

Motor Control For Designers (Part 10): Motor Magnetics Optimization

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As this series^[1-9] continues to explore the subject of motor design, we put aside the discussion of the torque vs. speed curves in the last part to analyze magnetics design factors. Overall, we follow a similar pattern of past discussions on power supply magnetics—optimize the core, then the winding design. Indeed this part leans heavily into analogies with power supply magnetics, and revisits the previously discussed magnetic analogs to Kirchoff's and Ohm's laws in discussing motor windings and the magnetic operating point.

However, differences emerge when discussing the magnetic characteristics of materials used in motors, which diverge from those of high-frequency transformers. Ultimately, we're working toward an understanding of how to optimally operate a permanent-magnet (hard ferrite), rather the soft ferrite core of a power supply magnet.

This part begins by analyzing power losses in cores and windings and how to balance them for best efficiency. This discussion draws parallels to previous ones on power supply magnetics. The effects of core size, air gap and winding window area as they relate to motor magnetic losses are discussed. The second section revisits the magnetic analog to KCL, and continues the analogies with power supply magnetics to describe the constraints on motor winding design.

Next the choice of magnetic materials, such as silicon steel, is explained in terms of magnetic properties such as permeability and saturation, and their impact on operating frequency and eddy current losses. There's also some musing on the possibility of going to higher operating frequencies in the future through use of ferrites and wire choices (litz) borrowed from the switching converter world.

The last section considers the B-H characteristics of permanent magnets, and manipulates a PM electrical model (again turning to magnetic Kirchoff's and Ohm's laws) to determine where on the B-H curve the magnet should be operated, which in turn suggests stator current and turns ratio.

Magnetic Power Losses

The most basic constraints on motor design are materials limitations, dominated by temperature, and hence design is power-loss limited. Magnetic power loss in the stator is $P_c = p_c(B, f, T_c) \cdot V$ where the power-loss density $p_c(B, f, T_c)$ depends on magnetic field density B , magnetic field frequency f , and the maximum allowable temperature T_c of the magnetic path through stator and rotor.

Maximum power transfer from electrical to mechanical sides of the motor via the magnetic field is achieved when electrical (winding) loss $P_w \approx$ magnetic (core) loss P_c at high efficiency, shown in Fig. 1.

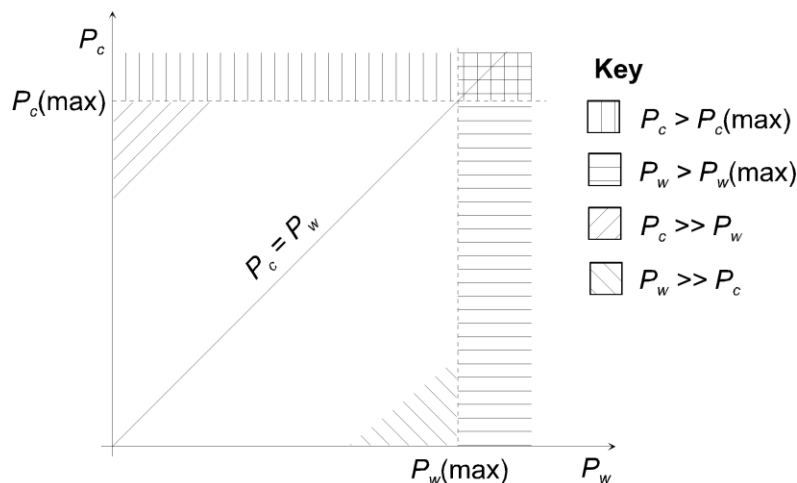


Fig. 1. Optimal loss for maximum power transfer at high efficiency has magnetic and electric losses about equal.

Within the maxima of magnetic “core”—or for motors it’s called the *armature*—and winding losses, $P_c(\text{max})$ and $P_w(\text{max})$ define suboptimal regions of design where P_c and P_w are much different in value.

Motor (magnetics) design, like transformer design, is iterative. (Design knowledge of transformer magnetics applies readily to motors.) Fig. 2 shows tradeoffs. A is stator (core) magnetic-path cross-sectional area and A_c is wire conductive area. The air gap between rotor magnets and stator teeth has length (or width) l_g between them.

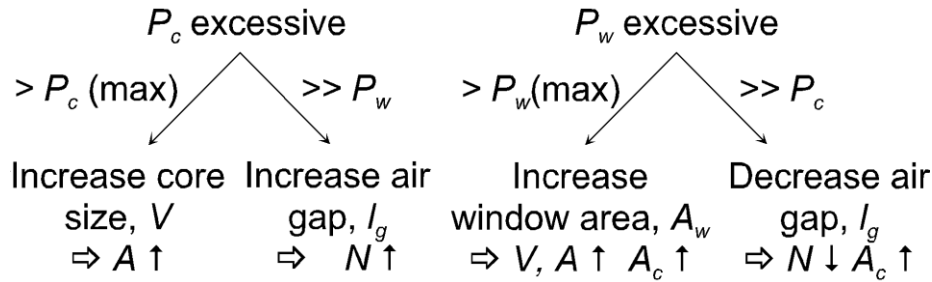


Fig. 2. Winding electrical and stator (core, armature) magnetic design tradeoffs.

Increasing air-gap length l_g decreases permeability μ . This decreases B and hence magnetic power loss P_c but increases H , N -is and electric winding loss P_w . Adjusting the gap length trades off between P_c and P_w . Fig. 3 illustrates this.

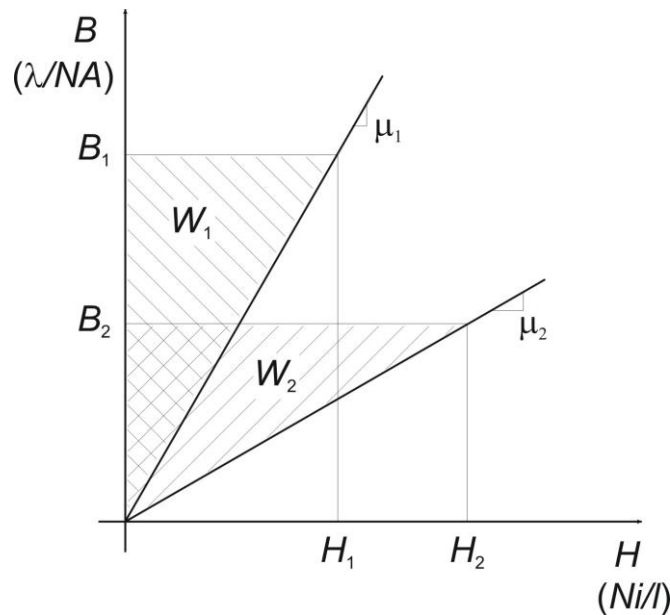


Fig. 3. The effect of air-gap length: $l_{g2} > l_{g1}$; $W_1 = W_2$. A wider gap lowers B but drops more H . P_c varies with B and P_w with H . With increased H , more turns are required at the same current and a longer winding has more loss.

The magnetic energy in the magnetic path is

$$W = \frac{1}{2} \cdot B \cdot H \cdot V$$

$$= \frac{1}{2} \cdot L \cdot i_s^2$$

Most of the magnetic-path energy W is in the gap because H is highest there, with B nearly the same around the closed magnetic path. The flux $\phi = B \cdot A$ is the same everywhere in the magnetic path, but reductions in cross-sectional area A can cause large increases in B and hence "hot spots" in the motor. Like voltage in a current loop, more H is dropped across l_g because air $\mu \ll$ stator μ of electrical steel.

For $W_1 = W_2$, $L_1 = L_2$ and $i_{s1} = i_{s2}$, $A_1 = A_2$, $l_1 = l_2$, then $P_{c1} > P_{c2}$ ($B_1 > B_2$), and $P_{w2} > P_{w1}$ ($N_2 > N_1$);
 A = magnetic path cross-section of B , l = magnetic path length of B loop, and W = magnetic energy stored in the magnetic path of core and gap. Winding resistance is (from part 7)

$$R_w = \frac{p \cdot N \cdot l_c}{\sigma \cdot A_c(f^2)} = \frac{p \cdot N \cdot (2 \cdot l / k_l)}{\sigma \cdot k_p \cdot A_w / (2 \cdot p \cdot N)} = \frac{(2 \cdot p \cdot N)^2}{\sigma \cdot \pi \cdot k_w \cdot k_p \cdot k_l} \cdot \frac{l}{r^2}$$

σ = winding conductivity, p = pole-pairs, l_c = winding length, A_c = conductive cross-sectional area (which varies with f^2), N = winding turns, and k_p is the wire packing factor, the ratio of conductor area to total winding area. From this, winding loss is $P_w = R_w \cdot i_s^2$; where i_s = rms i_s . $A_c(f^2)$ frequency dependencies are from skin and proximity effects, where the resulting A_c is the effective cross-sectional area of winding wire.

Motor Magnetics

Analogous to current being everywhere the same in a circuit loop, magnetic-field flux ϕ is constant around a magnetic loop (MKCL); B varies with magnetic path cross-sectional area A . The relevant magnetics equations are

$$H = \frac{N \cdot i_s}{l}$$

$$B = \mu \cdot H = \left(\frac{\mu}{l} \right) \cdot (N \cdot i_s)$$

The H equation is simplified Ampere's Law and the material relationship $B = \mu \cdot H$ is of the magnetic path. By definition of L , $\lambda = L \cdot i_s$; to the winding circuit

$$\lambda_s = L \cdot i_s = N^2 \cdot \underbrace{\left(\frac{\mu \cdot A}{l} \right)}_{\mathcal{L}} \cdot i_s$$

The circuit flux at the winding terminals is λ_s and is referred through turns N to the field flux ϕ as $\lambda = N \cdot \phi$. From this, the circuit inductance is derived as

$$L = \frac{d\lambda_s}{di_s} = N^2 \cdot \mathcal{L} = N^2 \cdot \left(\frac{\mu \cdot A}{l} \right)$$

These equations are not different from those found in the design of magnetics components for power converters. Where motor design departs from transformer design is the inclusion of the spatial variable θ relating to the motion of the rotor "secondary winding".

For induction and wound-rotor motors, there is a real winding creating the rotor flux, but for PMS and VR motors, a functional equivalent winding is modeled for the field source of the magnets. VR motors are based on

torque production from changing $L(\theta_r)$ that is caused by changes in the magnetic path with rotor rotation. (The resulting ΔL also causes ΔB and torque production.)

Although typical PMS motors are not perfectly *round-rotor*, meaning that their magnetic path is constant with rotation, the VR component of torque production in most PMS motors is small enough to be ignored. This applies especially to step-motors, which are essentially PMS but have typically a few percent of VR. (Motors with teeth, as presented previously in part 1 of the “Motor Theory” series, are called *salient-pole* motors.)

Consequently, the motor *electrical* (winding) *design* panorama can be viewed as a sequence:

$$\text{winding distribution } N(\theta_s) \rightarrow \text{current distribution } N(\theta_s) \cdot i_s \rightarrow B(\theta_s, \theta_r) \rightarrow \phi(\theta_s, \theta_r) \rightarrow \lambda_s \rightarrow L_{sr}$$

where L_{sr} is mutual inductance because it relates to the magnetic field coupling electrical energy to the mechanical side of the motor (as derived in the previous parts on motor theory from either the Lorentz or energy equations). There is also (as in transformers) leakage inductance across winding windows that adds to the total winding inductance. (Leakage inductance is in the series L in the motor model.)

The parameter referring circuit and field quantities is turns N . Stator turns design follows the sequence:

$$N(\theta_s) \rightarrow N(\theta_s) \cdot i_s \rightarrow \frac{\mu_g}{l_g(\theta_r)} \cdot N(\theta_s) \cdot i_s \rightarrow B \cdot A(r) \rightarrow N \cdot \phi \rightarrow N^2(\theta_s) \cdot \frac{\mu_g \cdot A(r)}{l_g(\theta_r)}$$

$l_g(\theta_r)$ is the gap length (or width) between rotor and stator in the magnetic path. Turns distribution around the stator $N(\theta_s)$ combines with the winding current i_s as the field current $N \cdot i_s$. When multiplied by μ/l_g , B results; MQL results in field flux $\phi = B \cdot A$, transformed to circuit flux $\lambda_s = N \cdot \phi$ and to winding inductance $L = N^2 \cdot \mathcal{L}$ where *field inductance* $\mathcal{L} = \mu \cdot A/l$, is the single-turn or per turn-squared inductance.

This sequence follows a “train of thought” intimating a design progression. It falls short of giving yet-to-be-derived design formulas for l_g , N , and $A(r)$, where $A(r)$ is the magnetic path cross-sectional area as a function of motor radius from shaft center, but instead shows how to arrive at them.

In the last expression, μ_g is the gap permeability. Most motors have air as surroundings. Air has a relative permeability of $\mu_r = 1$, relative to matter-free space; that is, $\mu_g \approx \mu_0$. The much higher μ of the materials of stator and rotor around the magnetic path constrain the field to identifiable paths plus some fringing of leakage L . Where the path traverses the rotor-stator gap, the length of the path across the gap is short, preventing the field from fringing much.

Magnetic Materials

With magnetic-path μ as an important factor in the magnetic design of motors, the most common magnetic material (like copper is to windings) is 3% Si (electrical) steel. Its $B(H)$ function is shown simplified in Fig. 4.

The motor stator (the older name is *armature*) is the magnetic *core* which is the stator plus whatever comprises the remainder of the magnetic path—the return steel on the outside of the magnets. In some step-motors, the magnetic path is axial and returns through the shaft and stator along the shaft. (To have bearings conduct magnetic fields or electric current is not generally good design practice.)

Ideally, the stator core has permeability $\mu \rightarrow \infty$ which is most closely approached by ferromagnetic materials (iron). Electrical steel, from Fig. 4, is deeply saturated at 2 T. The typical design limit is 1.6 T at 100 A/m. This saturation value is much higher than that of ferrites which are typically around 0.35 T.

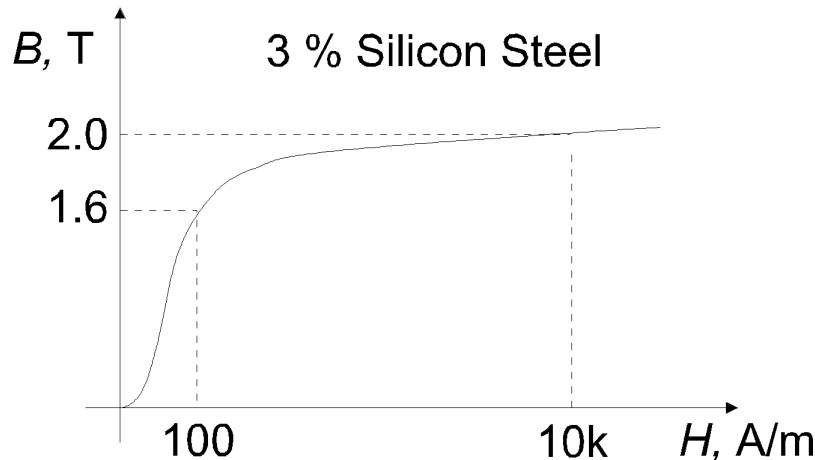


Fig. 4. Electrical (3% Si) steel has high saturation value ($\approx 2T$) and also high hysteresis loss. It begins to saturate significantly at around 1.6 T. (The unit of B , the tesla, $T \equiv V \cdot s/m^2$).

The 3% silicon added to iron reduces magnetic power-loss $P_c(f_{el})$ with frequency f_{el} . Step-motors have many pole-pairs (typically 50) and f_{el} can be high enough to cause eddy-current effects to begin to be significant. A more detailed graph with plots of different magnetic materials for motors is shown in Fig. 5. (This old graph is from GE in Schenectady, NY, the American origin of motor technology.) It shows a wide range of possibilities for $B(H)$.

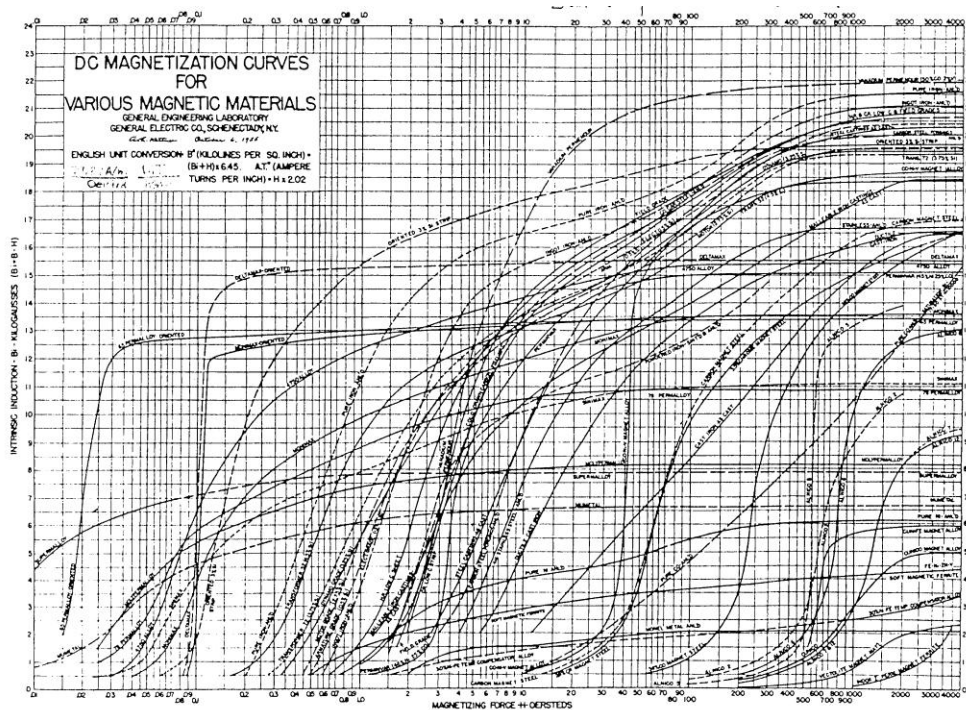


Fig. 5. GE data on magnetic materials for motors. Conversion factors: 79.34 A/m per Oersted; 10 Gauss/mT = 10,000 Gauss/T. "Magnetizing force" is correctly named magnetic field intensity H , expressed nowadays in A/m. (Source: GE)

As motors shrink in size to accommodate robotics and automation goals and to propel EVs, if motor speeds increase, motor design will tend in the direction of power-electronics converter magnetics, with either iron-powder or ferrite cores. There is nothing inherently restrictive of these materials, that they cannot be applied to

motor stators, which are typically made like 50/60-Hz transformer cores of *laminated* Si steel to reduce eddy-currents. They saturate at much less B which reduces transfer-power density, but they also have far less hysteresis loss, allowing them to be operated at much higher frequency.

Steel stators are made of laminations, thin slices of steel insulated from each other and stacked to reduce the conductive lateral (thickness) dimension and hence area within which eddy-currents develop. Laminated silicon steel cores are also in audio power transformers.

Furthermore, it is possible to envision *Litz wire*—bundled strands of twisted wire—as motor windings, to reduce eddy-current effects of the same kinds that appear in converter circuits. These high-frequency considerations will not be of concern here because most actual motors operate at speeds of not over $f_{me} = 20 \text{ krpm} = 333 \text{ Hz}$ me. With two pole-pairs the maximum $f_{el} = 667 \text{ Hz}$ —barely within the range of high-frequency effects.

A step-motor operating at $6 \text{ krpm} = 100 \text{ Hz me} = 5 \text{ kHz el}$ is beginning to enter the realm of high-frequency eddy-current losses, though not far enough for it to be a major design consideration beyond stator lamination.

Permanent Magnets

Magnets have a $B(H)$ relationship, as graphed in Fig. 6. B_r is the residual (*permanent*) magnetic field density with no external H field applied, such as from stator windings. The *coercivity* $H_c = B_r / \mu_m$ is the field intensity required to drive the magnet to zero field density. Additional opposing H ($-H_m$) drives the magnet to have B_m reduced from B_r . (This is usually not a good practice for magnets in that it tends to demagnetize them.) Between the extremes of $(0 \text{ A}, B_r)$ and $(-H_c, 0 \text{ V}\cdot\text{s}/\text{m}^2)$ is the “maximum energy” point for the magnet that results in maximum energy density in the magnetic path, at $(B_m \cdot H_m)_{\max}$. The (H_m, B_m) operating-point (op-pt) of the magnet is at

$$B_m = B_r - \mu_m \cdot H_m = \mu_m (H_c - H_m)$$

$-H_m$ applied to the magnet opposes the magnet field ϕ_m in a demagnetizing (negative) direction or polarity.

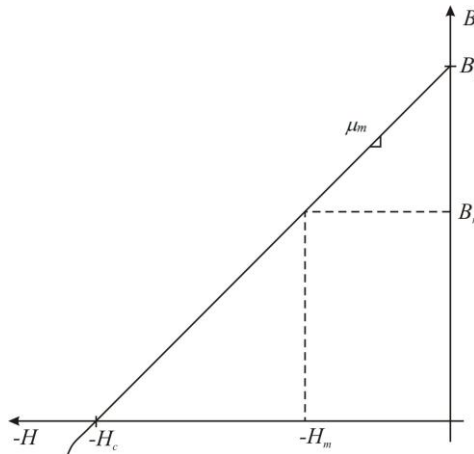


Fig. 6. B - H characteristics of linear PM. B_r is remanence; H_c is coercivity; $(B_m \cdot H_m)_{\max}$ is the energy product.

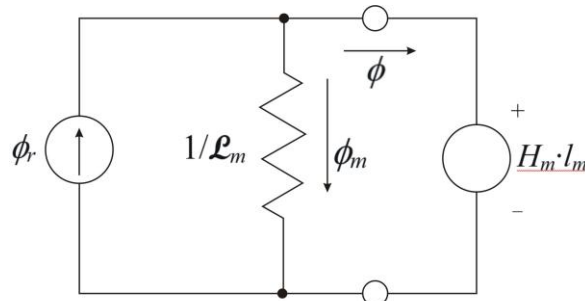
The permanent magnet (PM) plot of $B(H)$ as modeled in Fig. 6 is linear (in the second quadrant) but is also offset from the origin because of the *permanent* magnetic field built into it. The plot has a magnetic circuit PM source field flux $\phi_r = B_r \cdot A_m$ where A_m is the cross-sectional area of the PM magnetic path.

Magnetic circuits are based on the magnetic analogs from part 1, of ΩL , KVL, and KCL as $M\Omega L$, MKVL, and MKCL. What can be confusing at first is that in the magnetic analog of field current Ni is analogous to electric-circuit voltage, as in $M\Omega L$: $\phi = \mathcal{L} \cdot Ni$; field flux ϕ is like current in that it is the same everywhere in the closed loop of the magnetic path, field inductance $\mathcal{L}$ is like conductance, and Ni is like voltage.

A circuit diagram of a magnetic circuit is shown in Fig. 7. It simplifies the magnetic field by confining it to paths, much like electric current is confined to conductors. The PM circuit model is shown to the left of its port terminals.

The linear PM magnetic-circuit model shown in Fig. 7 has as a load for field-current the stator “voltage source” $Ni = H_m \cdot l_m$ constraining the “voltage” across the magnet to be at an op-pt at B_m and $-H_m$. The internal flux of the PM is ϕ_m ; ϕ is flux leaving the magnet with a path completed through $H_m \cdot l_m$ (a stator winding source). By MKCL and the PM $B_m(H_m)$ equation plotted in Fig. 6, output “current” from the PM to the stator is

$$\begin{aligned}\phi &= \phi_r - \phi_m = B_m \cdot A_m = (B_r - \mu_m \cdot H_m) \cdot A_m \\ &= B_r \cdot A_m - \left(\frac{\mu_m \cdot A_m}{l_m} \right) \cdot (H_m \cdot l_m) = \phi_r - \mathcal{L}_m \cdot (H_m \cdot l_m)\end{aligned}$$



Magnet Circuit Model

Fig. 7. Magnetic circuit of a PM loaded by field current $Ni = H_m \cdot l_m$. Field flux ϕ is like circuit current, circuit voltage like field current $H_m \cdot l_m$, and field inductance $\mathcal{L}_m$ like conductance.

The magnetic circuit in Fig. 8 includes the motor air-gap of length l_g with rotor magnet unloaded source flux ϕ_r and winding presenting $N \cdot i_s$ to the PM. The magnetic analogs of Kirchhoff's Laws apply;

$$H_m \cdot l_m = H_g \cdot l_g + N \cdot i_s \quad \text{MKVL}$$

$$\phi_g = B_g \cdot A_g = \phi_r - \phi_m = \phi_r - \mathcal{L}_m \cdot (H_m \cdot l_m) \quad \text{MKCL, M}\Omega$$

The magnetic circuit shown in Fig. 8 is modeled by a Norton equivalent circuit, and by Norton to Thevenin circuit conversion results in the Thevenin equivalent of Fig. 9.

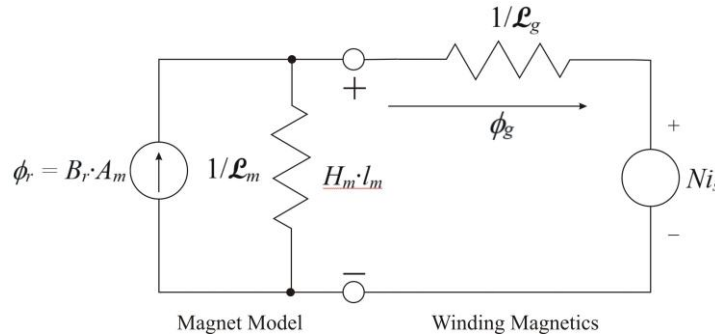


Fig. 8. Norton equivalent model of PM in a magnetic circuit with flux paths internal to the PM as $\phi_m = \mathcal{L}_m \cdot H_m \cdot l_m$ and external through the air gap as ϕ_g and opposed by the stator winding with its own source of Ni_s .

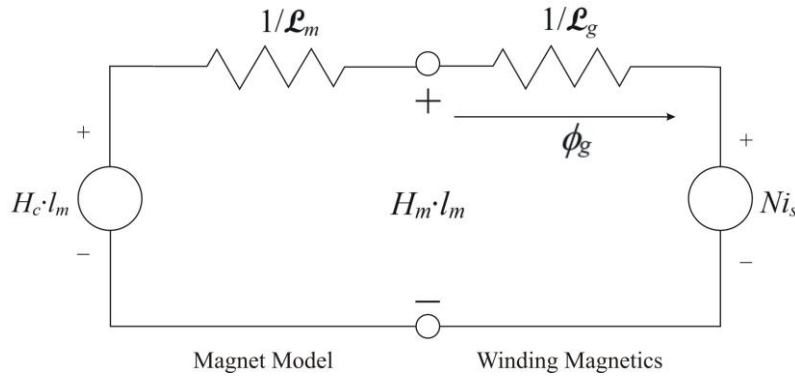


Fig. 9. Thevenin equivalent circuit model of Norton circuit of Fig. 8.

The Norton to Thevenin conversion starts with the magnet PM equation, and solving for B_m ,

$$B_m = \mu_m \cdot (H_c - H_m) \Rightarrow H_c = H_m + B_m / \mu_m$$

Multiply each side by l_m and solve for $H_m \cdot l_m$;

$$H_m \cdot l_m = H_c \cdot l_m - \phi_m / \mathcal{L}_m = H_c \cdot l_m - Ni_m$$

$H_c \cdot l_m$ is the voltage-source equivalent of ϕ_r in series with the drop across $1/\mathcal{L}_m$, resulting in $H_m \cdot l_m$.

Applying MKCL (and $M\Omega L$ to $1/\mathcal{L}_g$) to the Fig. 8 circuit,

$$\phi_r = B_r \cdot A_m = \phi_m + \phi_g = \mathcal{L}_m \cdot H_m \cdot l_m + \mathcal{L}_g \cdot (H_m \cdot l_m - Ni_s) = (\mathcal{L}_m + \mathcal{L}_g) \cdot H_m \cdot l_m - \mathcal{L}_g \cdot Ni_s$$

This can be put in a form interpretable as a superposition equation by solving for the node having field-current

$$H_m \cdot l_m = \frac{B_r \cdot A_m + \mathcal{L}_g \cdot Ni_s}{\mathcal{L}_m + \mathcal{L}_g} = \left(\frac{B_r \cdot A_m}{\mathcal{L}_m + \mathcal{L}_g} \right) + \left(\frac{\mathcal{L}_g}{\mathcal{L}_m + \mathcal{L}_g} \right) \cdot Ni_s$$

From the magnet model equation $B_r = \mu_m \cdot H_c$ and $\mu_m = \mathcal{L}_m \cdot l_m / A_m$. Substituting for B_r and μ_m and reducing,

$$H_m \cdot l_m = \left(\frac{\mathcal{L}_m}{\mathcal{L}_m + \mathcal{L}_g} \right) \cdot (H_c \cdot l_m) + \left(\frac{\mathcal{L}_g}{\mathcal{L}_m + \mathcal{L}_g} \right) \cdot Ni_s$$

The magnetic circuit can also be solved by applying MKVL to the Fig. 9 circuit; the magnet source is the sum of the $Ni = H \cdot l$ drops around the loop;

$$H_c \cdot l_m = \phi_g \cdot \left(\frac{1}{\mathcal{L}_m} + \frac{1}{\mathcal{L}_g} \right) + Ni_s = \frac{\phi_g}{\mathcal{L}_m \parallel \mathcal{L}_g} + Ni_s$$

Solving for $H_m \cdot l_m$ in Fig. 9 by superposition, the same expression results;

$$H_m \cdot l_m = H_c \cdot l_m \cdot \left(\frac{1/\mathcal{L}_g}{1/\mathcal{L}_m + 1/\mathcal{L}_g} \right) + Ni_s \cdot \left(\frac{1/\mathcal{L}_m}{1/\mathcal{L}_m + 1/\mathcal{L}_g} \right) = H_c \cdot l_m \cdot \left(\frac{\mathcal{L}_m}{\mathcal{L}_m + \mathcal{L}_g} \right) + Ni_s \cdot \left(\frac{\mathcal{L}_g}{\mathcal{L}_m + \mathcal{L}_g} \right)$$

The magnetic op-pt varies with stator current. The $N \cdot i_s$ range is designed to be centered at the $(H_m, B_m)_{\max}$ magnet op-pt for maximum B_g . Given the i_s range from motor-drive design, choose the (H_m, B_m) range around $(B_m \cdot H_m)_{\max}$ and calculate N for the design value of $N = N_{is} / i_s$.

In the next part, discussion continues about how large to make the magnets and gap, and a design formula for calculating leakage inductance.

References

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4. "[Motor Control For Designers \(Part 4\): PMS Motor Electrical Model](#)" by Dennis Feucht, Innovatia Laboratories, How2Power Today, November 2025.
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About The Author



Dennis Feucht has been involved in power electronics for 40 years, designing motor-drives and power converters. He has an instrument background from Tektronix, where he designed test and measurement equipment and did research in Tek Labs. He has lately been working on projects in theoretical magnetics and power converter research.

For further reading on motors and motor drives, see "[A Practical Primer On Motor Drives](#)".