

Motor Control For Designers (Part 11): Optimal Gap Sizing And Leakage Inductance In Motor Design

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In the previous part in this series,^[1-9] the discussion on motor design centered around the properties of motor magnetics. It explored how losses in cores and windings influence motor design choices and how magnetic material choices influence losses, power transfer density and operating frequency. A magnetic circuit model for a PM motor was also presented, enabling us to find an optimum operating-point on the B-H curve for the motor magnet, which influences choices of stator current and turns ratio.

Here in part 11, some of the more detailed formulas of motor magnetics design are presented based on the principles set down in the previous parts on motor design.^[9,10] We continue from part 10 and permanent magnets.

Specifically, we'll work out expressions for optimizing magnet volume and gap width. In this discussion, the nonlinear magnetic properties of AlNiCo magnets are introduced to analyze their influence on the magnet operating point and the air gap. However, focusing on the more common magnet types with linear properties, an expression for air gap width is derived with the objective of maximizing air-gap field density and thus maximizing transfer-power density from the electrical to mechanical sides of the motor.

Another consideration is the leakage inductance of the stator winding associated with the magnetic field within the winding windows. We'll derive an expression for this leakage inductance which is a function of the tooth and window dimensions. This inductance is not involved in electrical-mechanical conversion but rather a parasitic inductance that we must consider in the motor model for motor-drive design.

This part concludes with a brief explanation of how distributed phase windings with multiple loops may be used to reduce leakage inductance.

Magnet Size Equation

Some motor magnets are nonlinear and typically have a concave $B(H)$ curve as shown in Fig. 1. The linear magnet model of part 10 is valid only as an incremental model around a magnet $B(H)$ operating-point (op-pt), as shown in Fig. 1 at $(-H_m, B_m)$, and has validity if the stator field does not drive the magnet too far from its op-pt.

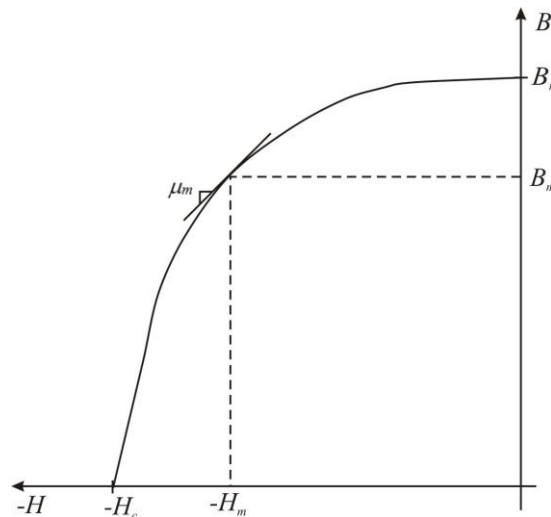


Fig. 1. $B(H)$ plot of nonlinear permanent magnet. The op-pt is chosen to be at or near the maximum-energy $(B_m, H_m)_{\max}$ point of the curve. Stator current i_s causes field current Ni_s to vary magnet operation around this point.

From the magnet circuit model of part 10, copied here as Fig. 2, can be derived a formula for magnet size as volume and with gap field density B_g at stator $Ni_s = 0$ A.

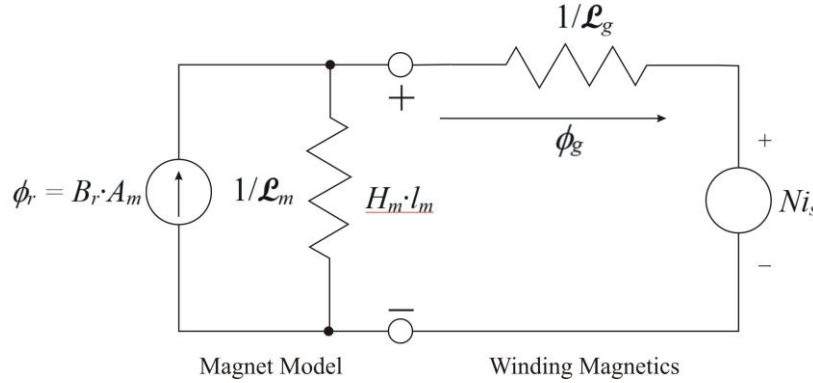


Fig. 2. Norton equivalent model of PM in a magnetic circuit with flux paths internal to the PM as $\phi_m = \mathcal{L}_m \cdot H_m \cdot l_m$ and external through the air gap as ϕ_g and opposed by the stator winding of source Ni_s .

By MKVL, the magnet field current Ni equals that across the air gap; $H_m \cdot l_m = H_g \cdot l_g$. The flux is equal throughout the magnetic loop, and by MKCL, $B_m \cdot A_m = B_g \cdot A_g$. Substituting for $H_g = B_g / \mu_g$,

$$H_m \cdot l_m = \frac{B_g}{\mu_g} \cdot l_g$$

Multiplying left and right sides of the MKVL and MKCL equations by $B_m \cdot A_m$ on the left and $B_g \cdot A_g$ on the right,

$$(B_m \cdot H_m) \cdot (l_m \cdot A_m) = B_m \cdot H_m \cdot V_m = B_g \cdot H_g \cdot (l_g \cdot A_g) = \frac{B_g^2}{\mu_g} \cdot V_g, B_m \cdot H_m \text{ constant}$$

Then solving for magnet volume V_m ,

$$V_m = A_m \cdot l_m = \left(\frac{B_g \cdot A_g}{B_m} \right) \cdot \left(\frac{H_g \cdot l_g}{H_m} \right) = \frac{B_g^2 \cdot V_g}{\mu_g \cdot (B_m \cdot H_m)}, (B_m \cdot H_m) \text{ constant}$$

For a gap of volume V_g , then the value of B_g that the magnet generates varies ($\sim$) with

$$B_g \sim \sqrt{V_m}$$

Whenever $\mu_m \gg \mu_g$, the field is considered non-fringing, $A_m = A_g$, and magnet thickness is

$$l_m = \frac{B_g^2 \cdot l_g}{\mu_g \cdot (B_m \cdot H_m)}$$

For a design goal of minimum V_m and maximum B_g :

$$\begin{aligned} \text{maximize } (B_m \cdot H_m) &\Rightarrow (B_m \cdot H_m) = (B_m \cdot H_m)_{\max} \text{ of magnet} \\ \text{minimize } V_g &\Rightarrow \text{narrow air-gap} \Rightarrow \text{minimum } l_g \Rightarrow \text{high } B_g \end{aligned}$$

By design $B_m \cdot H_m$ is optimized by setting it equal to $(B_m \cdot H_m)_{\max}$. For the same flux ϕ a magnet with greater B_r has less magnet area A_m , and larger H_c reduces magnet thickness l_m .

Nonlinear Magnets

Some common PMS motor magnet materials are listed in the following table.

Table: Magnet initial material characteristics.

Material type	B_r , T	H_c , A/m	$B(H)$
Ceramic	0.35	250	linear
SaCo	0.95	650	linear
NdFeB	1.25	925	linear
AlNiCo 5	1.25	50	nonlinear

A nonlinear magnet material more commonly used in the past is AlNiCo 5. Its $B(H)$ function is depicted in Fig. 3. Initially magnetized it has $B_{r0} = 1.25$ T, but because of hysteresis, it does not return to B_{r0} but to a lesser repeatable value of B_r , the value for an operating motor. AlNiCo has a typical $B_r = 1.125$ T.

An example of magnet length and cross-sectional path area are $l_m = 5$ cm, $A_m = 9$ cm² with $l_g = 1$ mm, $A_g = 8$ cm²; pole-pairs $p = 4$; $N/p = 140$ turns per pole-pair (70 turns/pole), maximum-energy op-pt at $B_m = 0.99$ T at $H_m = -40$ kA/m, and $B_r = 1.125$ T. Then along a curve between 0 A/m and H_m , what is the approximate μ of the magnet near $H = 0$ A/m? The linear approximation of $B(H)$ in Fig. 3 follows the recoil line from op-pt B_m to B_r with the same slope μ_i as the initial $B(H)$ at B_{r0} .

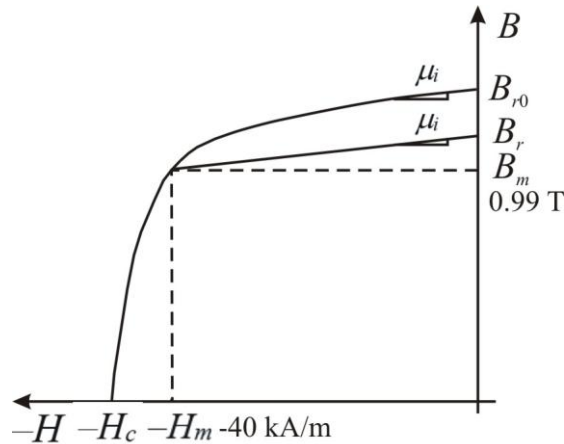


Fig. 3. AlNiCo magnet $B(H)$ graph: $B_r = 1.125$ T, and $(B_m, H_m)_{\max} = (0.99 \text{ T}, -40 \text{ kA/m})$ at maximum magnet energy. B_r is extrapolated to the B -axis at the same slope of μ_i as at B_{r0} .

On the B -axis side of $(B_m, H_m)_{\max}$ the incremental μ is

$$\mu_i \approx \frac{\Delta B}{\Delta H} = \frac{(1.125 \text{ T} - 0.99 \text{ T})}{(40 \text{ kA/m} - 0 \text{ kA/m})} = 3.375 \text{ } \mu\text{H/m}$$

The linearized $B(H)$ equation between B_m and B_r is

$$B_m = 1.125 \text{ T} - (3.375 \text{ } \mu\text{H/m}) \cdot H_m, H_m > 0 \text{ A/m}$$

An approximation of the incremental μ_m at $(B_m, H_m)_{\max}$ is made by finding a linearized μ_c of a line between $(B_m, H_m)_{\max}$ and $-H_c$;

$$\mu_c = \frac{B_m}{H_c - H_m} = \frac{0.99 \text{ T}}{50 \text{ kA/m} - 40 \text{ kA/m}} = \frac{0.99 \text{ T}}{10 \text{ kA/m}} = 99 \mu\text{H/m}$$

The nonlinear $B(H)$ at the op-pt of $(B_m, H_m)_{\max}$ has a piecewise-linear approximation of the an average slope of

$$\mu_m \approx \frac{\mu_i + \mu_c}{2} = 51.2 \mu\text{H/m}$$

What are the field inductances of the magnet and gap?

$$\mathcal{L}_m = \frac{\mu_m \cdot A_m}{l_m} \approx \frac{(51.2 \mu\text{H/m}) \cdot (9 \cdot 10^{-4} \text{ m}^2)}{5 \cdot 10^{-2} \text{ m}} = 0.922 \mu\text{H}$$

$$\mathcal{L}_g = \frac{\mu_g \cdot A_g}{l_g} = \frac{(1.26 \mu\text{H/m}) \cdot (8 \cdot 10^{-4} \text{ m}^2)}{10^{-3} \text{ m}} = 1.01 \mu\text{H}$$

In magnetic circuits, field inductance $\mathcal{L}$ is analogous to electrical conductance in electrical circuits. The magnet $\mu_m \gg \mu_g$, yet the air gap is a somewhat better magnetic flux conductor than the magnet itself because of its short length. The gap is linear but the magnet is not. The μ_m value approximated at the op-pt at B_m is dominated by the steeper part of the $B(H)$ curve.

The field inductances, as “magnetic conductances,” form a flux (“current”) divider in Fig. 2 (repeated below) with no winding current and $Ni_s = 0$ A. Then the Fig. 2 magnetic circuit motor (gap) flux is

$$\phi_g = \left(\frac{\mathcal{L}_g}{\mathcal{L}_m + \mathcal{L}_g} \right) \cdot \phi_m = [\mathcal{L}_g \parallel \mathcal{L}_m] \cdot (H_m \cdot l_m) = (482 \text{ nH}) \cdot [(40 \text{ kA/m}) \cdot (50 \text{ mm})] = (482 \text{ nH}) \cdot (2 \text{ kA}) = 0.964 \text{ mV} \cdot \text{s}$$

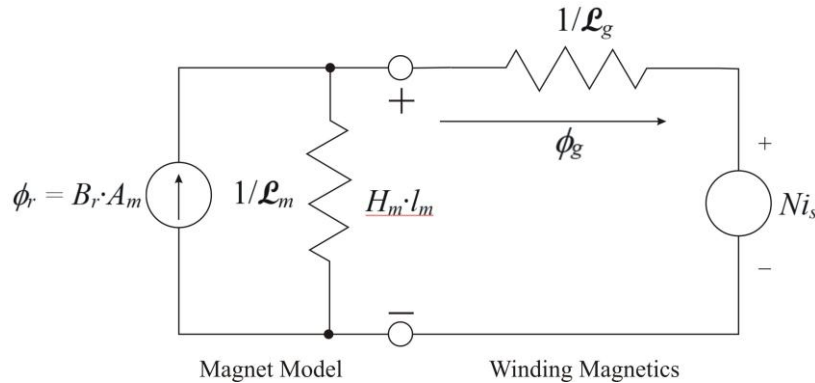


Fig. 2 again. Norton equivalent model of PM in a magnetic circuit.

The linearized magnetic circuit diagram is shown in Fig. 4 with the magnet model of Fig. 2 converted from a Norton to a Thevenin equivalent circuit. H_0 is the open-circuit magnet H_c . The model is accurate only at a given op-pt. Away from the op-pt, only the total-variable nonlinear model is valid.

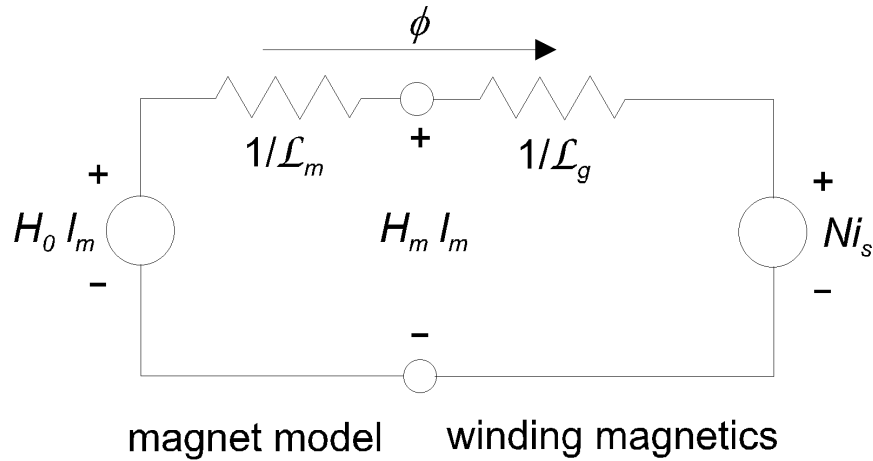


Fig. 4. Magnetic circuit with magnets and rotor-stator air gap between them.

What are the B values produced by the magnet? At the op-pt, $\phi = B_m \cdot A_m = 0.964 \text{ mV}\cdot\text{s}$; then

$$B_g = \frac{\phi}{A_g} = \frac{0.964 \text{ mV}\cdot\text{s}}{8 \text{ cm}^2} = 1.21 \text{ T} ; B_m = \frac{\phi}{A_m} = \frac{0.964 \text{ mV}\cdot\text{s}}{9 \text{ cm}^2} = 1.07 \text{ T}$$

From the continuing example of a PMS motor with eight poles (four pole-pairs) and 70 winding turns per pole, λ_{me} is

$$\lambda_{me} = p \cdot (N/p) \cdot \phi = 8 \cdot 70 \cdot (0.964 \text{ mV}\cdot\text{s}) = 0.54 \text{ V}\cdot\text{s} = (0.54) \cdot (2 \cdot \pi) \text{ V/Hz el} = 3.39 \text{ V/Hz el}$$

A mechanical speed in units of krpm is

$$s = 1/s^{-1} = 2 \cdot \pi / \text{Hz el} = 2 \cdot \pi \cdot p / \text{Hz me} = 2 \cdot \pi \cdot p / 60 \text{ rpm me} = 104.7 \cdot p / \text{krpm me}$$

Then $p = 4$ converts λ_{me} as

$$\lambda_{me} = 0.54 \text{ V}\cdot\text{s} = (0.54 \text{ V}) \cdot (419 / \text{krpm}) = 226 \text{ V/krpm}$$

where it is understood that the unit of krpm is always mechanical frequency or speed, in the mechanical reference-frame, and V/krpm refers to flux and voltage on the electrical side of the motor model scaled to mechanical speed—hence the name “electromechanical” flux for λ_{me} .

How much winding current will result in the minimum allowable B_m ? The flux corresponding to minimum B_m is $0.964 \text{ mV}\cdot\text{s}$. Then

$$\begin{aligned} \phi &= \mathcal{L}_g \cdot (H_m \cdot l_m - p \cdot N \cdot i_s) \\ 0.964 \text{ mV}\cdot\text{s} &= (1.01 \mu\text{H}) \cdot (2 \text{ kA} - 560 \cdot i_s) \end{aligned}$$

Solving, $i_s = 1046 \text{ A}/560 = 1.87 \text{ A}$.

Optimum Gap Width For Maximum B_g

By maximizing air-gap field density, transfer-power density from electrical to mechanical sides of the motor is increased. The prevalence of ceramic or NdFeB magnets in PMS motors leads to a focus on linear magnetic design. With linear $B(H)$ magnets as graphed in Fig. 5 the magnet field equations are

$$B = \left(\frac{B_r}{H_c} \right) \cdot H + B_r$$

$$H = H_c \cdot \left(\frac{B}{B_r} - 1 \right)$$

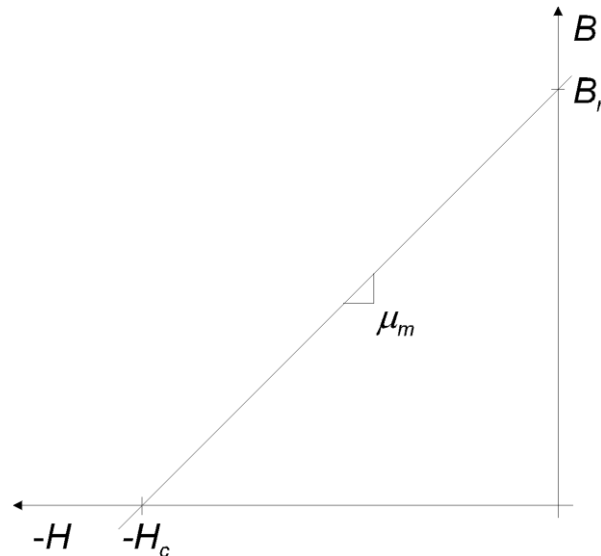


Fig. 5. Graph of linear $B(H)$ magnet. B_r and $-H_c$ are the axis intercepts and permeability $\mu_m = B_r/H_c$ is slope.

The derivation of maximum B_g assumes an inner rotor, outer stator, and the flux of the magnetic path is

$$\phi = B \cdot A(r) = \text{constant}$$

The area of the path varies along the radial (r) dimension of the motor, from shaft center to rotor OD and is a pie-shaped section of

$$\text{Area} = A(r) = (r \cdot \theta) \cdot l$$

where l is the motor dimension into the page. Fig. 6 shows the geometry and dimensions of an inner-rotor motor for the derivation of B_g .

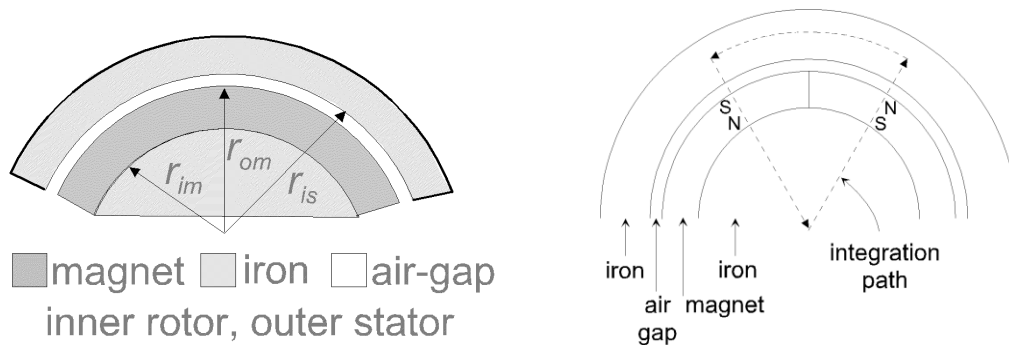


Fig. 6. Geometry of a section of motor from shaft center, through magnet inner (r_{im}) and outer (r_{om}) radii to inner stator (tooth) surface at r_{is} that completes the magnetic path.

The flux is $\phi = B \cdot A(r) = B \cdot l \cdot (\theta \cdot r)$. Then

$$B \cdot r = \frac{\phi}{\theta \cdot l} = \text{constant along radial}$$

The path consists of three materials: iron, magnets, and air. Ideal iron has $H_{Fe} \approx 0 \text{ A/m}$ ($\mu_{Fe} \rightarrow \infty$); ignore the Ni drops (analogous to voltage) along the electrical steel of the path and constrain the field to the non-fringing geometry. Because the motor is round, along the curved axial path through the (ideal) armature iron,

$$\int H_{\theta} \cdot dl \text{ along axial path} = 0$$

Then as previously applied,

$$H_m \cdot l_m = H_g \cdot l_g \text{ MKVL}$$

The symmetric radial paths of the magnetic loop in opposite directions are

$$\int_{r_{im}}^{r_{om}} H_m \cdot dr = \int_{r_{om}}^{r_{is}} H_g \cdot dr = - \int_{r_{is}}^{r_{om}} H_g \cdot dr$$

Substituting the magnet H equation for H_m ,

$$\int_{r_{im}}^{r_{om}} H_c \cdot \left(\frac{(\phi / \theta \cdot l)}{r \cdot B_r} - 1 \right) \cdot dr = - \int_{r_{om}}^{r_{is}} \frac{(\phi / \theta \cdot l)}{\mu_g \cdot r} \cdot dr$$

Perform integrations:

$$\frac{H_c \cdot (\phi / \theta \cdot l)}{B_r} \cdot \ln \left(\frac{r_{om}}{r_{im}} \right) - H_c \cdot (r_{om} - r_{im}) = - \frac{(\phi / \theta \cdot l)}{\mu_g} \cdot \ln \left(\frac{r_{is}}{r_{om}} \right)$$

B_g is maximum at B_{om} , at the face of the magnet. At r_{om} , $(\phi / \theta \cdot l) = B_{om} \cdot r_{om}$ and noting that $H_c/B_r = 1/\mu_m$,

$$B_{om} = \frac{H_c \cdot \left(1 - \frac{r_{im}}{r_{om}} \right)}{- \frac{1}{\mu_m} \cdot \ln \left(\frac{r_{im}}{r_{om}} \right) + \frac{1}{\mu_g} \cdot \ln \left(\frac{r_{is}}{r_{om}} \right)} = \frac{N}{D}$$

To solve for the air-gap width in magnet thickness (r_{om}/r_{im}) that produces the maximum B_{om} , set

$$\frac{dB_{om}}{d\left(\frac{r_{im}}{r_{om}}\right)} = \frac{D \cdot \frac{dN}{d\left(\frac{r_{im}}{r_{om}}\right)} - N \cdot \frac{dD}{d\left(\frac{r_{im}}{r_{om}}\right)}}{D^2} = 0 \Rightarrow D \cdot \frac{dN}{d\left(\frac{r_{im}}{r_{om}}\right)} = N \cdot \frac{dD}{d\left(\frac{r_{im}}{r_{om}}\right)} \Rightarrow$$

$$D \cdot (-H_c) = H_c \cdot \left(1 - \frac{r_{im}}{r_{om}}\right) \cdot \left[-\frac{1}{\mu_m} \cdot \frac{r_{om}}{r_{im}} + 0\right] \Rightarrow D = \left(1 - \frac{r_{im}}{r_{om}}\right) \cdot \frac{1}{\mu_m} \cdot \frac{r_{om}}{r_{im}}$$

Because r_{is} is independent of r_{im} , $d(r_{is}/r_{om}) = 0$ and the air-gap term, is zero. Then simplifying

$$\ln\left(\frac{r_{om}}{r_{im}}\right) + \frac{\mu_m}{\mu_g} \cdot \ln\left(\frac{r_{is}}{r_{om}}\right) = \frac{r_{om}}{r_{im}} - 1$$

Solving for r_{is}/r_{im} ,

$$\ln\left(\frac{r_{is}}{r_{om}}\right) = \left(\frac{\mu_g}{\mu_m}\right) \cdot \left(\frac{r_{om}}{r_{im}} - 1\right) + \ln\left(\frac{r_{om}}{r_{im}}\right)^{\left(-\frac{\mu_g}{\mu_m}\right)}$$

$$\frac{r_{is}}{r_{om}} = \left(\frac{r_{om}}{r_{im}}\right)^{\left(-\frac{\mu_g}{\mu_m}\right)} \cdot e^{\left(\frac{\mu_g}{\mu_m}\right) \cdot \left(\frac{r_{om}}{r_{im}} - 1\right)}$$

Given stator tooth radius r_{is} and r_{im} , the rotor magnet outer radius r_{om} is first determined for maximum B_g . Then magnet thickness = $r_{om} - r_{im}$ and

$$\text{Air-gap width} = l_g = r_{is} - r_{om} = r_{om} \cdot \left(\frac{r_{is}}{r_{om}} - 1\right)$$

Stator-Winding Leakage Inductance

Some of B_g in the stator strays from the high-permeability teeth and spreads out across the winding windows. This field contributes to leakage inductance and does not contribute to electrical-mechanical power conversion. Put another way, the flux from this field does not couple rotor and stator but is solely a part of the stator field. Yet it affects the motor model and hence motor-drive design, so it is included as series inductance in the motor model.

Analysis is simplified to be for one pole-pair—one loop of torque-producing winding. Multiply N by p for p pole-pairs. Then each of the symmetric phase-windings has this leakage inductance.

For this analysis of leakage inductance, the following parameters are defined as in Fig. 7:

$$\begin{aligned} N &= \text{turns in window} \\ r_o &= \text{window outer radius} \\ r_i &= \text{window inner radius} \\ \text{current density} &= J = N \cdot i_s / A(r_o) \end{aligned}$$

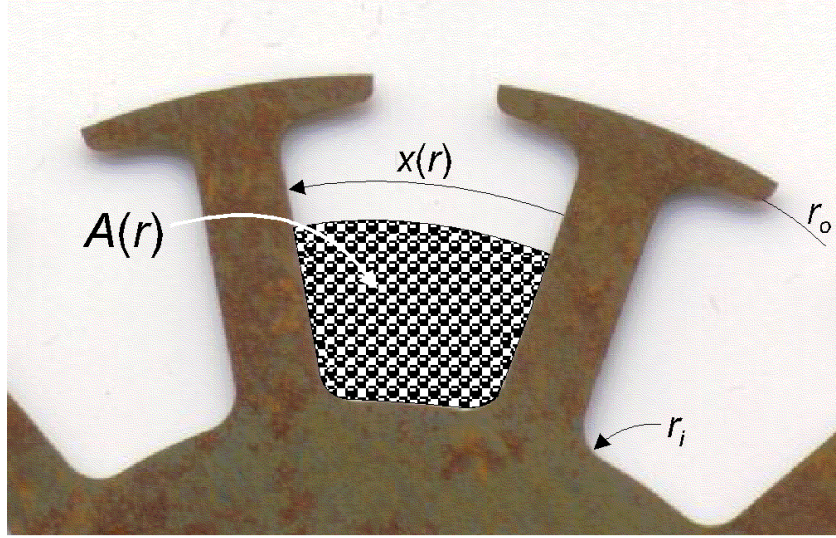


Fig. 7. Photo of an inner-stator lamination with dimensions added for leakage inductance analysis.

Both windows contain the poles of a winding pole-pair that contribute to leakage inductance of a winding loop. The two sides of the loop have a total of $2 \cdot N$ turns going through a total area of $2 \cdot A$ of window, leaving N/A in J . (The end-turns of the loop are disregarded in this analysis.)

Solve for $B(r)$ by integrating along the axial path $x(r)$ across the window in Fig. 7. The rest of the closed path is in steel ($H_{Fe} \cdot l_{Fe} \approx 0$ A). By Ampere's Law,

$$\begin{aligned} \int H \cdot dl &= \frac{B(r)}{\mu_0} \cdot x(r) = J \cdot A(r) = \frac{N \cdot i_s}{A(r_o)} \cdot A(r) \\ B(r) &= \frac{\mu_0 \cdot N \cdot i_s}{A(r_o)} \cdot \frac{A(r)}{x(r)} \end{aligned}$$

The window flux is

$$\phi(r) = l \cdot \int_{r_i}^{r_o} B(r) \cdot dr = \frac{\mu_0 \cdot l \cdot N \cdot i_s}{A(r_o)} \cdot \int_{r_i}^{r_o} \frac{A(r)}{x(r)} \cdot dr$$

In a wedge-shaped window (Fig. 8),

$$x(r) = \theta_w \cdot r$$

$$A(r) = \int_{r_i}^r (\theta_w \cdot r) \cdot (r - r_i) \cdot dr$$

$$= \theta_w \cdot \left(\frac{1}{3} \cdot r^3 - \frac{1}{2} \cdot r_i \cdot r^2 + \frac{1}{6} \cdot r_i^3 \right)$$

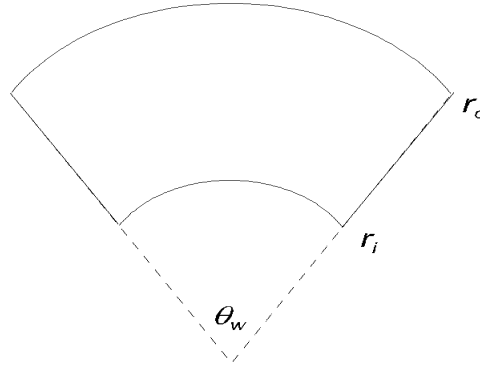


Fig. 8. The pole windings each have dimensions as shown, with window angle θ_w .

$$\int_{r_i}^{r_o} \frac{A(r)}{x(r)} \cdot dr = \int_{r_i}^{r_o} \left(\frac{1}{3} \cdot r^2 - \frac{1}{2} \cdot r_i \cdot r + \frac{1}{6} \cdot r_i^3 \cdot \frac{1}{r} \right) \cdot dr = \frac{1}{9} \cdot r^3 - \frac{1}{4} \cdot r_i \cdot r^2 + \frac{1}{6} \cdot r_i^3 \cdot \ln r \Big|_{r_i}^{r_o}$$

$$= \frac{1}{9} \cdot (r_o^3 - r_i^3) - \frac{1}{4} \cdot (r_i \cdot r_o^2 + r_i^3) + \frac{1}{6} \cdot r_i^3 \cdot \ln(r_o / r_i)$$

Thus, the flux is

$$\phi(r) = l \cdot \int_{r_i}^{r_o} B(r) \cdot dr = \frac{\mu_0 \cdot l \cdot N \cdot i_s}{A(r_o)} \cdot \int_{r_i}^{r_o} \frac{A(r)}{x(r)} \cdot dr$$

$$= \frac{\mu_0 \cdot l \cdot N \cdot i_s}{A(r_o)} \cdot \left\{ \frac{1}{9} \cdot (r_o^3 - r_i^3) - \frac{1}{4} \cdot (r_i \cdot r_o^2 + r_i^3) + \frac{1}{6} \cdot r_i^3 \cdot \ln(r_o / r_i) \right\}$$

Inductance = $L = \lambda / i_s = 2 \cdot N \cdot \phi / i_s$. For conductor window length l (into the page) and a full turn (one loop) wound through two windows (ignoring the end-turns flux), each window is a half-loop having N conductors. The entire loop contributes to the inductance from both windows for a total $\lambda = 2 \cdot \phi$ where ϕ is the window field flux.

By integrating up the winding from r_i to r_o , not only does ϕ vary with r , the turns also vary. The turns of each winding layer vary with r as $N(r) = x(r) \cdot N / A(r_o)$, where N is the total winding turns. Then inductance is found by integrating from r_i to r_o the circuit flux λ , for which both $N(r)$ and $\phi(r)$ vary with r ;

$$L = \frac{\lambda}{i_s} = \frac{2 \cdot \int_{r_i}^{r_o} \left(\frac{N}{A(r_o)} \cdot x(r) \right) \cdot \phi(r) \cdot dr}{i_s} = \frac{2 \cdot N \cdot \theta_w}{A(r_o)} \cdot \frac{\int_{r_i}^{r_o} r \cdot \int_{r_i}^{r_o} \frac{A(r)}{x(r)} \cdot dr \cdot dr}{i_s}$$

$$= \frac{2 \cdot \mu_0 \cdot l \cdot N^2 \cdot \theta_w}{A(r_o)^2} \cdot \left\{ \frac{1}{9} \cdot (r_o^3 - r_i^3) - \frac{1}{4} \cdot (r_i \cdot r_o^2 + r_i^3) + \frac{1}{6} \cdot r_i^3 \cdot \ln(r_o / r_i) \right\} \cdot \int_{r_i}^{r_o} r \cdot dr$$

The expression in brackets is a geometric constant and can be pulled out of the integral. Then leakage inductance of a window-pair or pole-pair or one winding loop is thus

$$L = \frac{\mu_0 \cdot l \cdot N^2 \cdot \theta_w}{A(r_o)^2} \cdot \left\{ \frac{1}{9} \cdot (r_o^3 - r_i^3) - \frac{1}{4} \cdot (r_i \cdot r_o^2 + r_i^3) + \frac{1}{6} \cdot r_i^3 \cdot \ln(r_o / r_i) \right\} \cdot (r_o^2 - r_i^2)$$

Although this equation can be used with a calculator, magnetics design quickly lends itself to computer calculation of design formulas!

Distributed Versus Concentrated Windings

Another design choice for motor windings is whether to construct a phase-winding as a single loop or as multiple, distributed loops around the stator. Concentrated phase-windings are single loop; distributed phase-windings are multiple loops. The inductance of a concentrated winding (one pole-pair) is

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

where $\mathcal{L}$ is the field inductance. By distributing N/p turns of p windings in the same total area A , inductance is

$$L_d = p \cdot \left[\left(\frac{N}{p} \right)^2 \cdot \mathcal{L} \right] = \frac{L_c}{p}$$

Distributed windings of p inductors in series has $1/p^2$ times the inductance of a single (concentrated) winding and is a way of reducing leakage inductance.

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5. "[Motor Control For Designers \(Part 5\): Deriving Force Production From Magnetic Energy](#)" by Dennis Feucht, Innovatia Laboratories, How2Power Today, January 2026.
6. "[Motor Control For Designers \(Part 6\): Three-Phase Motor Waveforms](#)" by Dennis Feucht, Innovatia Laboratories, How2Power Today, April 2026.
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9. "[Motor Control For Designers \(Part 9\): Motor Design Parameters](#)" by Dennis Feucht, Innovatia Laboratories, How2Power Today, July 2026.
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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](#)".