unit axial (stack) length of the machine. This is a function of the lamination thickness.

TABLE II. TYPICAL LAMINATION STACKING FACTOR FOR DIFFERENT LAMINATION THICKNESSES [4]

Lamination Thickness	**Stacking Factor**
0.65mm	0.97
0.5mm	0.96
0.35mm	0.95
0.1mm	0.90
0.025mm	0.74

Lamination materials used for electric machines operating at line frequency (50 or 60Hz) normally use lamination material of 0.50 to 0.65mm thickness. Machines operating at higher frequencies (say, hundreds of Hz) generally use 0.5mm or 0.35mm laminations. Special machines may use even thinner laminations, e.g. a high-speed switched reluctance machine [5] used 0.1mm laminations.

As shown in (20), the eddy-current loss of magnetic materials can also be reduced by increasing their resistivity. Pure iron is a good conductor and to increase its resistivity impurities are added to it. Adding about 3% percent of silicon can increase the resistivity of iron by a factor of about four times, which reduces the eddy-current losses proportionally. Note adding too much silicon affects the physical properties of the steel as well as reduces its saturation flux density as there is basically less iron to carry magnetic flux.

V. HARD MAGNETIC MATERIALS

While soft magnetic materials readily allow the magnetic flux in them to change, hard magnetic materials are selected on their ability to *retain* magnetic fields. They are more commonly known as permanent magnets.

Permanent magnet materials need to be *magnetised* by exposing them to a strong external magnetic field. This magnetic field is usually generated by passing a large current through a coil in a magnetic circuit.

A. BH Loop for Permanent Magnets and Parameters

Fig. 7 illustrates the BH loop for a permanent magnet.

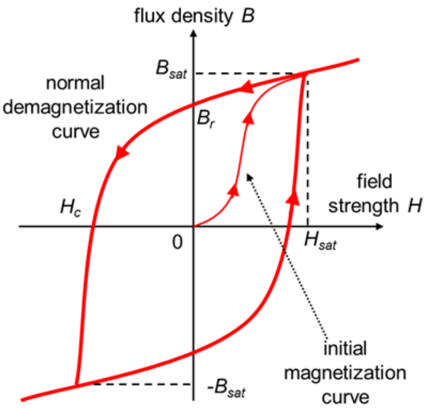

Fig. 7. Example magnetization curve and BH loop for a permanent magnet.

The magnetic field strength or intensity H is applied by, for instance, a current flowing in a magnetic circuit placed around the magnet. Initially the magnet is unmagnetised and has zero flux density B. As the magnetic field intensity H is increased, the flux density in the magnet B increases along what is called the *initial magnetisation curve* until the magnet saturates with a value of flux density B_{sat}. Further increases in H yield only small increases in B.

Decreasing H causes the flux density to fall until it reaches the *remanent flux density* B_r, when H = 0. This is also called the remanence point. Reversing the direction of H and increasing its magnitude produces the *normal demagnetisation curve* where the value of B decreases. The value of H where B = 0 is the coercivity H_c of the material. Further increases in H drives the magnet into saturation in the reverse direction.

The flux density in the magnet, B is the sum of that produced by the external field ($\mu_0 H$) and that produced by the magnet. The latter is called the magnetisation, M. Thus,

$$B = \mu_0 H + M \quad [\text{T}] \tag{22}$$

This is illustrated in Fig. 8 which shows that the normal curve is the sum of the curves for air and the intrinsic magnetisation curve. The normal and intrinsic curves intersect where H = 0. The intrinsic coercivity H_{ci} is the point where the magnetisation is reduced to zero and is a little larger than the coercivity H_c.

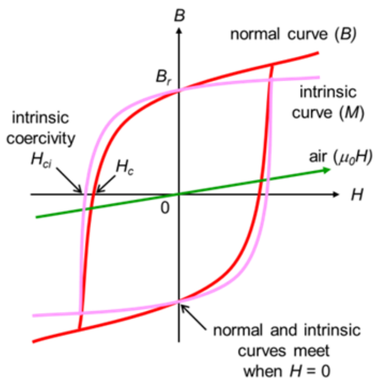

Fig. 8. Normal and intrinsic permanent magnet magnetization curves.

Fig. 9 shows an enlargement of the second-quadrant of the BH curve in Fig. 8. The shape of the curves are more realistic than that illustrated in Fig. 8. It also illustrates the knee point H_k, which is the value of magnetic field strength which the intrinsic (and the normal) BH curve becomes non-linear and represents the onset of *demagnetisation*. This is discussed further in subsection D.

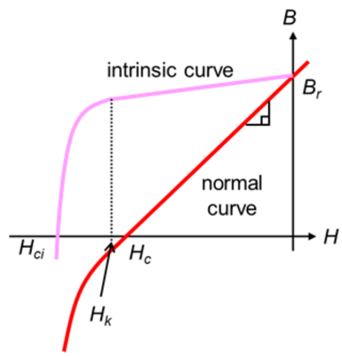

Fig. 9. Definition of key permanent magnet properties.

B. PM Properties and Key Types

There are four key properties of permanent magnet materials: remanent flux density, intrinsic coercivity, maximum energy product and recoil permeability.

Remanent flux density, B_r in Telsa (T), is the "stored" value of maximum flux density (magnetisation) in the permanent magnet in Telsa. It corresponds to the magnetic flux density when the applied magnetic field strength is zero. Typical values are in the range 0.2 to 1.4T.

Intrinsic coercivity, H_{ci} in Amps/m (A/m) corresponds to the value of reverse magnetic field strength which reduces the magnetisation of the magnet to zero. Thus the magnet becomes fully *demagnetised*.

Maximum energy product is the largest value of the product of B and H in the BH curve of the permanent magnet material. A high value minimizes the amount of magnet to produce a given flux density in the airgap

Recoil permeability is the permeability of the magnet near the remanance point (that is, the slope of the BH curve at this point), a value close to unity is preferable for high maximum energy product.

The properties of the main types of permanent magnet materials are summarized in Table III and Fig. 10.

TABLE III. PERMANENT MAGNET MATERIAL PROPERTIES [6]

Property	**Remanence Br [T]**	**Intrinsic Coercivity H_{ci} [kA/m]**	**Max. Energy Product [kJ/m^3]**	**Relative Recoil Perme-ability**
Alnico	0.6 – 1.35	40 – 130	20 – 100	1.9 – 7
Ferrite anisotropic	0.35 – 0.53	180 – 400	24 – 36	1.05 – 1.15
SmCo sintered	0.7 – 1.05	800 – 1500	1600 – 4000	1.02 – 1.07
NdFeB sintered	1.0 – 1.5	800 – 1900	2000 – 3000	1.04 – 1.1

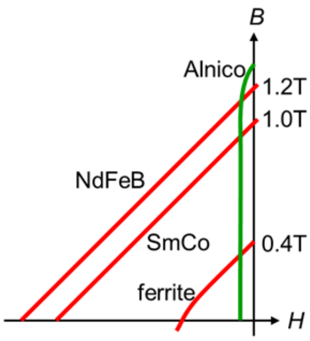

Fig. 10. Typical BH curves for different magnet material types.

Alnico is one of the oldest type of permanent magnets, and has high values of remanent flux density but relatively low intrinsic coercivity and energy product. It is rarely used in electrical machines.

Sintered ferrite magnets have low remanent flux density and moderate coercivity and energy product. Its key advantage is low cost which makes it widely used for high volume, modest performance applications.

The two main types of rare-earth magnets are neodymium iron boron (NdFeB) and samarium cobalt (SmCo) which both have high remanent flux density, intrinsic coercivity and energy product. NdFeB is widely used in high performance PM machines in application such as electric traction and renewable energy generation. SmCo has better high temperature performance than NdFeB but is more expensive and less commonly used.

Magnet materials come in two forms: sintered magnets which are made from pure magnet powder and bonded magnets which have an insulating binder mixed with the magnet powder. Bonded magnets have lower remanent flux density and co-ercivity than the corresponding sintered magnet, typically roughly half. Adding the binder to rare-earth magnets substantially increases its resistivity which reduces its eddy-current losses. The binder also can improve the mechanical properties allowing more complex shapes to be produced.

C. PM Analysis

The airgap flux density produced by a permanent magnet of remanent flux density B_r, length l_M and recoil relative permeability μ_R in a magnetic circuit with an airgap of length l_G is given by,

$$B_g = B_r \frac{l_M / \mu_R}{l_M / \mu_R + l_G} \quad [\mathrm{T}] \qquad (23)$$

This equation assumes that the cross-sectional area of the airgap and magnet in the magnetic circuit are the same.

For example, consider a NdFeB magnet with B_r = 1.2T, μ_R = 1.1T, l_M = 3mm in a magnetic circuit with an airgap l_G = 0.8mm. The airgap flux density B_g is,

$$B_g = B_r \frac{l_M / \mu_R}{l_M / \mu_R + l_G} = 1.2\mathrm{T}\frac{3\mathrm{mm}/1.1}{3\mathrm{mm}/1.1 + 0.8\mathrm{mm}} \qquad (24)$$
$$= 0.928\mathrm{T}$$

The corresponding value of magnetic field strength H_{mg} in the magnet due to the presence of the airgap is given by,

$$H_{mg} = -\frac{B_r - B_g}{\mu_R \mu_0} = -\frac{1.2\mathrm{T} - 0.928\mathrm{T}}{1.1 \bullet 4\pi \times 10^{-7}\ \mathrm{H/m}} = -197\mathrm{kA/m} \qquad (25)$$

The magnet will also see a demagnetising magnetic field strength from the stator which can be described in terms of an equivalent stator turns N_{eq} and the (negative) stator *d*-axis current I_d, as,

$$H_{ma} = \frac{N_{eq} I_d}{l_M / \mu_R + l_G} \qquad (26)$$

The largest negative *d*-axis current may correspond to transient fault conditions, e.g. a 3ph short circuit fault on the machine terminals [7,8].

For instance, consider $N_{eq}I_d$ = -500A, then as,

$$H_{ma} = \frac{N_{eq} I_d}{l_M / \mu_R + l_G}$$
$$= -\frac{500\mathrm{A}}{0.003\mathrm{m}/1.1 + 0.0008\mathrm{m}} = -142\mathrm{kA/m} \qquad (27)$$

The total magnetic field strength in the magnet is given by the sum of that due to the airgap and the stator demagnetising field,

$$H_m = H_{mg} + H_{ma} = -197\mathrm{kA/m} - 142\mathrm{kA/m}$$
$$= -339\mathrm{kA/m} \qquad (28)$$

It is important that this does not exceed the knee point H_k as shown in Fig. 9 to avoid demagnetisation. This is discussed in the next subsection.

D. Demagnetisation and Temperature Performance

If any part of the magnet is exposed to excessive negative values of magnetic field strength this can cause demagnetisation (loss of magnetisation) of those parts. This results in the magnitude of the back-emf reducing and its shape changing.

The demagnetising BH curve of a magnet from the remanence point B_r, starts off with a linear slope of relative permeability μ_R (see Fig. 9). At a particular value of magnetic field strength *H*, the curve becomes non-linear. It is important not to exceed this value of *H* as this will cause demagnetisation.

The value of *H* corresponding to demagnetisation is sensitive to temperature (see Fig. 11). For rare-earth magnets such as NdFeB, it decreases significantly with temperature and so generally worst case operation occurs at the maximum operating temperature. On the other hand, for ferrites, the worst case demagnetisation case generally occurs at the lowest operating temperature.

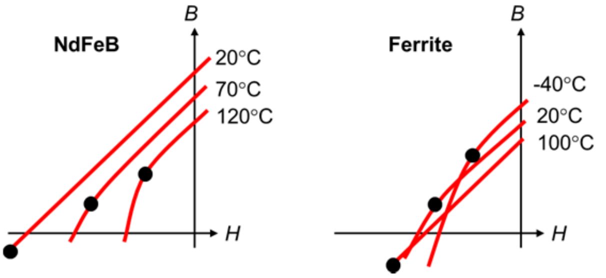

Fig. 11. Effect of temperature on BH curves for NdFeB and ferrite magnets. The dots show the onset of demagnetisation.

The high temperature demagnetisation performance of NdFeB magnets can be improved by adding small quantities of elements such as dysprosium to it. This however adds significantly to the magnet cost. SmCo magnets have better high temperature performance but are significantly more expensive than NdFeB and so are not so widely used.

In summary, demagnetisation analysis consists of two parts. Firstly, the highest negative value of magnetic field strength in the magnet is estimated based on the machine design and the worst-case expectation for the negative I_d current. Secondly, this is compared to the magnet BH curves at the worst case magnet operating temperature. Note that estimating this temperature can be challenging.

VI. Conclusions

Soft and hard magnetic materials are used widely in electrical machines and understanding their properties, performance and analysis is important.

VII. References

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A Word for Today

“Trust in the Lord with all your heart and lean not on your own understanding; in all your ways submit to him, and he will make your paths straight.”

Proverbs 3:5-6 (NIV)