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**ScuDo**

Scuola di Dottorato - Doctoral School  
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Doctoral Dissertation

Doctoral Program in Energy Engineering (38<sup>th</sup> cycle)

# **Design Methods for Axial Flux Machines in E-Aircraft Applications**

By

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\*\*\*\*\*

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## Declaration

I hereby declare that, the contents and organization of this dissertation constitute my own original work and does not compromise in any way the rights of third parties, including those relating to the security of personal data.

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2026

\* This dissertation is presented in partial fulfillment of the requirements for **Ph.D. degree** in the Graduate School of Politecnico di Torino (ScuDo).



*I would like to dedicate this thesis to my loving parents*

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## **Abstract**

The transition toward sustainable transportation systems presents both significant challenges and opportunities, particularly in the integration of electric architectures. International organizations and governments have established ambitious targets for the reduction of greenhouse gas emissions, with the transportation sector accounting for more than 20% of global emissions. Within this sector, aviation plays a critical role, making the electrification of aircraft powertrains and onboard systems a key enabler for achieving these sustainability goals.

In this context, different levels of aircraft electrification have been proposed, with the More Electric Aircraft (MEA) and All Electric Aircraft (AEA) concepts representing the most promising approaches. The MEA paradigm focuses on replacing conventional hydraulic, pneumatic, and mechanical systems with electrically powered alternatives, thereby improving efficiency, reliability, and operational flexibility. This approach has already been adopted in modern large-scale aircraft, contributing to reduced operating costs. In contrast, the AEA concept aims at the complete electrification of the propulsion system. While currently feasible only for small-scale aircraft, its extension to regional and commercial aviation remains limited by the performance of existing technologies.

Achieving full electrification requires substantial advancements across all powertrain subsystems, including energy storage, power electronics, power distribution, and electromechanical energy conversion. In particular, electric machine technology still presents considerable opportunities for improvement in terms of topology, materials, and thermal management.

Among the various machine topologies, axial flux permanent magnet machines (AFMs) appear as strong candidates for aerospace applications due to their high specific torque and compact geometry. These advantages are primarily attributed to the higher number of magnetic poles that can be accommodated for PM configura-

tions, as well as the feasibility of multi-stage and coreless topologies. Furthermore, synchronous AFMs are well suited for concentrated winding arrangements, which offer reduced copper usage, lower Joule losses, and enhanced fault tolerance, features that are particularly attractive for aircraft propulsion systems.

The development of fully electrified aircraft architectures involves long research and development cycles, making the availability of simple and accurate design strategies essential for optimizing electric machines. Within this context, this thesis presents a set of methodologies for the fast sizing and analysis of axial flux machines, aimed at facilitating their integration into design workflows, accelerating prototyping, and reducing time-to-market for electrified propulsion systems.

First, an analytical sizing procedure is developed based on self-consistent electromagnetic formulations to determine the main machine dimensions. The approach also estimates circuit parameters, enabling the calculation of the number of conductors per phase required to meet performance targets. The methodology is validated through finite element simulations and experimental measurements, showing good agreement between predicted and measured results.

Second, an electromagnetic analysis method based on magnetic equivalent circuit (MEC) modeling is introduced. The proposed approach accounts for nonlinear material behavior, rotor motion, leakage fluxes, and multi-stage configurations. Detailed formulations of the reluctance network and the nonlinear solution procedure are provided. The model is validated against finite element simulations and experimental data from axial flux prototypes, demonstrating high accuracy in predicting machine performance.

Finally, a thermal analysis methodology based on a lumped parameter thermal network is presented to estimate temperature distribution within axial flux machines. The model is specifically tailored for early-stage design, requiring limited input data while maintaining sufficient accuracy. Analytical formulations are developed for conduction and convection elements, including correlations adapted to disk-type geometries. Validation against thermal simulations and experimental measurements confirms the capability of the model to predict both steady-state and transient thermal behavior with satisfactory accuracy.

Overall, the proposed methodologies provide a comprehensive and computationally efficient framework for the design and analysis of axial flux machines,

supporting the development of high-performance electric propulsion systems for future sustainable aircraft.

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# Nomenclature

## Acronyms/Abbreviations

1S2R One-Stator Two-Rotor

2S1R Two-Stator One-Rotor

AEA All Electric Aircraft

AF Axial flux

AFM Axial Flux Machine

AFPM Axial Flux Permanent Magnet

AM Additive Manufacturing

CFD Computational Fluid Dynamics

CNT Carbon Nanotube

DL Double Layer

DW Distributed Windings

ESS Energy Storage Systems

FEA Finite Element Analysis

FSCW Fractional Slot Concentrated Windings

GCD Greatest Common Divisor

HTS High Temperature Superconductor



---

|      |                                      |
|------|--------------------------------------|
| IM   | Induction Motor                      |
| IPM  | Interior Permanent Magnet            |
| IRMA | Inner Rotor Modeling Approach        |
| LCM  | Least Common Multiple                |
| LMMA | Linear Motor Modeling Approach       |
| LPTN | Lumped Parameter Thermal Network     |
| MEA  | More Electric Aircraft               |
| MEC  | Magnetic Equivalent Circuit          |
| MMF  | Magnetomotive Force                  |
| ORMA | Outer Rotor Modeling Approach        |
| PM   | Permanent Magnet                     |
| PMSM | Permanent Magnet Synchronous Machine |
| RFPM | Radial Flux Permanent Magnet         |
| SL   | Single Layer                         |
| SMPM | Surface Mounted Permanent Magnets    |
| SR   | Switched Reluctance                  |
| SynR | Synchronous Reluctance               |
| TRL  | Technology Readiness Level           |
| WFSM | Wound Field Synchronous Machine      |

**Greek Symbols**

|               |                                                                   |
|---------------|-------------------------------------------------------------------|
| $\alpha$      | Thermal diffusivity, ( $m^2/s$ )                                  |
| $\alpha_B$    | Temperature coefficient for remanence variation, ( $1/^\circ K$ ) |
| $\alpha_{cu}$ | Temperature coefficient of resistivity, ( $1/^\circ K$ )          |

---

|                                      |                                                             |
|--------------------------------------|-------------------------------------------------------------|
| $\alpha_{PM}$                        | PM to pole pitch ratio, (-)                                 |
| $\beta$                              | Thermal expansion coefficient, (-)                          |
| $\beta_H$                            | Temperature coefficient for coercivity variation, (1/°K)    |
| $\beta_{ax}$                         | Winding geometrical angle, (rad)                            |
| $\delta_{load}$                      | Load angle of the machine, (rad)                            |
| $\delta_{skin}$                      | Skin depth, (m)                                             |
| $\varepsilon$                        | Slot pitching factor of the winding in FSCW, (-)            |
| $\varepsilon_s$                      | Emissivity of the surface, (-)                              |
| $\lambda_d$                          | d-axis flux linkage, (Wb)                                   |
| $\lambda_q$                          | q-axis flux linkage, (Wb)                                   |
| $\lambda_1, \lambda_2, \lambda_{12}$ | Slot coefficient values, (-)                                |
| $\lambda_{abc}$                      | Linked flux in the <i>abc</i> reference frame, (Wb)         |
| $\lambda_{AFM}$                      | Aspect ratio of an axial flux machine, (-)                  |
| $\lambda_{ew}$                       | End-winding leakage permeance factor, (-)                   |
| $\lambda_{slot}$                     | Slot leakage coefficient, (-)                               |
| $\Lambda_{sp}$                       | Flux linked by a single phase, (Wb)                         |
| $\lambda_{tt}$                       | Tooth rip leakage permeance factor, (-)                     |
| $\mu_{air}$                          | Dynamic viscosity of the air, ( $Pa \cdot s$ )              |
| $\mu_{cond}$                         | Conductor material permeability, (H/m)                      |
| $\mu_{fluid}$                        | Dynamic viscosity of the fluid, ( $Pa \cdot s$ )            |
| $\mu_{r,env}$                        | Relative permeability of the surroundings, (-)              |
| $\mu_{r,PM}$                         | Relative permeability of the permanent magnet material, (-) |
| $\mu_{r,st,j}$                       | Relative permeability of the <i>j</i> flux tube, (-)        |

---

|                 |                                                                        |
|-----------------|------------------------------------------------------------------------|
| $\nu$           | Kinematic viscosity, ( $m^2/s$ )                                       |
| $\Omega_{mot}$  | Rotational speed, (m/s)                                                |
| $\Phi_0$        | Initial estimate of the loop flux vector, (Wb)                         |
| $\phi_\sigma$   | Leakage flux, (Wb)                                                     |
| $\phi_m$        | Main magnetizing flux, (Wb)                                            |
| $\phi_{pole1}$  | Amplitude of fundamental component of the magnetic flux per pole, (Wb) |
| $\phi_{pole}$   | Flux per pole, (Wb)                                                    |
| $\Phi_{sy}$     | Stator yoke flux vector, (Wb)                                          |
| $\rho_i$        | Material density of the body in node $i$ , ( $kg/m^3$ )                |
| $\rho_{air}$    | Air density, ( $kg/m^3$ )                                              |
| $\rho_{cond}$   | Conductor electrical resistivity, ( $\Omega m$ )                       |
| $\rho_{cu}$     | Winding electrical resistivity, ( $\Omega m$ )                         |
| $\rho_{fluid}$  | Density of the fluid, ( $kg/m^3$ )                                     |
| $\rho_{load}$   | Load angle of the machine, (rad)                                       |
| $\sigma_\delta$ | Harmonic air gap leakage factor, (-)                                   |
| $\sigma_b$      | Stefan-Boltzmann constant, (-)                                         |
| $\tau_p$        | Pole pitch, (m)                                                        |
| $\tau_s$        | Slot pitch, (m)                                                        |
| $\theta_{cogg}$ | Cogging torque period, (deg)                                           |

### Roman Symbols

|                  |                                       |
|------------------|---------------------------------------|
| $\dot{Q}_{cond}$ | Heat rate of conduction, (W)          |
| $\dot{Q}_{conv}$ | Rate of convective heat transfer, (W) |
| $\dot{Q}_{rad}$  | Rate of radiation heat transfer, (W)  |

---

|                               |                                                         |
|-------------------------------|---------------------------------------------------------|
| $\hat{B}$                     | Peak value of the flux density, (T)                     |
| $\mathbf{A}_r$                | Reluctance matrix, ( $H^{-1}$ )                         |
| $\mathbf{A}_s$                | Stator reluctance sub-matrix, ( $H^{-1}$ )              |
| $\mathbf{A}_{\text{gap,PM}}$  | Air gap - PM reluctance sub-matrix, ( $H^{-1}$ )        |
| $\mathbf{A}_{\text{gap,rot}}$ | Air gap - rotor reluctance sub-matrix, ( $H^{-1}$ )     |
| $\mathbf{A}_{\text{gap}}$     | Air gap reluctance sub-matrix, ( $H^{-1}$ )             |
| $\mathbf{A}_{\text{PM,rot}}$  | PM - rotor reluctance sub-matrix, ( $H^{-1}$ )          |
| $\mathbf{A}_{\text{PM}}$      | PM reluctance sub-matrix, ( $H^{-1}$ )                  |
| $\mathbf{A}_{\text{rot}}$     | Rotor reluctance sub-matrix, ( $H^{-1}$ )               |
| $\mathbf{A}_{s,\text{gap}}$   | Stator - air gap reluctance sub-matrix, ( $H^{-1}$ )    |
| $\mathbf{A}_{s,\text{PM}}$    | Stator - PM reluctance sub-matrix, ( $H^{-1}$ )         |
| $\mathbf{B}_g$                | Vector of magnetic flux density in the air gap, (T)     |
| $\mathbf{C}$                  | Thermal capacitance matrix, (J/K)                       |
| $\mathbf{F}_{\text{wdg}}$     | Vector of MMF produced by the windings, (A)             |
| $\mathbf{G}$                  | Thermal conductance matrix, (W/K)                       |
| $\mathbf{i}_{\text{abc}}$     | Vector of the phase currents, (A)                       |
| $\mathbf{N}_{\text{wdg}}$     | Matrix that describes the winding distributions, (-)    |
| $\mathbf{Q}$                  | Heat source vector, (W)                                 |
| $\mathbf{T}$                  | Node temperature vector, (K)                            |
| $\mathbf{u}$                  | Velocity vector field of the fluid, (m/s)               |
| $A_{\text{conv}}$             | Surface area through which convection occurs, ( $m^2$ ) |
| $A_{\text{cu}}$               | Copper area in the slot, ( $m^2$ )                      |
| $A_c$                         | Cross sectional area of conduction, ( $m^2$ )           |

---

|              |                                                              |
|--------------|--------------------------------------------------------------|
| $A_{ew,top}$ | End-winding upper surface area, ( $m^2$ )                    |
| $A_{ew}$     | End-winding surface area, ( $m^2$ )                          |
| $A_{us}$     | Useful slot area, ( $m^2$ )                                  |
| $B_g$        | Magnitude of air gap flux density, (T)                       |
| $B_r$        | Permanent magnet remanence, (T)                              |
| $B_{g1}$     | Fundamental air gap flux density, (T)                        |
| $B_{r0}$     | PM remanence at a reference temperature, (T)                 |
| $C$          | Empirical coefficient of air velocity, (-)                   |
| $C_i$        | Thermal capacitance of node $i$ , (J/K)                      |
| $c_i$        | Specific heat capacity of the material in node $i$ , (J/kgK) |
| $c_p$        | Thermal specific heat capacity, (J/kgK)                      |
| $c_{air}$    | Specific heat of the air, (J/kgK)                            |
| $C_{ec}$     | Eddy current losses coefficient, (-)                         |
| $C_{exc}$    | Excess losses coefficient, (-)                               |
| $C_{FE}$     | Thermal capacitance of the stator iron lamination, (J/K)     |
| $C_{hsg}$    | Thermal capacitance of the housing, (J/K)                    |
| $C_{hyst}$   | Hysteresis losses coefficient, (-)                           |
| $c_{ly}$     | Number of winding layers in the slot, (-)                    |
| $C_{PM}$     | Thermal capacitance of the PMs, (J/K)                        |
| $C_{ry}$     | Thermal capacitance of the rotor iron lamination, (J/K)      |
| $C_{sh}$     | Thermal capacitance of the shaft, (J/K)                      |
| $C_{th}$     | Thermal capacitance, (J/K)                                   |
| $C_{wdg}$    | Thermal capacitance of the winding, (J/K)                    |

---

|               |                                                            |
|---------------|------------------------------------------------------------|
| $d\mathbf{S}$ | Vector surface element, ( $m^2$ )                          |
| $d\mathbf{F}$ | Tangential force element, (N)                              |
| $d\mathbf{T}$ | Torque vector element, (Nm)                                |
| $D_1$         | Stator lamination inner diameter, (m)                      |
| $D_2$         | Stator lamination outer diameter, (m)                      |
| $D_e$         | Most outer diameter, (m)                                   |
| $D_g$         | Air gap diameter, (m)                                      |
| $D_i$         | Arbitrary diameter for evaluation, (m)                     |
| $D_{im}$      | Most inner diameter, (m)                                   |
| $D_{PM1}$     | PM inner diameter, (m)                                     |
| $D_{PM2}$     | PM outer diameter, (m)                                     |
| $D_{sh}$      | Shaft diameter, (m)                                        |
| $dT/dx$       | Temperature gradient across the medium, (K/s)              |
| $dT_{em}$     | Electromagnetic torque differential element, (Nm)          |
| $E_0$         | No load back EMF, (V)                                      |
| $E_s$         | Stored energy in the slot, (J)                             |
| $f_{op}$      | Operating frequency, (Hz)                                  |
| $g_{acc}$     | Gravitational acceleration, ( $m/s^2$ )                    |
| $Gr$          | Grashof number, (-)                                        |
| $H_c$         | Permanent magnet coercivity, (A/m)                         |
| $h_c$         | General convective heat transfer coefficient, ( $W/m^2K$ ) |
| $H_{c0}$      | PM coercivity at a reference temperature, (A/m)            |
| $h_{conv}$    | Convection heat transfer coefficient, ( $W/m^2K$ )         |

---

|                  |                                                                 |
|------------------|-----------------------------------------------------------------|
| $h_{f,ew}$       | Forced convection coefficient on the end-winding, ( $W/m^2K$ )  |
| $h_{gap}$        | Air gap thickness, (m)                                          |
| $h'_{gap}$       | Air gap equivalent thickness, (m)                               |
| $h_{nat,ew}$     | Natural convection coefficient on the end-winding, ( $W/m^2K$ ) |
| $h_{us}$         | Useful slot height, (m)                                         |
| $h_{yr}$         | Rotor yoke thickness, (m)                                       |
| $h_{ys}$         | Stator yoke thickness, (m)                                      |
| $i_d$            | d-axis current, (A)                                             |
| $i_q$            | q-axis current, (A)                                             |
| $I_s$            | Stator phase current, (A)                                       |
| $I_{DC}$         | DC current for the thermal test, (A)                            |
| $I_{sl}$         | Slot current, (A)                                               |
| $J_{rms}$        | RMS slot current density, ( $A/mm^2$ )                          |
| $k_d$            | Distribution factor for windings, (-)                           |
| $k_p$            | Pitch factor for windings, (-)                                  |
| $K_s$            | Linear current density, (A/m)                                   |
| $k_s$            | Skew factor for windings, (-)                                   |
| $K_s(r)$         | Linear current density at a given radius, (A/m)                 |
| $k_w$            | Winding factor, (-)                                             |
| $k_{1h}, k_{2h}$ | Factor to compute the slot permeance coefficient, (-)           |
| $k_{cu}$         | Slot fill factor, (-)                                           |
| $k_{fluid}$      | Thermal conductivity of the fluid, (W/mK)                       |
| $k_{gap}$        | Portion of air gap section facing a stator tooth, (-)           |

---

|              |                                                           |
|--------------|-----------------------------------------------------------|
| $k_{hold}$   | Rotor holder material thermal conductivity, (W/mK)        |
| $k_{ins}$    | Insulation system equivalent thermal conductivity, (W/mK) |
| $k_{lam}$    | Iron lamination thermal conductivity, (W/mK)              |
| $k_{PM}$     | PM material thermal conductivity, (W/mK)                  |
| $K_{s,max}$  | Maximum linear current density constraint, (A/m)          |
| $k_{sh}$     | Shaft material thermal conductivity, (W/mK)               |
| $k_{skin}$   | Skin effect factor, (-)                                   |
| $k_{th}$     | Thermal conductivity of the material, (W/mK)              |
| $k_{w,h}$    | winding factor of the h-th spatial harmonic, (-)          |
| $k_{w1}$     | Synchronous winding factor, (-)                           |
| $KC$         | Carter Factor, (-)                                        |
| $L_c$        | Carachteristic lenght, (m)                                |
| $L_m$        | Magnetizing inductance, (H)                               |
| $l_{body}$   | Thickness of the body, (m)                                |
| $L_{c,ew}$   | Characteristic length of the end-winding, (m)             |
| $L_{core}$   | Iron stack length, (m)                                    |
| $L_{ew,ave}$ | Average winding turn length, (m)                          |
| $L_{ew}$     | End-winding leakage inductance, (H)                       |
| $L_{gap}$    | Air gap leakage inductance, (H)                           |
| $l_{ins}$    | Insulation system equivalent thickness, (m)               |
| $L_{m,fund}$ | Fundamental magnetizing inductance, (H)                   |
| $L_{PM}$     | Radial length of the PM, (m)                              |
| $L_{slot}$   | Slot leakage inductance, (H)                              |



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|                |                                                            |
|----------------|------------------------------------------------------------|
| $l_{slot}$     | Slot perimeter length (m)                                  |
| $L_{sync}$     | Synchronous inductance, (H)                                |
| $L_{tt}$       | Tooth tip leakage inductance, (H)                          |
| $L_{turn,ave}$ | Average winding turn length, (m)                           |
| $m$            | Number of phases, (-)                                      |
| $m_{cu}$       | Mass of the winding, (kg)                                  |
| $m_{fe}$       | Mass of the iron lamination, (kg)                          |
| $m_{PM}$       | Mass of the PM material, (kg)                              |
| $N_c$          | Number of slots, (-)                                       |
| $n_s$          | Number of stator teeth within one machine periodicity, (-) |
| $N_{cs}$       | Number of conductor in one slot, (-)                       |
| $n_{gap}$      | Number of air gap segments, (-)                            |
| $n_{mot}$      | Motor rotational speed, (rpm)                              |
| $N_{turns}$    | Number of turns in series per phase, (-)                   |
| $Nu$           | Nusselt number, (-)                                        |
| $p$            | Number of poles, (-)                                       |
| $p_0$          | Search direction for the CG solver, (-)                    |
| $P_{cu}$       | Stator copper losses, (W)                                  |
| $P_{Fe,r}$     | Rotor iron losses, (W)                                     |
| $P_{Fe,s}$     | Stator iron losses, (W)                                    |
| $p_{FE}$       | Specific iron losses, (W/kg)                               |
| $p_{fluid}$    | Pressure field, (Pa)                                       |
| $P_{mot}$      | Rated power of the motor, (kW)                             |

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|                |                                                                    |
|----------------|--------------------------------------------------------------------|
| $P_{PM}$       | PM losses, (W)                                                     |
| $pp$           | Pole pairs, (-)                                                    |
| $Pr$           | Prandtl number, (-)                                                |
| $q_s$          | slot per pole per phase, (-)                                       |
| $Q_{th}$       | Volumetric heat source, ( $W/m^3$ )                                |
| $r_0$          | Initial residual, (-)                                              |
| $R_1$          | Stator lamination inner radius, (m)                                |
| $R_2$          | Stator lamination outer radius, (m)                                |
| $r_i$          | Specific radius for computation, (m)                               |
| $R_m$          | Generic magnetic reluctance, ( $H^{-1}$ )                          |
| $R_s$          | Stator phase resistance, ( $\Omega$ )                              |
| $R_{ac}$       | AC electrical resistance of the winding, ( $\Omega$ )              |
| $r_{cond}$     | Conductor radius, (m)                                              |
| $R_{dc}$       | DC electrical resistance of the winding, ( $\Omega$ )              |
| $R_{fringe}$   | Reluctance of the fringe flux paths, ( $H^{-1}$ )                  |
| $R_{gap,min}$  | Minimum air gap reluctance, ( $H^{-1}$ )                           |
| $R_{gap}$      | Magnetic reluctance of the air gap, ( $H^{-1}$ )                   |
| $R_{PM\sigma}$ | PM to PM leakage reluctance, ( $H^{-1}$ )                          |
| $R_{r\sigma}$  | PM to rotor yoke leakage reluctance, ( $H^{-1}$ )                  |
| $R_{s,0}$      | Stator winding resistance at a reference temperature, ( $\Omega$ ) |
| $R_{ss}$       | Slot leakage reluctance, ( $H^{-1}$ )                              |
| $R_{st}$       | Magnetic reluctance of the stator tooth, ( $H^{-1}$ )              |
| $R_{sy}$       | Magnetic reluctance of the stator yoke, ( $H^{-1}$ )               |

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|                 |                                                                 |
|-----------------|-----------------------------------------------------------------|
| $R_{th(i,j)}$   | Thermal resistance between nodes $i$ and $j$ , (K/W)            |
| $R_{th,ew,ia}$  | End-winding - inner air convection thermal resistance, (K/W)    |
| $R_{th,f,ew}$   | End-winding forced convection thermal resistance, (K/W)         |
| $R_{th,gap,ew}$ | End-winding - air gap convection thermal resistance, (K/W)      |
| $R_{th,gap,r}$  | Rotor - air gap convection thermal resistance, (K/W)            |
| $R_{th,gap,s}$  | Stator - air gap convection thermal resistance, (K/W)           |
| $R_{th,hold}$   | Rotor holder radial conduction thermal resistance, (K/W)        |
| $R_{th,hsg}$    | Housing - ambient convection thermal resistance, (K/W)          |
| $R_{th,nat,ew}$ | End-winding natural convection thermal resistance, (K/W)        |
| $R_{th,PM}$     | PM axial conduction thermal resistance, (K/W)                   |
| $R_{th,rot}$    | Rotor holder - end cap air convection thermal resistance, (K/W) |
| $R_{th,ry}$     | Rotor yoke conduction thermal resistance, (K/W)                 |
| $R_{th,sh}$     | Shaft axial conduction thermal resistance, (K/W)                |
| $R_{th,st}$     | Stator tooth conduction thermal resistance, (K/W)               |
| $R_{th,sy}$     | Stator yoke conduction thermal resistance, (K/W)                |
| $R_{th,vent}$   | Ventilation hole convection thermal resistance, (K/W)           |
| $R_{th,wdg}$    | Winding conduction thermal resistance, (K/W)                    |
| $Ra$            | Rayleigh number, (-)                                            |
| $Re$            | Reynolds number, (-)                                            |
| $S$             | Cross-sectional area ( $m^2$ )                                  |
| $T_0$           | Reference temperature, ( $^{\circ}K$ )                          |
| $t_p$           | Periodicity, (-)                                                |
| $T_r$           | Rated torque of the machine, (Nm)                               |

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|             |                                                                        |
|-------------|------------------------------------------------------------------------|
| $T_{abs}$   | Absolute temperature of the surrounded surface, ( $^{\circ}\text{K}$ ) |
| $T_{amb}$   | Ambient temperature, ( $\text{K}$ )                                    |
| $T_{em}$    | Electromagnetic torque, ( $\text{Nm}$ )                                |
| $T_{hsg}$   | Temperature of the housing surface, ( $^{\circ}\text{C}$ )             |
| $T_{op}$    | Temperature of operation, ( $^{\circ}\text{K}$ )                       |
| $T_{surf}$  | Surface temperature, ( $\text{K}$ )                                    |
| $V_i$       | Volume of the body corresponding to node $i$ , ( $m^3$ )               |
| $V_r$       | Rated voltage, ( $\text{V}$ )                                          |
| $V_{air}$   | Air velocity at the end-winding section, ( $\text{m/s}$ )              |
| $V_{cu}$    | Volume of the winding, ( $m^3$ )                                       |
| $V_{DC}$    | DC Voltage for the thermal test, ( $\text{V}$ )                        |
| $V_{PM}$    | Volume of the PM, ( $m^3$ )                                            |
| $V_{slots}$ | Volume of the slots, ( $m^3$ )                                         |
| $V_{tip}$   | Rotor tip speed, ( $\text{m/s}$ )                                      |
| $V_{tot}$   | Total volume of the motor, ( $m^3$ )                                   |
| $V_{ts}$    | Volume of the stator teeth, ( $m^3$ )                                  |
| $V_{yr}$    | Volume of rotor yoke, ( $m^3$ )                                        |
| $V_{ys}$    | Volume of the stator yoke, ( $m^3$ )                                   |
| $w_s$       | Slot width, ( $\text{m}$ )                                             |
| $w_{so}$    | Slot opening, ( $\text{m}$ )                                           |
| $w_{t1}$    | Inner tooth width, ( $\text{m}$ )                                      |
| $w_{t2}$    | Outer tooth width, ( $\text{m}$ )                                      |
| $y$         | Slot per pole, (-)                                                     |

---

|          |                                                      |
|----------|------------------------------------------------------|
| $y_s$    | Coil span, (-)                                       |
| $Z_c$    | Number of turns in series per phase in one slot, (-) |
| $Z_{ph}$ | Conductors in series per phase, (-)                  |
| SP       | Specific power, (kW/kg)                              |

# Chapter 1

## Introduction

### 1.1 Context and motivation

Achieving carbon neutrality requires the electrification of multiple productive sectors to meet the decarbonization targets established by governmental institutions around the world [1]. Among these sectors, transportation electrification represented 21% of global CO<sub>2</sub> emissions in 2024 as shown in fig.1.1, with air transport playing a central role in global emission reduction efforts.

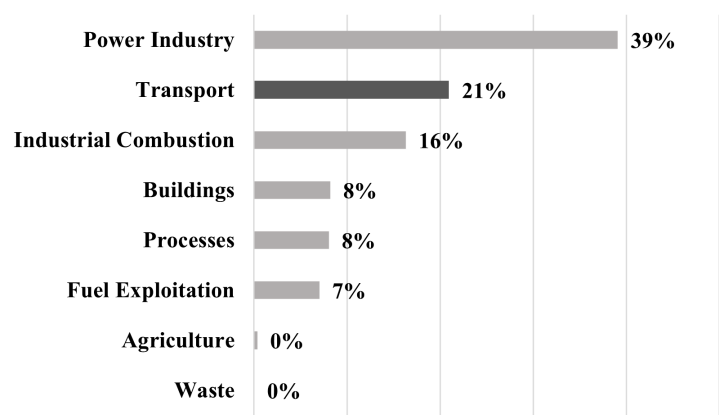


Fig. 1.1 CO<sub>2</sub> emission by sector in 2024 [2].

In this context, the electrification of aviation represents a key strategy for reducing emissions and achieving the sustainability targets established by countries worldwide in the near future. To this end, the concepts of More Electric Aircraft (MEA) and All

Electric Aircraft (AEA) have emerged over recent decades as promising approaches to enhancing sustainability within the air transport sector. This transition is reflected in the increasing electrical power requirements of commercial aircraft over time, as illustrated in Fig. 1.2.

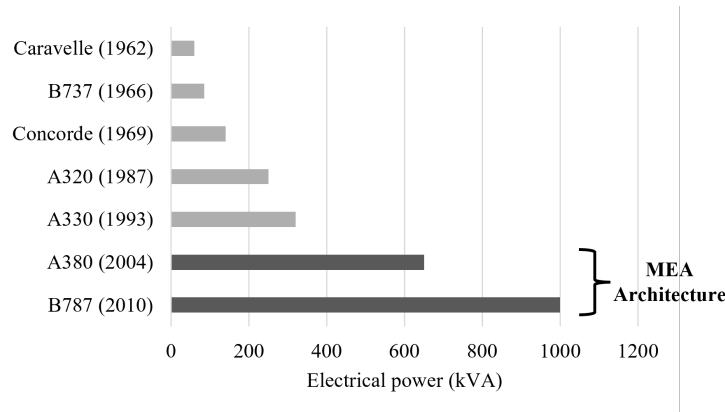


Fig. 1.2 Evolution of electrical power demand by aircraft through the years [2].

Commercial aircraft comprise multiple interconnected systems that are carefully integrated and regulated to ensure optimal performance and safety. Consequently, the introduction of new technologies requires a rigorous systems integration approach to prevent adverse impacts on overall aircraft performance.

The adoption of novel architectures, such as electric propulsion, represents a complex systems engineering challenge. In particular, motors, generators, and associated electrical components generate significant heat that must be effectively dissipated through robust thermal management systems, which can, in turn, influence overall aircraft performance.

More Electric Aircraft architectures have already been implemented in modern platforms such as the Boeing 787 and the Airbus A380, enabling reductions in fuel consumption and operating costs. However, achieving net-zero emissions in the medium term necessitates the electrification of the propulsion powertrain. In this context, energy storage systems (e.g., batteries and fuel cells) and electric machines emerge as the key enabling technologies, offering the greatest potential for further improvement.

With regard to motor technology, increasing specific power (kW/kg) and efficiency is essential to enable long-range commercial electric flight. Among the

various machine topologies, axial flux permanent magnet (AFPM) machines have emerged as one of the most promising solutions for aircraft propulsion systems [3]. Their inherently high specific torque and suitability for low speed operation, enables direct-drive operation without the need for a gearbox, thereby improving overall system efficiency and reliability.

However, achieving the performance targets required for aircraft propulsion electrification demands fast and robust design methodologies for electrical machines. Such approaches are necessary to efficiently explore a wide range of design candidates. A typical electric machine design workflow is illustrated in Fig. 1.3.

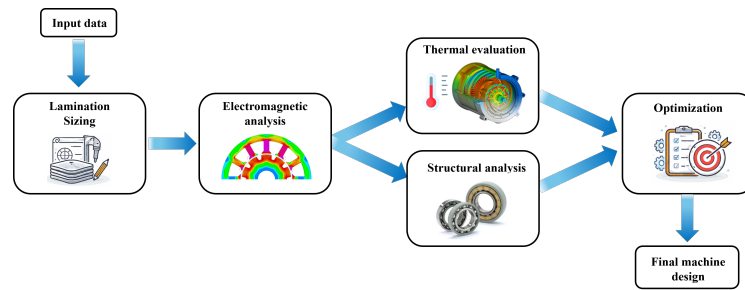


Fig. 1.3 Typical design workflow for an electric machine.

With reference to the design steps illustrated in Fig. 1.3, namely electromagnetic sizing, electromagnetic analysis, and mechanical and thermal evaluation, the design of axial flux machines is particularly challenging due to their inherently three-dimensional electromagnetic behavior and the iterative nature of the design process. Since each design iteration may require the repeated evaluation of multiple physical domains, relying exclusively on computationally intensive numerical methods can result in prohibitively long design cycles. Consequently, simplified and computationally efficient methodologies are essential to enable rapid design iterations, facilitate integration with optimization algorithms, and accelerate the transition from concept to prototype. Within this framework, analytical approaches based on theoretical formulations and empirically derived relationships provide an effective compromise between computational efficiency and predictive accuracy, making them particularly well suited for the early-stage design of axial flux machines.



## 1.2 Research goals

To facilitate the transition from fuel-based aviation to fully electrified propulsion systems, significant improvements in the performance of powertrain subsystems are required to enable zero-emission air transport, such as specific power and efficiency values. Among these subsystems, electric machines play a critical role in determining the feasibility of regional and intercontinental commercial aviation.

However, electric machine design is inherently a multiphysics problem that requires multiple iterations to achieve an optimal, prototype-ready solution. This challenge is particularly pronounced for axial flux (AF) configurations, where the three-dimensional nature of the magnetic flux paths typically necessitates computationally intensive numerical simulations for electromagnetic analysis. As a result, the design process, from initial concept to final implementation, can be time-consuming and resource-intensive.

In this context, the work presented in this thesis focuses on the development of design tools based on analytical and empirical formulations, simplifying the design procedure for less experienced machine designers and significantly reducing the time required to achieve a viable prototype. These methodologies rely on straightforward yet accurate approaches with low computational requirements and are validated through both numerical simulations and experimental measurements.

The contributions of this thesis are presented in the following chapters and summarized below.

- Chapter 1 presented the environmental and sustainability challenges currently faced by the aviation sector. It also contextualized the scope of decarbonization targets established by international and governmental organizations, highlighting the need to advance existing electric machine technologies to enable electric propulsion.
- Chapter 2 reviews aircraft electrification concepts, outlining the main onboard subsystems. It covers the transition from fuel-based systems to More Electric Aircraft architectures and examines the All Electric Aircraft concept, including current technologies and required performance targets. Finally, a survey and data analysis of electric machines for propulsion are presented to highlight historical trends and future performance projections.

- Chapter 3 provides a comprehensive review of the fundamental principles of axial flux permanent magnet machines, which serve as the basis for the subsequent developments in this thesis. In addition, a comparison with radial flux topologies is presented to further justify the relevance of the proposed work. Finally, a state-of-the-art review of modeling techniques and thermal management approaches is included.
- Chapter 4 presents a sizing tool specifically developed for axial flux permanent magnet machines. The procedure is described step by step, relying exclusively on analytical formulations and incorporating sufficient simplifications to make it accessible to designers with limited experience. In addition, the tool is validated to assess the accuracy and applicability of the proposed methodology.
- Chapter 5 presents an electromagnetic analysis tool for axial flux permanent magnet machines based on the magnetic equivalent circuit approach. The network topology is described, along with the advantages and limitations of the proposed method. All relevant formulations for the reluctance elements are provided, together with a discussion of their role within the circuit. Finally, the methodology is validated through comparison with prototype measurements and finite element simulations.
- Chapter 6 presents a thermal analysis approach based on the lumped parameter thermal network methodology. A thermal network tailored for a single-stage axial flux machine is introduced, and each of its elements is described together with the corresponding analytical formulations. In addition, analytical models for the evaluation of convection coefficients are provided, making the approach suitable for early-stage design. The methodology is validated through numerical simulations and experimental temperature measurements on a prototype.
- Chapter 7 presents the main conclusions of the thesis and outlines potential directions for future work.

## Chapter 2

# Aircraft Electrification: Current State

Before the development of heavier-than-air aircraft, flight was achieved using balloons and airships. Early airships were initially powered by steam engines, later replaced by electric motors. The first documented electric flight is attributed to the Tissandier brothers with their La France airship in 1883. The successful flight of the Wright brothers in 1903 marked the beginning of modern aviation and subsequent technological advancements.

An early concept of electrically assisted propulsion was proposed by A.N. Lodygin in 1914, where internal combustion engines drove generators supplying electric motors—an architecture now recognized as hybrid electric aircraft. At that time, onboard electrical systems were limited in power and mainly used for ignition and communication, with energy supplied by batteries and small generators.

As aircraft systems evolved, increasing power demands led to the adoption of engine-driven generators. However, DC-based systems faced limitations in terms of power capability and reliability, particularly due to commutation issues at high altitudes. This prompted the transition toward AC power systems, which offered higher power density, improved efficiency, and reduced weight.

The introduction of semiconductor technology further enabled the development of variable-speed generation and more efficient power conversion. These advancements laid the foundation for modern electrified aircraft architectures and the renewed interest in electric and hybrid-electric propulsion systems. This led to the adoption of new electrified architectures onboard such as the more electric aircraft and all electric aircraft.

More electric aircraft which utilize electrical power to supply non-propulsive on-board systems, represents a paradigm shift from conventional mechanical, hydraulic, and pneumatic architectures toward electrically driven solutions aimed at improving efficiency and reducing operating and maintenance costs.

This transition reached a significant milestone with the introduction of the Airbus A380 (2007) and the Boeing 787 (2011), which featured substantially increased onboard electrical power. These aircraft adopted variable frequency generators directly coupled to the engines, replacing traditional integrated drive generators, and introduced bleed-less architectures that eliminate the use of engine bleed air for secondary systems. These innovations improved propulsion efficiency and overall system performance, contributing to notable reductions in operating costs, such as the approximately 14% decrease achieved by the Boeing 787 compared to the Boeing 767.

Finally, the idea of electric propulsion in aviation appeared as early as the 1970s, when advances in electric motors and power electronics enabled small-scale demonstrations. Key characteristics of early attempts:

- limited by low energy density of batteries;
- restricted to small aircraft or short-duration flights;
- mainly experimental and not commercially viable.

Early developments demonstrated the feasibility of electric propulsion, but also highlighted the critical limitations of energy storage technologies, which remained the primary bottleneck for several decades. More recently, advances in power electronics, materials, and electric machine design have renewed interest in hybrid and fully electric propulsion systems. In this context, high power-density machines, such as axial flux permanent magnet (PM) motors, have emerged as key enabling technologies.

## 2.1 More electric aviation

Aircraft can be classified according to their size, capacity, and intended operational use. General aviation refers to all civil aircraft operations excluding commercial air

transport. Aircraft in this category are typically designed to accommodate a limited number of passengers (generally up to 10) and to operate over relatively short flight ranges (typically up to 1000 km). In contrast, commercial aviation accounts for the majority of passenger and cargo traffic and, consequently, represents the largest share of greenhouse gas emissions within the aviation sector. Within this category, narrow-body aircraft are generally single-aisle configurations with a passenger capacity of up to approximately 300 and are primarily employed on short- to medium-haul routes. Wide-body aircraft, on the other hand, typically feature twin-aisle configurations, accommodate between 200 and 850 passengers, and are predominantly used for long-haul intercontinental flights.

In conventional large commercial aircraft, fuel serves not only as the primary source of propulsion but also as the energy source for secondary onboard systems. As illustrated in Fig. 2.1, the main engine provides power to multiple subsystems, including electrical systems via onboard generators, as well as mechanical, pneumatic, and hydraulic systems required for functions such as fuel and oil pumping and aircraft actuation.

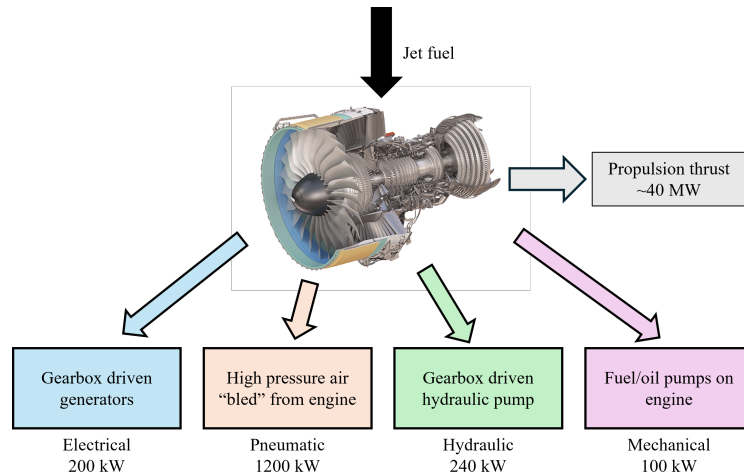


Fig. 2.1 Traditional aircraft power systems [4].

Although this architecture is widely implemented in commercial aviation, it presents notable drawbacks, including limited overall efficiency and increased weight due to the reliance on multiple dedicated power sources for different onboard subsystems [4]. Among these, electrical systems offer distinct advantages in terms of efficiency and operational flexibility across a wide range of applications. Consequently, the More Electric Aircraft concept has emerged as an attractive alternative.

This approach is based on the centralized generation and distribution of electrical power to supply all secondary systems through onboard electrical generators, as illustrated in Fig. 2.2.

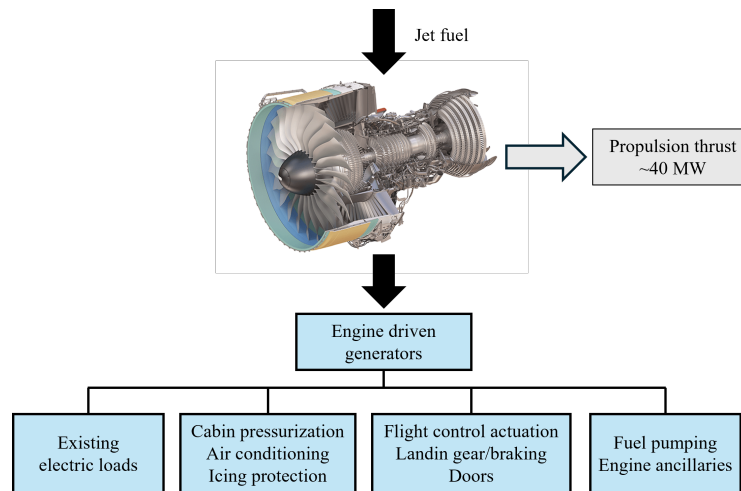


Fig. 2.2 More electric aircraft power systems [4].

Since all secondary systems must be supplied by a single type of energy source, it is necessary to incorporate mechanisms capable of adapting both the supply configuration (AC or DC) and the voltage levels to meet the requirements of the variety of loads present in the aircraft. In this context, the integration of power electronics is essential to ensure the efficient and reliable distribution of electrical power to the wide variety of onboard systems.

In modern large aircraft (e.g. Boeing 787), several voltage and frequency levels can be found (see Fig. 2.3). The common distribution levels are listed below:

- 28V DC - For low power devices (fuel pumps, igniters, etc.) and avionics;
- 270V DC - For secondary systems such as environmental control system fans and compressors, engine start systems, among others;
- 115V AC, 400 Hz - For larger loads such as cooling fans and heating systems;
- Newer models for wide body aviation includes higher voltage levels (540V DC and 230V AC with variable frequency) to reduce cable size, decreasing overall weight.

Table 2.1 reports the different distribution levels and power electronics components present in both conventional aircraft and airplanes models that incorporate the MEA architectures.

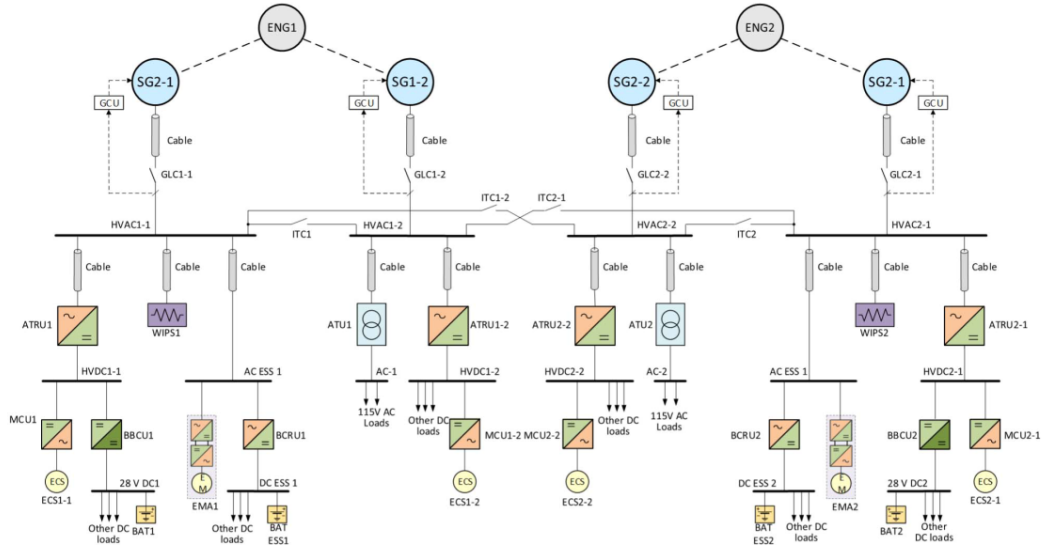


Fig. 2.3 MEA distribution system [5].

Table 2.1 Electrical distribution levels and power electronics in conventional aircraft and MEA [6].

| Aircraft type    | Conventional civil |                    | MEA             |
|------------------|--------------------|--------------------|-----------------|
| Aircraft example | Airbus A320        | Airbus A380        | Boeing 787      |
| Generated power  | 2 x 90 kVA         | 4 x 150 kVA        | 4 x 250 kVA     |
| AC bus           | 115 V / 400 Hz     | 115 V / 400 Hz     | 115 V / 400 Hz  |
|                  |                    | 115 V / 360-800 Hz | 230 V / 360-800 |
| HV DC Bus        | -                  | $\pm 135V$         | $\pm 270V$      |
| LV DC bus        | 28 V               | 28 V               | 28 V            |
| Power converter  | AC / DC            | AC / DC            | AC / DC         |
|                  |                    | DC / AC            | DC / AC         |
|                  |                    | AC / AC            | AC / AC         |

## 2.2 All electric aircraft

In all electric aircraft, all onboard power must be supplied by an energy storage system (ESS), such as a battery pack or a fuel cell. Furthermore, the propulsion thrust required by the aircraft is generated entirely by electric motors. The basic layout of this architecture is illustrated in Fig. 2.4.

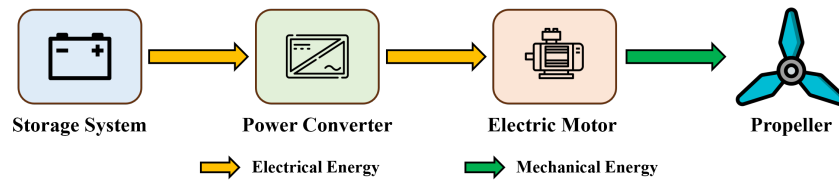


Fig. 2.4 All electric aircraft layout.

Despite the numerous advantages of AEA, such as zero in-flight emissions, reduced noise levels, and potentially lower operating costs, several significant challenges remain to be addressed before this configuration can be considered viable for commercial aviation. At present, limitations in energy density, as well as constraints related to power generation and voltage distribution, restrict the implementation of AEA to small-scale prototypes. Substantial advancements in subsystem technologies are therefore required to enable the full electrification of aircraft operating on regional and intercontinental routes. Table 2.2 summarizes the estimated power and voltage distribution requirements for different classes of electrified aircraft.

Table 2.2 Electrical Power and voltage distribution requirements for AEA [6].

| Aircraft category    | Powertrain power | HV DC bus     |
|----------------------|------------------|---------------|
| General aviation     | < 300 kW         | < 600 V       |
| Narrow body aircraft | 500 kW — 3 MW    | < 800 V       |
| Wide body aircraft   | > 2 MW           | 1000 — 3000 V |

In the following, a brief overview of the current state of the art and the corresponding performance requirements for each subsystem is provided. A more detailed discussion of electric machines is presented in Section 2.3.



### 2.2.1 Energy storage systems

The energy and power densities of an energy storage system are key metrics that determine the range and overall performance of an electrically propelled aircraft. Currently, the most widely adopted technologies for energy storage in electric aircraft propulsion systems are electrochemical batteries and fuel cells (FCs).

Among the available battery technologies, lithium-ion (Li-ion) batteries represent the most practical solution for aircraft electrification at present. They typically exhibit energy density values of approximately 250 Wh/kg at the cell level, which decrease to around 110–160 Wh/kg at the battery pack level (For reference, conventional jet fuel has a gravimetric energy density of approximately 12,000 Wh/kg, nearly two orders of magnitude higher than current battery technologies. However, gas turbine propulsion systems typically operate with overall efficiencies below 30%, whereas electric propulsion systems can achieve efficiencies above 90% [7]). In terms of power density, values can reach up to 1500–2000 W/kg, with energy efficiencies of up to 80% [8], [9]. Although ongoing technological developments are expected to further improve these figures in the near future, substantial advancements are still required, particularly in energy density, to enable the exclusive use of batteries in aircraft with significant flight range [10], [11].

Alternatively, energy storage systems based on low-temperature fuel cells, such as proton exchange membrane (PEM) fuel cells, offer higher energy densities in the range of 600–800 Wh/kg [8], [12]. However, compared to batteries, fuel cell systems generally exhibit lower power densities and efficiencies, typically not exceeding 100 W/kg and 50–60%, respectively [11], [13].

Hybrid energy storage systems are therefore often proposed to leverage the complementary advantages of both technologies. In such configurations, the fuel cell supplies the steady-state power required during cruise conditions, while the battery is used to meet peak power demands, such as those occurring during take-off and climb phases.

### 2.2.2 Power electronics

The variety of aircraft architectures, together with the presence of multiple electrical distribution systems, requires the use of several types of power electronic converters.

For example, AC/DC converters are employed in hybrid and turbo-electric architectures to interface onboard electric generators with the DC distribution bus. AC/AC converters enable the interconnection of systems operating at different voltage and frequency levels while avoiding the need for bulky intermediate DC-link capacitors or inductors. DC/DC converters are widely used across electrified architectures to supply DC loads, interface additional onboard power sources, and adapt voltage levels for low-voltage subsystems [8]. In contrast, DC/AC converters are responsible for supplying AC electric actuators and driving the propulsion motors in electrified aircraft.

Depending on the aircraft category, the power ratings and voltage levels of DC/AC converters for propulsion applications can vary significantly. In general aviation, low-voltage ( $< 600$  V) and low-power ( $< 300$  kW) propulsion systems are typically sufficient to meet thrust requirements. However, for larger and long-range aircraft, these values can increase to 1–3 kV and exceed 2 MW, respectively. In many cases, power converters for propulsion are implemented in parallel configurations to ensure system redundancy and to overcome the power limitations of individual power electronic components [6].

More recently, wide band-gap semiconductor devices, such as silicon carbide (SiC) and gallium nitride (GaN), have emerged as promising alternatives to conventional silicon-based technologies. Their ability to operate at higher switching frequencies enables a reduction in the size and weight of passive components, thereby improving overall system power density [13]. Another key parameter influencing the size of DC/AC converters in propulsion applications is the power factor of the electric machine. For a given delivered power, a lower power factor results in higher reactive power circulation, which in turn increases the required converter rating and overall system size.

### 2.2.3 Electrical distribution systems

The onboard power distribution system of conventional commercial aircraft has traditionally consisted of a low-power 28 V DC bus for avionics and battery-supported services, alongside a high-power 115 Vrms, 400 Hz AC supply for larger loads. However, the continuously increasing electrical power demand of modern more electric aircraft has driven a transition toward alternative onboard distribution architectures,

characterized by higher voltage levels and the introduction of variable-frequency operation [4]. For example, in the Airbus A380, the conventional 28 V DC and 115 V/400 Hz systems are supplemented by a 115 Vrms variable-frequency distribution operating in the range of 360–800 Hz, while the Boeing 787 incorporates 230 Vrms variable-frequency systems and a 540 V DC ( $\pm 270$  V) distribution network [9], [6].

The substantial power demand associated with fully electrified propulsion systems is expected to further drive the evolution of onboard electrical distribution toward medium-voltage DC systems, with voltage levels likely exceeding the kilovolt range [8], [13], [14]. This transition toward DC distribution is primarily motivated by the elimination of skin and proximity effect losses, as well as the reduced corona effects compared to AC systems [8]. Furthermore, increasing the distribution voltage enables a reduction in current for a given power level, thereby decreasing cable size and weight.

Nevertheless, the selection of an optimal voltage level must also account for insulation-related challenges, particularly under high-altitude operating conditions. These include increased susceptibility to partial discharges, insulation aging, and arcing phenomena [14], [15]. This effect is illustrated by Paschen's curve (Fig. 2.5), which describes the variation of breakdown voltage in insulation systems as a function of insulation thickness and operating altitude [16]. Such issues are further exacerbated in electric propulsion systems, where motors are typically supplied by pulse-width modulated drives with high voltage slew rates. Moreover, the chosen distribution voltage level directly impacts the characteristics, dimensions, and weight of the energy storage system, further highlighting the need for a careful and integrated design approach.

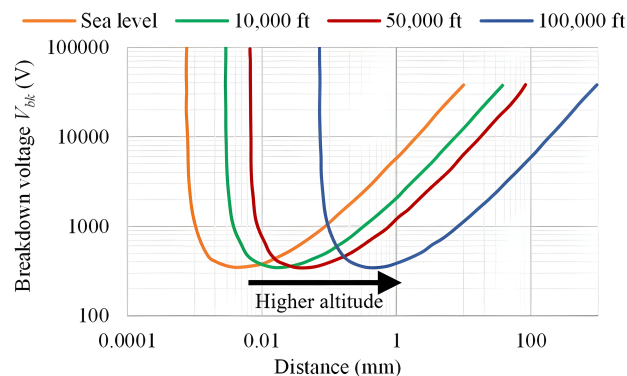


Fig. 2.5 Breakdown voltage at different altitudes [16].

## 2.3 Electric machines for aircraft propulsion

In order to select a suitable electric machine type and topology for aircraft propulsion applications, several key requirements must be satisfied to ensure their viability. More generally, any onboard device must meet the following performance criteria:

- **Specific power:** This is one of the most critical parameters in relation to operational cost, as aircraft are constrained by a maximum take-off weight. It is estimated that each additional kilogram of system weight increases costs by approximately US\$1000 [17]. For reference, [11] projects that specific power values exceeding 6.5 kW/kg are required to enable all-electric architectures for general and regional aviation.
- **Reliability:** All onboard components must comply with stringent safety and certification requirements in order to meet the strict regulations governing air transport.
- **Efficiency:** The losses associated with the operation of electric machines are directly related to the energy consumption of the primary energy source (e.g., fuel or batteries). Lower efficiency leads to increased operational costs.
- **Thermal management:** The ability to effectively dissipate heat from the device has a direct impact on the design of the cooling system, potentially increasing system complexity and weight.
- **Initial Cost:** In addition to operational expenses, the initial cost must also be considered, particularly with respect to material selection and manufacturing processes.

Based on the criteria described above, a brief overview of the different electric machines types is provided to assess their potential for use in aircraft propulsion systems.

### 2.3.1 Induction motors

Induction motors (IMs) are a well-established and reliable technology, offering simple construction and relatively low cost. They are capable of safely handling

overload conditions for limited durations and can be fully de-energized in the event of faults. This latter characteristic is particularly important in safety-critical applications, as it prevents uncontrolled generation operation of the machine due to induced voltages in the machine terminals.

However, compared to alternative machine types, IMs generally exhibit lower power factor and specific power, which often limits their suitability for aircraft propulsion applications. In addition, the presence of rotor currents increases thermal losses within the rotor, making heat dissipation and overall cooling more challenging.

Nevertheless, the literature reports the use of induction motors in aircraft starter and generator systems, owing to their capability to operate at high speeds and elevated temperatures compared to permanent magnet machines. They are also employed in applications such as fuel pumps, where high reliability and low maintenance requirements are critical [18–20].

### **2.3.2 Permanent magnet synchronous motors**

Permanent magnet synchronous machines (PMSMs) offer high power density and specific power due to the presence of permanent magnets in the rotor. This results in high efficiency and elevated power factor values, which in turn contribute to a reduction in the size of the associated power converter. Furthermore, PMSMs are well suited for high pole-pair configurations, enabling a reduction in the stator and rotor yoke dimensions and, consequently, in the overall weight and volume of the machine [21].

This type of machine can be implemented using either surface-mounted permanent magnets (SMPM) or interior permanent magnet (IPM) configurations. SMPM machines are characterized by high specific power and typically exhibit lower inductance values compared to IPM machines. This leads to a higher power factor but also results in increased short-circuit currents [22], [23]. However, in high-speed applications, mechanical retention systems such as retaining sleeves or rings may be required to ensure the structural integrity of the magnets [24]. Additionally, SMPM machines generally exhibit limited flux-weakening capability, leading to a reduced constant power speed range compared to IPM configurations.

Despite their advantages, PMSMs present certain limitations. In particular, they are susceptible to demagnetization under high-temperature conditions or in the

presence of large stator currents, such as during fault scenarios. Moreover, due to the permanent excitation provided by the magnets, these machines cannot be fully de-energized under fault conditions, which represents a potential drawback for safety-critical applications.

### 2.3.3 Reluctance motors

Synchronous reluctance (SynR) and switched reluctance (SR) machines are characterized by relatively simple rotor structures compared to other machine topologies. In their basic configurations, these machines do not include windings or permanent magnets on the rotor, resulting in enhanced robustness, as well as good overload capability and inherent fault tolerance.

According to the literature, SR machines represent a promising candidate for aircraft electrification applications due to their inherent modularity and improved fault-tolerant behavior, which is largely attributed to their concentrated winding layout [25]. Furthermore, SR machines are well suited for high-speed operation owing to their mechanically robust rotor structure. Their structural simplicity also contributes to reduced manufacturing costs and facilitates large-scale production.

However, SR machines are often limited by relatively low specific power and power factor compared to other machine types. Additionally, their rotor geometry can lead to unbalanced radial forces, which may increase mechanical stress and accelerate bearing wear [26]. Finally, as their operation relies on variations in magnetic reluctance, SR machines typically exhibit high torque ripple and increased acoustic noise during operation, which can be critical drawbacks in aerospace applications.

### 2.3.4 Wound field synchronous machines

Wound field synchronous machines (WFSMs) are widely used in conventional aircraft, primarily as onboard generators mechanically driven by the engine through a gearbox. In motoring operation, the ability to independently control the rotor field current allows this machine type to achieve an extended constant power speed range while maintaining high efficiency and power factor. Furthermore, the capability to regulate the field excitation enhances the overall controllability and fault tolerance of the machine.

However, WFSMs present several limitations. The presence of rotor currents complicates thermal management, resulting in more challenging cooling requirements. In addition, supplying current to the rotor necessitates the use of slip rings, multi-stage machine configurations, or wireless power transfer systems [27]. These additional components, along with the associated power electronics, negatively impact the specific power and increase the overall system cost.

Considering the advantages and disadvantages of each machine type discussed above, a qualitative comparison table can be constructed assigning a grade and corresponding color to represent how suitable is that machine for aircraft propulsion application regarding the category analyzed. Table 2.3 depicts this comparison where color green is associated with a good (G) performance, yellow with average (A) performance and red with poor (P) performance.

Table 2.3 Qualitative comparison of machines across categories.

| Category           | Induction | Permanent magnets | Reluctance | Wound Field |
|--------------------|-----------|-------------------|------------|-------------|
| Specific power     | A         | G                 | P          | A           |
| Reliability        | G         | A                 | G          | G           |
| Efficiency         | P         | G                 | A          | A           |
| Thermal management | P         | G                 | G          | P           |
| Cost               | G         | P                 | G          | A           |

## 2.4 Survey on electric motors for aircraft propulsion

It is relevant to review electric machines that have been employed in all-electric aircraft applications or specifically developed for such purposes. Several motors and generators reported in the literature, ongoing projects, and commercially available solutions have been investigated, and their main performance indicators are summarized. A comprehensive list of the relevant details is provided in Table A.1 in Appendix A. The selected machines were included only if sufficient data were available, such as rated power, specific weight, and rated speed, or if these parameters could be derived from the reported information.

Once the complete dataset of surveyed machines was established, a data analysis was performed to identify trends and patterns in key performance indicators, such as specific power and, when available, technology readiness level (TRL) [28–53]. In

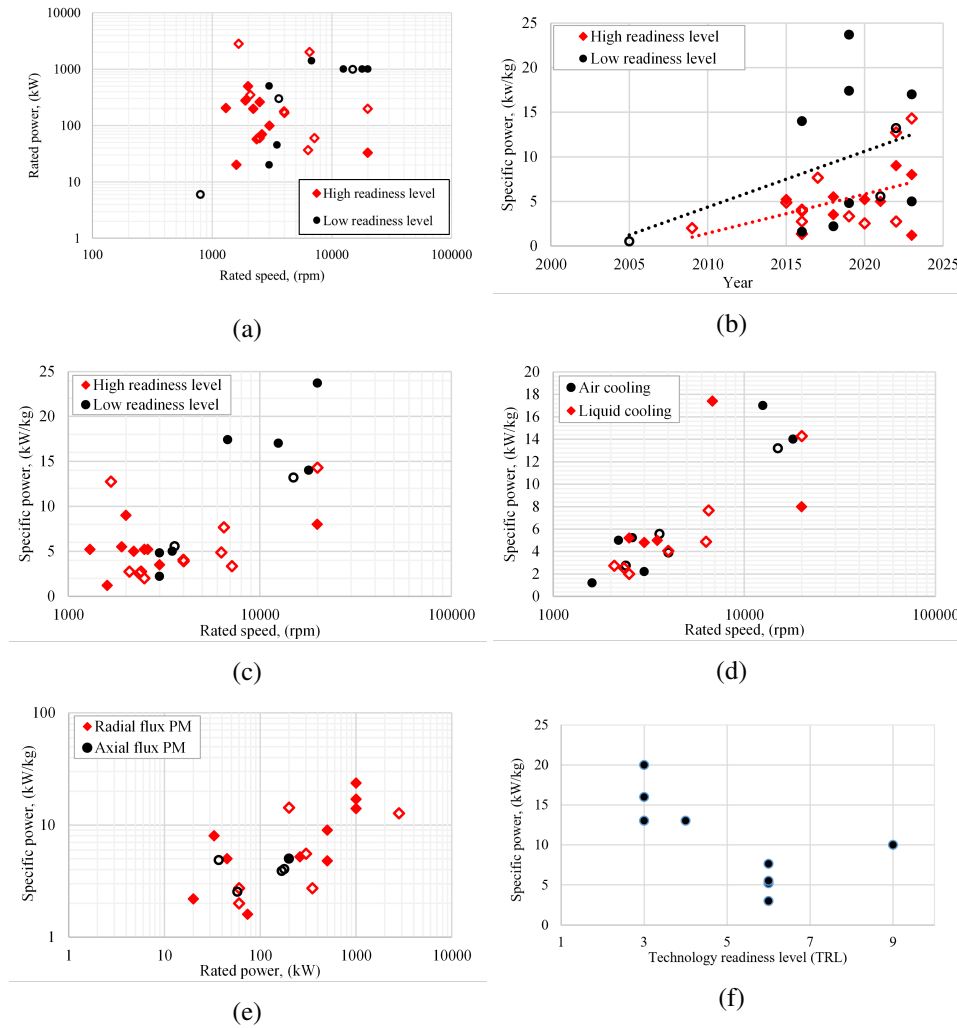


Fig. 2.6 Motor survey data findings. (a) Rated power vs. rated speed, (b) Specific power yearly growth, (c) Specific power vs. rated speed, (d) Specific power vs. rated speed for different cooling systems, (e) Specific vs. rated power for radial and axial flux PM motors and (f) Specific power vs. TRL

this context, Fig. 2.6 presents the results of the analysis, which are discussed in the following paragraphs.

As summarized in Fig. 2.6a, the rated power of the considered case studies ranges from tens of kilowatts up to several megawatts, with rated speeds approximately between 1-20 krpm. The figure also distinguishes between systems with high and low technological readiness levels. The former includes machines that have been installed on aircraft demonstrators or tested on laboratory rigs, while the latter comprises designs that remain at the research or simulation stage.



Figure 2.6b presents the relationship between specific power and development year. Although the available data are relatively scattered—particularly for low-TRL case studies—the figure includes linear trend lines to highlight general tendencies. Despite the low correlation observed in the data, these trends suggest that research-stage developments are typically 5–10 years ahead of machines that have been prototyped or are ready for testing on flight demonstrators.

The plot in Fig. 2.6c illustrates the relationship between specific power and rated speed for the analyzed solutions. As in the previous figures, the data are categorized according to the technology readiness level. The observed trend indicates that an increase in rated speed generally leads to higher specific power values, resulting in more compact machine designs. This is consistent with the fact that machine size is primarily determined by the required torque, which decreases with increasing speed for a given power rating.

However, at rotational speeds beyond those considered in the analyzed case studies, the increase in power density may be limited by additional factors, such as higher frequency losses in magnetic and conductive materials, as well as reduced heat dissipation capability due to smaller machine dimensions.

It is also worth noting that most of the analyzed solutions are concentrated in the lower speed range. This is likely due to the possibility of avoiding mechanical transmission systems to match load torque-speed requirements, thereby improving system efficiency and reliability while reducing weight and maintenance needs. Additionally, higher rotational speeds introduce challenges related to the dynamic and structural stability of the rotor.

Since machine performance and dimensions are strongly influenced by the adopted cooling method, Fig. 2.6d presents the relationship between specific power and rated speed for both air-cooled and liquid-cooled machines, where such information is available. The results indicate that liquid cooling is the preferred solution for achieving higher speeds and power densities. It should be noted that the few air-cooled machines exhibiting high speed and high specific power correspond to designs that are still at the research stage.

Figure 2.6e indicates that case studies with lower power ratings tend to exhibit lower specific power values. This observation is particularly relevant in the context of distributed propulsion systems [54], where the required propulsive power is provided by multiple smaller machines rather than a single large motor.

Nevertheless, [21] presents a broader survey of electrical machines, not limited to aircraft applications. Although the data are scattered and no clear trend can be definitively established, a general decrease in specific power with increasing rated power can be observed. This behavior can be attributed to the need to reduce electromagnetic and electrical loadings in order to maintain acceptable loss densities and comply with thermal constraints [55]. The data in Fig. 2.6e also indicate that axial flux machines achieve specific power values comparable to those of radial flux machines. However, for higher power levels, radial flux machines appear to be the preferred solution, at least among the considered case studies.

The reviewed research studies suggest that even higher specific power values are achievable; however, many of these developments remain at an early stage. This is further supported by Fig. 2.6f, which presents data from [56] on electrical machines suitable for aircraft propulsion, along with their corresponding technology readiness levels. The data clearly show that machines at higher stages of development tend to exhibit lower specific power values. This observation highlights that ongoing research and technological advancements in electrical machine design are essential to meet the specific power targets required for increasing the level of electrification in aircraft propulsion systems.

## 2.5 Why Axial Flux PM Machines for AEA?

The discussion presented throughout this chapter highlights that electric propulsion systems for aircraft must satisfy demanding and often competing requirements, including high efficiency, low weight, high reliability, fault-tolerant operation, and effective thermal management. These stringent requirements considerably reduce the number of electrical machine technologies suitable for fully electric aviation.

Among the available alternatives, axial flux permanent magnet (AFPM) machines have emerged as promising candidates for aircraft propulsion. Their achievable specific power is comparable to that of high-performance radial flux permanent magnet machines, particularly at low rotational speeds where direct coupling to the propeller is possible. Eliminating the gearbox not only reduces system weight but also improves drivetrain reliability and decreases maintenance requirements.

Beyond their competitive specific power, AFPM machines possess several characteristics that closely match the requirements of electric aircraft. Their disc-shaped geometry facilitates the design of machines with a high number of poles, allowing a reduction of the stator and rotor yokes while maintaining the required magnetic loading. Furthermore, in [57] is showed that, for appropriate pole/slot combinations with more than ten poles, the magnetic coupling between phases becomes negligible, improving the fault-tolerant capability of the propulsion system.

The axial flux topology also makes more effective use of the available active materials. Its larger effective air-gap radius increases the torque produced for a given machine volume, while multi-stage configurations enable the implementation of yokeless or coreless structures, further reducing weight and increasing specific power. In addition, multiple rotor-stator stages can be stacked while sharing the same magnetic circuit, providing a scalable solution without significantly increasing the machine diameter.

The compact axial geometry is another attractive feature for aerospace applications. The reduced axial length facilitates the integration of the propulsion unit within nose-mounted propeller systems and distributed electric propulsion architectures [54], where installation space is limited. AFPM machines are also well suited to fractional-slot concentrated windings, whose shorter end windings reduce copper usage and Joule losses, potentially improving efficiency, although at the expense of increased harmonic content and additional design challenges, as discussed in Chapter 3.

Despite these attractive characteristics, AFPM machines present additional challenges compared with conventional radial flux topologies. Their three-dimensional magnetic flux paths, complicate both the electromagnetic and thermal design processes, making accurate analytical and numerical modeling particularly important. The following chapters review the main technological aspects of AFPM machines, as well as presenting modeling and analysis approaches and their associated limitations for electromagnetic and thermal transfer phenomena.

## 2.6 Enabling technologies for AEA

Throughout this chapter, it has been mentioned the big challenge that is to reach a fully electrified commercial aircraft. With current technology available in the market is not feasible to apply the all electric aircraft concept for regional or intercontinental flights for commercial applications. However, several research regarding new technologies of the different powertrain subsystems are under development and appears as promising solutions to reach carbon neutrality in the air transport sector. In the following several of these technologies are presented and briefly discussed.

### 2.6.1 Battery technology

A primary limitation of current battery technologies for aircraft applications is their relatively low energy density. This results in heavier battery packs, which significantly constrain the achievable flight range. As discussed in [58] and [59], a minimum energy density in the range of 600–800 Wh/kg is considered necessary for regional commercial aviation. This target corresponds to approximately four times the performance of the current state of the art described in Section 2.2.1.

Several emerging battery chemistries are being investigated to bridge this gap. Among them, solid-state batteries are particularly promising, with theoretical energy densities exceeding 800 Wh/kg [60]. However, their large-scale manufacturing remains a significant challenge. Another potential candidate is lithium–air battery technology, which, according to [61], can achieve theoretical energy densities approaching 3000 Wh/kg, although these values are expected to decrease at the system level when considering complete battery pack configurations.

For this reason, lithium–air technology has attracted interest from major aerospace companies. In particular, Airbus and EADS have explored its potential application in the VoltAir concept, a fully electric aircraft envisioned for market entry around 2035 [62].

### 2.6.2 Electric motor technology

Regarding electric machines, the development of high-temperature superconductors (HTS) offers the potential to significantly increase achievable specific power levels.

Various machine configurations have been proposed, depending on the placement of the superconducting material within the device [63]. The main challenges associated with this technology include the complexity of the required cooling systems and the sensitivity of HTS materials to operating conditions. However, the increasing investment in hydrogen as a fuel for transportation may provide new opportunities, as cryogenic systems could be integrated and shared across different subsystems, including propulsion components. In terms of performance, superconducting machines are expected to achieve specific power values exceeding 40 kW/kg, with efficiencies above 99% [64]. These characteristics make them highly attractive for future aircraft propulsion systems, where both efficiency and weight are critical factors.

Bearing-less motor technology represents another promising approach to improving powertrain performance. By eliminating mechanical bearings and lubrication systems through electromagnetic suspension, these machines can enhance reliability while reducing mechanical losses. Additionally, they enable operation at higher rotational speeds, contributing to increased specific power [65]. Finally, the adoption of cryogenic cooling solutions for traditional electric machines may further enhance performance, potentially achieving specific power values of up to 25 kW/kg [66].

### **2.6.3 New materials developments**

Regarding advanced materials, several developments are expected to significantly impact the performance of all electric aircraft subsystems.

Among these, novel conducting materials such as carbon nanotubes (CNTs) represent a promising alternative to conventional conductors like copper and aluminum. CNTs exhibit theoretical electrical conductivity values that can exceed those of silver, along with superior thermal conductivity, which enhances heat dissipation. Additionally, their density is approximately six times lower than that of copper, offering substantial potential for increasing the specific power of electrical devices. However, the technology is not yet mature, and current production costs remain significantly higher than those of traditional materials. A practical intermediate solution involves combining CNTs with conventional conductors to improve overall performance while mitigating cost and manufacturability challenges [67].

Additive manufacturing (AM) also represents a promising approach for the design and production of electric aircraft propulsion subsystems. Commonly re-

ferred to as 3D printing, AM is a computer-controlled process in which components are built layer by layer. This technique offers several advantages over traditional manufacturing methods [68]:

- reduced material waste compared to conventional subtractive processes;
- enhanced design flexibility, enabling topology optimization and weight reduction;
- cost-effective production for low-volume and highly complex components;
- improved integration of multiple components, simplifying manufacturing and assembly processes.

Additive manufacturing can be applied using a wide range of materials, including polymers, metallic powders, and ceramic composites, making it suitable for various components of electric propulsion systems. As a result, AM shows significant potential for the development of lighter and more efficient electric machines.

## 2.7 Chapter conclusion

This chapter has reviewed the key aspects of aircraft electrification relevant to the scope of this thesis. From the historical perspective presented, a clear trend toward increasing onboard electrical power demand can be observed. This evolution has led to the adoption of the MEA concept, in which secondary systems are electrically powered to improve efficiency and reduce operating costs.

In recent years, growing interest in electric propulsion has further driven the development of AEA, which are already viable for general aviation and small-scale applications. However, extending this approach to commercial aviation requires significant advancements in powertrain subsystems, particularly in electric machine technology. From a qualitative standpoint, permanent magnet machines emerge as the most promising candidates due to their high efficiency, superior power and torque density, possibility to tailor its geometry in order to fulfill volumetric constraints and to reduce the weight adopting specific topology arrangements, as will be discussed in Chapter 3.

Within this context, the survey of electric machines for propulsion applications presented highlights axial flux topologies as particularly promising. Their compact geometry and inherently high specific power, especially in low-speed, direct-drive configurations, make them strong candidates for future aerospace propulsion systems, where weight and efficiency are critical design aspects.

## **Chapter 3**

# **Axial Flux Permanent Magnet Machine Technology**

The concept of axial flux electrical machines predates the widespread adoption of conventional radial flux topologies. Although modern interest in axial flux permanent magnet machines has surged only in recent decades, their origins can be traced back to the earliest developments in electrical machine design during the 19th century.

The earliest electrical machines, including Faraday's disc generator (1831), inherently exhibited an axial flux configuration. In these early devices, the magnetic flux flowed parallel to the axis of rotation, and torque was produced through interactions in a disc-shaped geometry. These designs were mechanically simple but suffered from low efficiency, poor material utilization, and limited electromagnetic performance due to the absence of advanced magnetic materials.

During the late 19th and early 20th centuries, axial flux machines were explored alongside radial flux machines. However, radial flux topologies quickly became dominant due to their superior manufacturability, robustness, and compatibility with laminated iron cores, which significantly reduced eddy current losses. The industrial standardization of cylindrical machine structures further marginalized axial flux configurations.

Between the 1920s and 1970s, axial flux machines received relatively limited attention in mainstream industrial applications. This decline can be attributed to several interconnected factors. First, material limitations—particularly the absence of high-energy permanent magnets—restricted achievable torque density and overall



performance. At the same time, manufacturing challenges played a significant role, as disc-type structures required precise mechanical alignment and were difficult to mass-produce using the technologies available at the time. Thermal management also posed difficulties, since cooling disc-shaped machines was less straightforward than in conventional cylindrical radial flux configurations. In parallel, the rapid development and industrial adoption of radial flux machines, especially induction and synchronous types, further reduced the appeal of axial flux alternatives. Consequently, axial flux machines were largely confined to niche or experimental applications during this period.

This situation began to change with the development of improved magnetic materials. The invention of Alnico PMs in 1931, followed by ferrite magnets in the 1950s, marked important milestones. However, the most significant breakthrough occurred with the introduction of rare-earth permanent magnets such as samarium-cobalt (SmCo) in the 1970s and neodymium-iron-boron (NdFeB) in the 1980s. These materials offered significantly higher energy density, enabling compact machine designs with improved torque output. This development, combined with growing demand for high power density and compact electrical machines, laid the foundation for the re-emergence of AFPM machines in modern engineering applications.

### **3.1 Axial flux machine configurations**

Axial flux machines have magnetic paths that cross the air gap in the axial direction as they close themselves in the rotor and stator yokes. They are characterized by a high diameter to length ratio, suitability for high pole count, and high power density. Axial flux machines can be designed to function with single or multiple air gaps (also referred as single- or multi-stage configurations, respectively) depending on the application requirements.

The most basic configuration is the single-stage where only one stator and one rotor is considered (fig. 3.1). The biggest drawback of this arrangement is the unbalanced magnetic pull that complicates the mounting process and increases stress and fatigue on the mechanical parts such as bearings, the rotor disk and housing.

The distinctive disk-type geometry of axial flux machines provides significant flexibility in terms of stator and rotor configurations, as well as in the arrangement

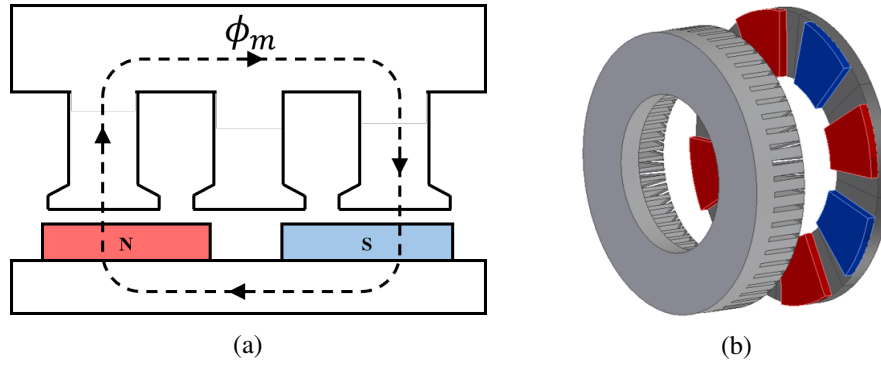


Fig. 3.1 Single-stage AFM. (a) 2D drawing with magnetic flux paths, (b) 3D representation.

of permanent magnets. Among the available designs, the double-stage axial flux machines (AFM) is the most commonly adopted in commercial applications, primarily due to the elimination of axial magnetic forces and the possibility of implementing coreless stator or rotor structures depending on the permanent magnet polarity distribution. Figure 3.2 illustrates the four possible configurations of double-stage AFMs.

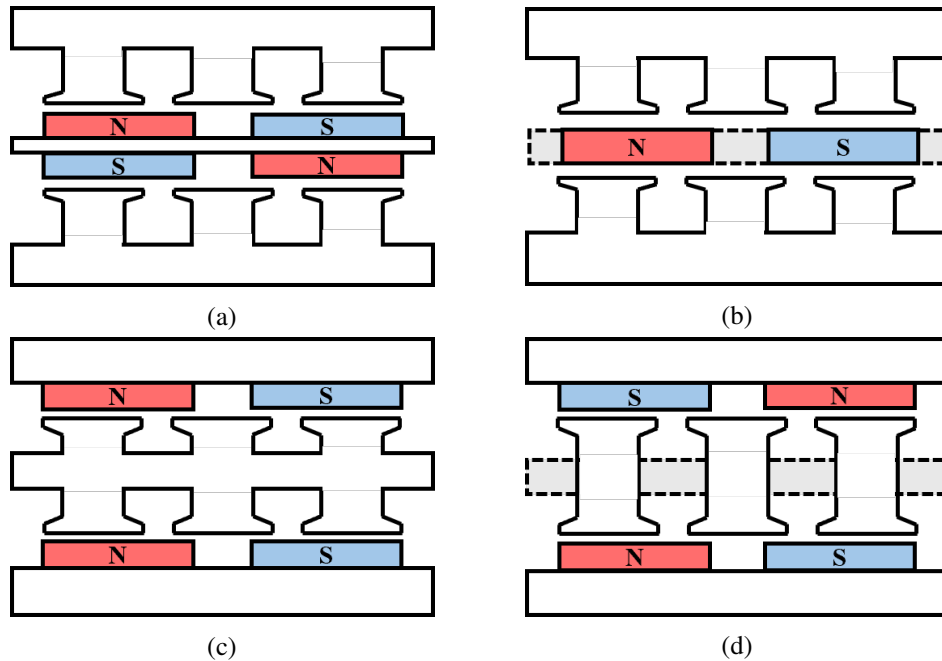


Fig. 3.2 AFM double stage configurations. (a) 2S1R-NS, (b) 2S1R-NN, (c) 1S2R-NN and (d) 1S2R-NS.

Based on the configurations shown in Fig. 3.2, four possible double-stage AFM topologies can be identified. Figure 3.2.a illustrates the two-stator, one-rotor (2S1R) configuration with opposing magnet polarities facing each other (NS). This arrangement requires magnetic cores in both the stator and rotor, leading to increased iron losses. In contrast, Fig. 3.2.b presents a 2S1R configuration with identical magnet polarities (NN) facing at the center, which eliminates the need for a rotor yoke, thereby reducing rotor losses and potentially decreasing machine weight.

The alternative configurations consist of a one-stator, two-rotor (1S2R) arrangement. In Fig. 3.2.c, the magnets facing each other have the same polarity, requiring a stator yoke to close the magnetic flux paths. Conversely, Fig. 3.2.d shows a 1S2R configuration with opposing magnet polarities, enabling a yokeless stator design and resulting in improved specific power and efficiency.

Finally, the inherent flexibility of axial flux topologies allows for the stacking of multiple rotor and stator stages to increase torque production, as illustrated in Fig. 3.3.

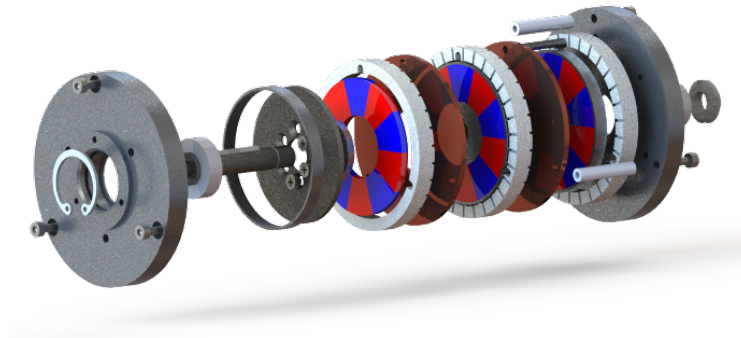


Fig. 3.3 Multistage AFM principle [69].

## 3.2 Radial flux versus axial flux machines

An important aspect to consider is the comparison between axial flux and radial flux machine topologies, in order to identify the application domains in which AFMs can outperform their radial counterparts. Based on the available literature, the main characteristics of axial flux machines can be summarized as follows [70], [71], [72]:

- higher diameter to length ratio than radial machines;

- incorporation of adjustable air gaps [73], [74], allowing more flexible operation over a wider speed range, although introducing significant mechanical challenges, particularly in larger machines [75].
- due to the larger outer diameters, AFMs can incorporate more pole pairs, potentially reaching higher torque density and making them more suitable for low speed applications [76], [77];
- better usage of active materials. Much of the rotor core around the shaft (usually the rotor yoke) is hardly utilized as a magnetic circuit.

In [70], a methodology for comparing double-stage AFPM machines with radial flux permanent magnet (RFPM) machines is presented. The comparison is based on simplified thermal considerations, evaluating both topologies in terms of electromagnetic torque and torque density under identical conditions of total machine volume, loss density per cooling surface, and air-gap flux density. The analysis focuses on low-speed, direct-drive applications, such as aircraft propulsion and electric vehicle traction.

The results indicate that AFPM machines are particularly advantageous in configurations with a high number of poles and short axial length. These findings are consistent with the data reported in [3] and [21], where axial flux machines demonstrate superior performance in low-speed operating regions, in some cases outperforming the surveyed radial flux machines.

Additionally, [72] presents a performance comparison between a conventional radial flux permanent magnet machine and several axial flux configurations across five different power levels. The results indicate that AF machines achieve lower volume and reduced active material mass for a given power rating compared to their radial counterpart.

### 3.3 Winding configurations for AFM

The operating principle of electrical machines is fundamentally based on the interaction between magnetic fields and electric currents within the machine windings. The specific configuration of the windings, including their spatial arrangement and electrical connections, in conjunction with the applied currents and voltages, governs

the mode of operation and ultimately determines the classification and performance characteristics of the electrical machine.

The type of winding can be determined with the slot per pole per phase  $q_s$  given by (3.1).

$$q_s = \frac{N_c}{p \cdot m} \quad (3.1)$$

Where,  $N_c$  is the number of slots,  $p$  the number of poles and  $m$  represent the number of phases. Based on the value computed with (3.1), the following possibilities will define the type of winding in an electric machine.

- $q_s \geq 1$ , integer: Distributed windings (DW);
- $q_s \geq 1$ , fractional: fractional slot distributed windings;
- $q_s = 1$ , integer: Concentrated winding;
- $q_s \leq 1$ , fractional: fractional slot concentrated winding (FSCW).

The selection of the this number is a critical aspect of electric machine design, as different values can lead to significantly different machine characteristics. In particular, key performance parameters such as the winding factor, cogging torque, vibration levels, rotor losses, and machine inductance are strongly influenced by the chosen combination of slot and pole numbers.

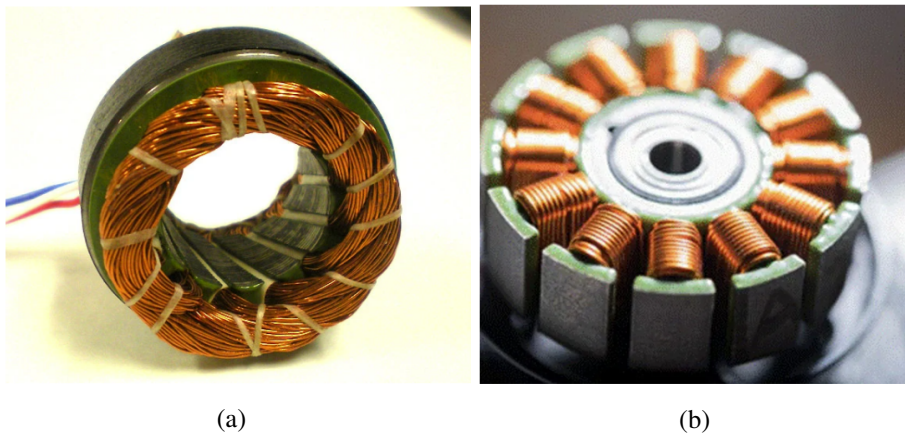


Fig. 3.4 Examples of Winding arrangements. (a) Distributed windings and (b) fractional slot concentrated windings [78].

In most applications, DW and FSCW (see Fig. 3.4) are the most common arrangements based on the typical slot/pole combination usually considered at the beginning of the design process. therefore, this work will focus on these arrangements and their corresponding features.

In distributed winding, the coils are spread across multiple slots, along the air gap periphery. This spatial distribution results in a near-sinusoidal magnetomotive force waveform (MMF), which is beneficial for achieving smooth torque production and minimizing harmonic content. Consequently, machines employing distributed windings generally exhibit low torque ripple, reduced acoustic noise, and lower additional losses associated with space harmonics. These features make distributed windings particularly suitable for applications requiring high efficiency and smooth operation, such as industrial drives and grid-connected machines. However, the distributed nature of the coils leads to longer end-windings, increasing copper usage and associated resistive losses, as well as overall machine size.

Concentrated windings means each coil is wound around a single tooth of the stator, instead of spreading it across multiple slots. This topology enables a highly compact design with significantly reduced end-winding length, resulting in lower copper consumption and improved power density. FSCW configurations are therefore widely adopted in applications where compactness and high torque density are critical, such as electric vehicles and axial flux machines. Nevertheless, the localized nature of the windings produces a non-sinusoidal MMF distribution with pronounced space harmonics. These harmonics can lead to increased torque ripple, higher acoustic noise, and additional rotor losses, particularly in PM machines due to eddy currents induced in the magnets.

Regarding axial flux machine topologies, the higher number of poles typically employed leads to a correspondingly large number of stator slots. However, the available space for slot allocation is limited due to the reduced pole pitch. As a consequence, achieving a slots per pole per phase ratio greater than unity—typically desired to obtain a more sinusoidal MMF and thus lower torque ripple and losses is often not feasible in high-pole-count AFMs. As reported in [79], when the stator axial length is smaller than the air gap diameter (diameter at the average of the outer and inner dimensions), fractional-slot non-overlapping concentrated windings become a favorable solution. In particular, [79] evaluates the ratio between air gap diameter and axial length for fractional-slot machines to determine the conditions

under which FSCW configurations are advantageous. The study compares full-pitch overlapping windings with fractional-slot non-overlapping windings, considering both torque production capability and copper losses. The results indicate that, for machines with a high number of poles, non-overlapping windings provide superior performance when the stator axial length is smaller than the air gap diameter.

To summarize the advantages and drawbacks of each winding topology, Table 3.1 presents a comparison of the different configurations with respect to key performance indicators.

Table 3.1 Comparative performance of distributed windings and FSCW

| Aspect                   | Distributed windings           | FSCW                                             |
|--------------------------|--------------------------------|--------------------------------------------------|
| MMF waveform             | Nearly sinusoidal              | Highly non-sinusoidal                            |
| Torque ripple            | Low                            | Higher (due to harmonics)                        |
| Copper usage             | Higher (longer end-windings)   | Lower                                            |
| Thermal management       | More distributed heat          | Heat concentrated in smaller area                |
| Manufacturing complexity | More complex winding insertion | Simpler, suitable for automation                 |
| Suitability for AFM      | Less suitable                  | More suitable (generally higher amount of poles) |
| Rotor losses             | Lower                          | Higher (due to harmonics)                        |

### 3.3.1 Number of winding layers

Another important concept related to the windings topology in electric machines is the number of layers. In general, the most common choices regarding this aspect are listed below.

- Single layer winding (SL): Each slot allocates conductors only from one coil (Fig. 3.5a).
- Double layer winding (DL): Each slot allocates conductors from two different coils (Fig. 3.5b).

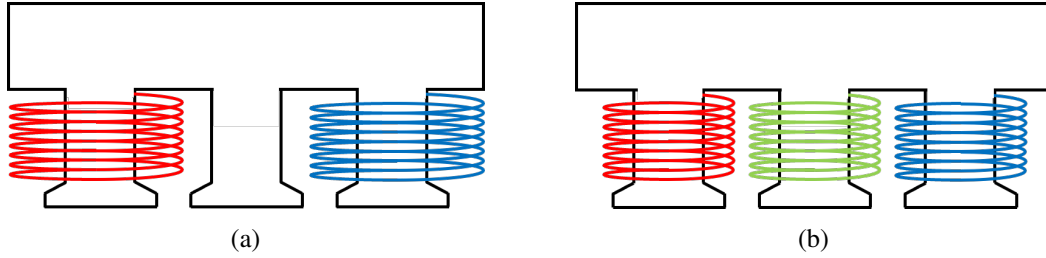


Fig. 3.5 Winding layout. (a) Single layer winding and (b) double layer winding [80].

Additionally, in [81] and [82] four layer windings has also been considered to further improve the performance of axial flux machines.

Single layer windings are particularly suitable for applications requiring high fault-tolerant capability. In this configuration, the phases are electrically and thermally isolated, short-circuit currents are inherently limited due to the high self-inductance, and magnetic coupling between phases is minimized as a result of the low mutual inductance. Furthermore, owing to their higher inductance compared to double-layer windings, SL configurations are preferred in applications that require a wide constant power speed range.

In some SL winding designs, teeth that do not carry coils are made thinner than those accommodating conductors. This approach allows the effective coil pitch to approach the pole pitch, leading to performance improvements such as increased torque capability—due to enhanced flux linkage and a higher fundamental winding factor—as well as reduced torque ripple [83], [84].

On the other hand, DL windings are generally preferred when minimizing torque ripple and losses is a primary objective. Additionally, DL configurations offer greater flexibility in selecting pole and slot combinations compared to SL windings. A comparison between SL and DL configurations is reported in Table 3.2.

It is important to note that not all  $q_s$  values are compatible with single-layer fractional-slot concentrated windings. As discussed in [85] and [86], both geometri-



Table 3.2 Comparison between single and double layer windings [80]

| Aspect            | Single layer    | Double layer    |
|-------------------|-----------------|-----------------|
| Winding factor    | Higher          | Lower           |
| Slot fill factor  | Higher          | Lower           |
| EMF               | Less sinusoidal | More sinusoidal |
| MMF harmonics     | Higher          | Lower           |
| Self inductance   | Higher          | Lower           |
| Mutual inductance | Lower           | Higher          |

cal and electrical constraints must be satisfied to enable a valid transformation from double-layer to single-layer winding configurations.

From a geometrical perspective, the number of slots  $N_c$  must be even, since the transformation involves removing half of the coils, and in DL windings the number of coils is equal to the number of slots. If this condition is not met, the resulting configuration leads to an uneven distribution, with some slots remaining empty while others contain conductors. Additionally, the slot pitch must be an odd number to ensure a uniform distribution of conductors across all slots after the transformation.

From an electrical standpoint, the feasibility of the transformation depends on the machine periodicity  $t_p$ , defined as  $t_p = GCD(N_c, pp)$ , where  $pp$  denotes the number of pole pairs and  $GCD$  is the Greatest Common Divisor.

### 3.3.2 Pole/slot selection and winding factor

The number of poles is directly related to the nominal speed or operating frequency of the machine. However, it must be selected in conjunction with the number of stator slots. In multiphase machines, only specific combinations of slot and pole numbers enable the realization of balanced winding configurations. These combinations are determined by the following condition:

$$\frac{N_c}{GCD(N_c, pp)} = m \cdot k \quad (3.2)$$

where  $k$  is a positive integer.

The average electromagnetic torque of a permanent magnet synchronous machine is directly proportional to the winding factor. A lower winding factor results in reduced torque production capability, which must be compensated by increasing either the current or the number of turns, both of which are inversely related to the winding factor. The winding factor of a given winding configuration is defined as the ratio between the magnetic flux effectively linked by the winding and the flux that would be linked by an ideal reference winding. This reference is typically a single-layer, full-pitch, non-skewed, integer-slot winding with the same number of turns and one slot per pole per phase.

The winding factor  $k_{w,h}$  for the  $h$ -th harmonic can be expressed as the product of three factors, the pitch factor  $k_{p,h}$  and the distribution factor  $k_{d,h}$  as shown in (3.3):

$$k_w = k_p \cdot k_d \quad (3.3)$$

The pitch factor accounts for the fact that windings are often not fully pitched; that is, the coil span is intentionally reduced to shorten the end-winding length and does not extend over a full pole pitch. The distribution factor, reflects the distribution of phase windings over multiple slots. Since the electromotive forces induced in different slots are not in phase, their phasor sum is smaller than their arithmetic sum. The winding factor can be calculated due to the contribution of each harmonic generated by the winding, thus both the pitch factor and distribution factor can be computed for the  $h$ -th as shown by (3.4) and (3.5) respectively.

$$k_{p,h} = \sin \left( h \frac{y}{y_s} \frac{\pi}{2} \right) \quad (3.4)$$

$$k_{p,h} = \frac{\sin \left( h \frac{\pi}{2m} \right)}{q_s \sin \left( h \frac{\pi}{2mq_s} \right)} \quad (3.5)$$

In (3.4)  $y$  is the coil span defined by the number of slots (which accounts for the short pitch, if any) and  $y_s$  is the number of slots per pole.

For each combination of pole and slot numbers, multiple winding layouts can be obtained by arranging the phase coils in different configurations within the slots. The selection of a specific winding arrangement has a direct impact on machine performance, as the synchronous-component winding factor ( $k_{w1}$ ) plays a key role

in the calculation of the average electromagnetic torque. In [80] and [87] a full description on the calculation on the winding factor considering all the harmonics is presented, along with tables with  $k_{w1}$  already computed for several slot/pole combinations.

## 3.4 Modeling and simulation tools

### 3.4.1 Electromagnetic torque production

In axial flux permanent magnet machines, the geometric dimensions vary with the radius (or diameter), and consequently, electromagnetic torque is produced over a continuous radial distribution rather than at a constant radius, as in conventional cylindrical machines. The pole pitch ( $\tau_p$ ) on an AFM with  $p$  number of poles evaluated at a given diameter  $D_i$  is defined as follows:

$$\tau_p = \frac{\pi D_i}{p} \quad (3.6)$$

A similar approach can be applied to determine the slot pitch ( $\tau_s$ ) for an axial flux machine with  $N_c$  slots, as shown in (3.7).

$$\tau_s = \frac{\pi D_i}{N_c} \quad (3.7)$$

For an axial flux machine, the tangential force acting on the rotor disk can be computed on the basis of the Ampere's equation as in (3.8), shown below [88].

$$d\mathbf{F} = K_s(r)(d\mathbf{S} \times \mathbf{B}_g) \quad (3.8)$$

where  $K_s(r)$  is the linear current density at a given radius,  $d\mathbf{S}$  is the vector surface element and  $\mathbf{B}_g$  is the vector of magnetic flux density in the air gap. Then, the torque can be computed from the tangential force at a specific radius  $r_i$  as:

$$d\mathbf{T} = r_i \times d\mathbf{F} \quad (3.9)$$

The magnitude of magnetic flux density in the air gap can be obtained from the PM remanence  $B_r$ , the magnet relative permeability  $\mu_{r,PM}$  and the air gap thickness  $h_{gap}$  as well as the air gap equivalent thickness  $h'_{gap}$  corrected by the Carter factor as follows,

$$B_g = \left( \frac{h_{gap}/\mu_{r,PM}}{h_{gap}/\mu_{r,PM} + h'_{gap}} \right) B_r \quad (3.10)$$

The Carter factor ( $KC$ ) is a dimensionless number that quantify the increase in air gap reluctance, and consequently the reduction in air gap magnetic flux—caused by the presence of stator or rotor slots. This number is define such that

$$h'_{gap} = KC \cdot h_{gap} \quad (3.11)$$

where  $KC$  can be computed with the slot dimension and air gap dimension. Different formulations of the Carter factor is presented in [89] and [90]. In this thesis, the formulation presented in (3.12) will be used.

$$KC = \frac{\tau_s}{\tau_s - \frac{2w_{so}}{\pi} \left( \arctan \frac{w_{so}}{2h_{gap}} - \frac{h_{gap}}{w_{so}} \ln \left[ 1 + \left( \frac{w_{so}}{2h_{gap}} \right)^2 \right] \right)} \quad (3.12)$$

Where,  $w_{so}$  is the slot opening. From (3.10), the maximum value of the fundamental air gap flux density  $B_{g1}$  (see Fig. 3.6), can be calculated as:

$$B_{g1} = \frac{4}{\pi} \sin \left( \alpha_{PM} \cdot \frac{\pi}{2} \right) B_g \quad (3.13)$$

Where  $B_g$  is the air gap flux density and  $\alpha_{PM}$  is the PMs occupation over the polar pitch which is considered constant with the radius. Hence, the fundamental air gap flux density value can be considered constant with the radius as well.

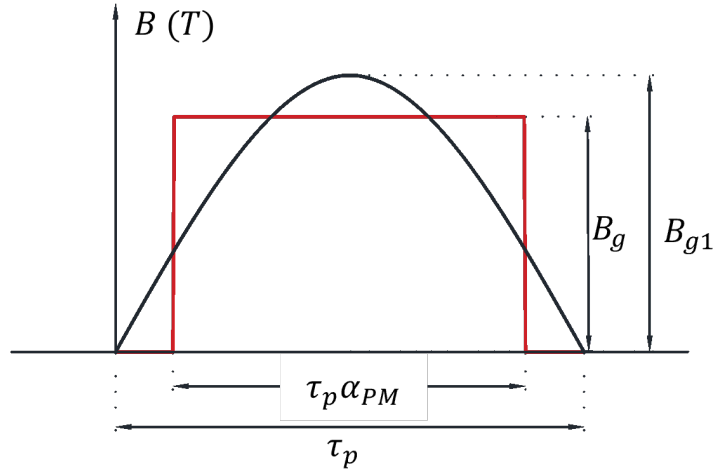


Fig. 3.6 Air gap flux density waveform over a pole pitch.

Then, the fundamental component of the linear current density at the diameter  $D$  can be computed as in

$$K_{s1} = k_{w1} J_{slot} \frac{A_{cu}}{\pi D} \quad (3.14)$$

with  $J_{slot}$  being the slot current density and  $A_{cu}$  being the copper area in the slot.

Finally, to get the electromagnetic torque produced by the AFM, (3.8) and (3.9) can be rewritten to obtain (3.15) that computes the electromagnetic torque element  $dT_{em}$  as:

$$dT_{em} = 2\pi K_s(r) B_g(\theta) r^2 dr \quad (3.15)$$

Considering only the fundamental components of the linear current density and the air gap magnetic flux density, and assuming that both are in phase to maximize torque production, the following expressions can be written:

$$B_g(\theta) = B_{g1} \sin \theta \quad ; \quad K_s(\theta) = \sqrt{2} K_{s1} \sin \theta \quad (3.16)$$

It should be noted that, in (3.16), the linear current density is expressed in terms of its RMS value and varies with the radial position, as described in (3.14). By

considering the average product of  $B_g$  and  $K_s$ , and substituting (3.14) into (3.15), the elemental electromagnetic torque becomes

$$dT_{em} = \frac{\sqrt{2}}{2} k_{w1} A_{cu} J_{slot} B_{g1} r, dr \quad (3.17)$$

The total electromagnetic torque is then obtained by integrating over the active radial region of the machine, from the inner radius  $R_1$  to the outer radius  $R_2$ :

$$T_{em} = \int_{R_1}^{R_2} \frac{\sqrt{2}}{2} k_{w1} A_{cu} J_{slot} B_{g1} r, dr \quad (3.18)$$

Finally, evaluating the integral and expressing the result in terms of the inner and outer diameters yields

$$T_{em} = \frac{\sqrt{2}}{16} k_{w1} A_{cu} (D_2^2 - D_1^2) B_{g1} J_{slot}. \quad (3.19)$$

with  $D_1$  and  $D_2$ , being the inner and outer diameter of the stator iron lamination respectively. In Chapter 4 the analytical procedure is further developed to derive a sizing methodology for determining the main dimensions of axial flux permanent magnet machines.

### 3.4.2 Cogging torque and torque ripple

When a permanent magnet synchronous machine is running, the torque produced is characterized by an oscillatory behavior for different reasons:

- cogging torque arises from the variation of the magnetic reluctance of the stator teeth as seen by the PMs as the rotor rotates;
- interaction between the spatial harmonics in the magnetic field produced by the permanent magnets and those generated by the stator windings;
- time harmonics induced in the phase current generated by power electronic converters;
- imperfections in the machine construction, such as eccentricity, non-uniform PM magnetization, and manufacturing tolerances.

Cogging torque is a periodic phenomenon whose period is determined by the least common multiple (LCM) of the number of stator slots and the number of poles. The cogging period,  $\theta_{cogg}$ , expressed in mechanical degrees, can be computed as follows:

$$\theta_{cogg} = \frac{360}{LCM(N_c, p)} \quad (3.20)$$

From (3.20), it is evident that the selection of the pole–slot combination during the machine design directly influences the cogging period. Additionally, several techniques have been investigated in the literature to reduce cogging torque, including design-based approaches, such as optimized winding configurations, magnet shaping as well as control strategies [91]. Among these methods, the introduction of intentional asymmetries in either the stator or the rotor has emerged as a particularly effective solution. By altering the magnetic flux distribution, such asymmetries can significantly reduce cogging torque and enhance the overall performance of the machine [92].

However, a low cogging torque does not necessarily imply a low torque ripple. Torque ripple is defined as the periodic variation of the instantaneous electromagnetic torque about its average value during machine operation. Unlike cogging torque, which is present under no-load conditions and results from the interaction between the permanent magnets and the stator slotting, torque ripple is typically evaluated under load and arises from the combined effects of several electromagnetic phenomena, including:

- spatial harmonics generated by the permanent magnets and those produced by the stator windings;
- magnetic flux saturation;
- time harmonics from power electronics;

The superposition of these phenomena produces periodic torque oscillations, leading to increased vibration and acoustic noise, reduced positioning accuracy, and potentially accelerated mechanical fatigue. Consequently, minimizing torque ripple is an important design objective, particularly in applications requiring smooth

operation. In general, single-layer windings exhibit higher torque ripple than double-layer configurations due to their less favorable MMF harmonic content.

### 3.4.3 Simulation approaches

In general, conventional radial flux machines can be effectively analyzed using two-dimensional finite element analysis (FEA), as their electromagnetic behavior is predominantly planar. In contrast, axial flux permanent magnet machines exhibit inherently three-dimensional flux paths, making 3D FEA the preferred approach for accurately predicting their electromagnetic behavior. While 3D simulations provides high-fidelity results, it entails significantly higher computational cost and simulation time compared to 2D analyses.

Furthermore, the computational effort is strongly influenced by the complexity of the machine design, including factors such as geometry, topology (e.g., cored or coreless structures), number of poles, number of stators, slot configuration, operating frequency, and rotor arrangement. As a result, the development of simplified and computationally efficient modeling approaches, capable of providing reasonably accurate predictions with reduced simulation time, is highly desirable for practical design and optimization processes.

For the two-dimensional modeling of axial flux machines, a commonly adopted approach consists of decomposing the axial geometry into multiple planar sections and performing separate FEA simulations for each slice [93]. In this method, each section is treated as an equivalent linear machine and analyzed independently. The overall performance of the axial flux permanent magnet machine is then obtained by combining the contributions of all slices through superposition. This procedure is illustrated in Fig.3.7.

In [94], an alternative methodology is proposed to simulate axial flux machines using multiple two-dimensional finite element models. As illustrated in Fig. 3.8, the original three-dimensional geometry is progressively reduced by exploiting symmetry and decomposed into several 2D representations. In addition to the linear motor modeling approach (LMMA), the method introduces two further modeling strategies.

The first is the inner rotor modeling approach (IRMA), in which the axial flux geometry is transformed into a set of equivalent 2D inner-rotor radial machines



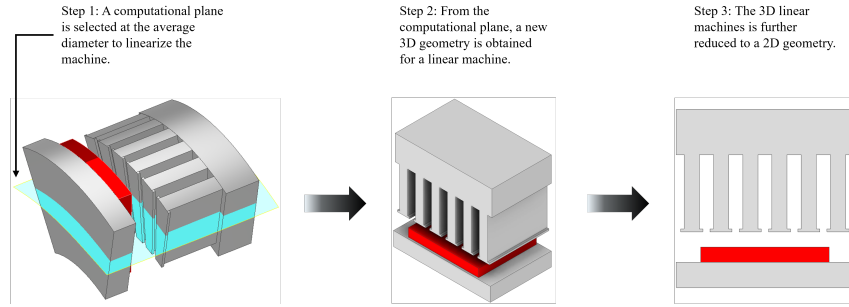


Fig. 3.7 FEA 2D modeling approach for Axial flux machines [93].

evaluated at different radial positions. The second is the outer rotor modeling approach (ORMA), where the machine is represented as a series of equivalent outer-rotor radial machines, also defined at different radii. The overall performance of the axial flux permanent magnet machine is then obtained by combining the contributions of these individual models through superposition.

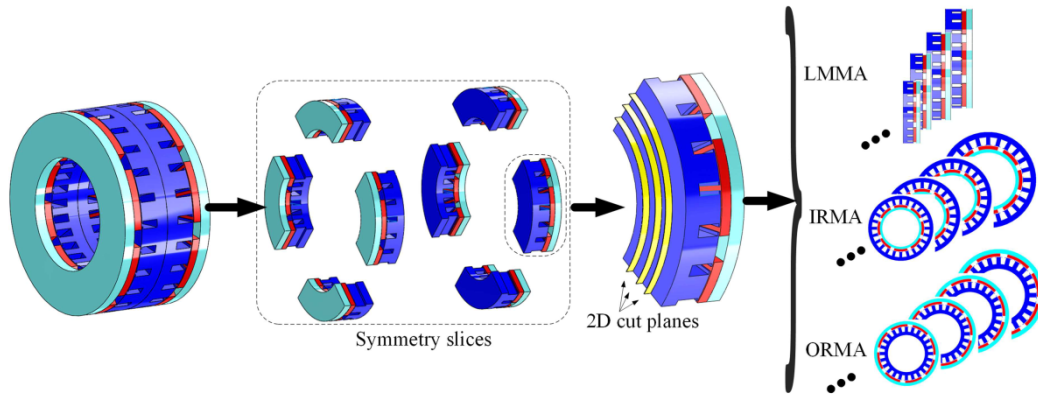


Fig. 3.8 FEA 2D modeling approach for AFM [94].

These methodologies demonstrate promising performance, offering a favorable trade-off between accuracy and computational efficiency. In particular, they achieve results comparable to full 3D simulations while requiring only a fraction of the computational time, making them attractive for design and optimization applications. Finally, although the IRMA and ORMA approaches achieve accuracy and computational performance comparable to those of the linear machine model, they offer the additional advantage of exploiting the 2D FEA templates widely available in commercial software, thereby accelerating the modeling process.

## 3.5 Thermal management

### 3.5.1 Losses in PM electric machines

The losses present in axial flux permanent magnet machines can be divided into the following categories:

- winding losses (Joule losses);
- iron losses (both in the stator and the rotor, when appropriate);
- permanent magnet losses;
- mechanical losses.

Figure 3.9 depicts a power diagram illustrating the losses for an axial flux machine working as a motor.

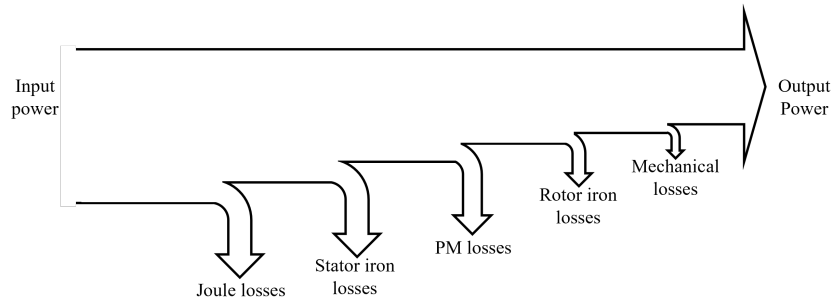


Fig. 3.9 Power flow diagram of a PM axial flux motor.

#### Stator losses

Copper losses correspond to the resistive (Joule) losses occurring in the stator windings of the machine. In a three-phase AC machine, the stator copper losses are expressed as follows:

$$P_{cu} = 3 \cdot R_s \cdot I_s^2 \quad (3.21)$$

where in (3.21),  $R_s$  is the stator winding phase AC resistance and  $I_s$  is the stator phase current.

In AFPM machines, when using concentrated windings, have the advantage of reduced length for the end-windings. Since the active part of the winding is arranged radially in a disc, the inactive end-turns are significantly shorter than in cylindrical machines.

In AC machines, additional phenomena such as skin and proximity effects can further increase the effective resistance, especially at higher frequencies.

The skin effect refers to the tendency of alternating current to concentrate near the surface of a conductor as the frequency increases. This phenomenon reduces the effective cross-sectional area available for current flow, thereby increasing the AC resistance. The main parameter governing this behavior is the skin depth  $\delta_{skin}$ , defined as in (3.22) [95]:

$$\delta_{skin} = \sqrt{\frac{\rho_{cond}}{\pi \cdot f \cdot \mu_{cond}}} \quad (3.22)$$

where  $\rho_{cond}$  is the conductor material electrical resistivity,  $f$  is the operating frequency and  $\mu_{cond}$  is the conductor material magnetic permeability.

When the conductor radius  $r_{cond}$  is comparable to or larger than the skin depth, the current density becomes highly non-uniform, leading to a significant increase in effective resistance, the AC resistance ( $R_{ac}$ ) can be computed by (3.23).

$$R_{ac} = k_{skin} \cdot R_{dc} \quad (3.23)$$

where,  $R_{dc}$  is the DC resistance of the winding (which can be measured or computed based on the conductor geometry).  $k_{skin}$  is the skin effect factor that can be estimated as follows [89],

$$k_{skin} \approx \frac{r_{cond}}{2\delta_{skin}} \quad (3.24)$$

The proximity effect arises from the influence of time-varying magnetic fields produced by adjacent conductors. These fields induce circulating eddy currents within a conductor, distorting the current distribution and further increasing the effective resistance.

The severity of the proximity effect depends on:

- conductor spacing and arrangement;
- operation frequency;
- magnitude of surrounding magnetic fields.

The impact of this phenomenon is more complex to quantify. Nevertheless, a methodology for estimating its effect is presented in [96].

For axial flux machines, although these phenomena are governed by the same physical principles, their design characteristics often exacerbate their effects. In particular, the use of high pole counts, fractional-slot concentrated windings, and inherently three-dimensional magnetic field distributions can increase AC copper losses. As a result, these losses become a more critical design consideration in AFPM machines than in comparable radial flux machines.

### **Iron losses**

Iron losses in the magnetic components of an electric machine, also referred to as core or magnetic losses, arise from the time-varying magnetic fields within the stator and rotor cores. In conductive ferromagnetic materials, these losses are typically decomposed into hysteresis losses and eddy current losses. Hysteresis losses are associated with the intrinsic magnetic properties of the material. When a magnetic material is magnetized by an increasing magnetic field and subsequently demagnetized by an opposing field, the magnetization does not follow the same path, resulting in a hysteresis loop, as illustrated in Fig. 3.10. The area enclosed by this loop corresponds to the energy dissipated per unit volume during each magnetization cycle and is therefore directly proportional to the hysteresis losses.

Eddy currents are induced in conductive magnetic materials as a result of time-varying magnetic flux, generating additional dynamic losses. These circulating currents oppose changes in the magnetic field, both in magnitude and direction. As a consequence, they contribute to increased energy dissipation and lead to an apparent broadening of the hysteresis loop, effectively increasing the magnetic coercivity of the material. These losses are reduced by using thin lamination or high-resistivity materials.

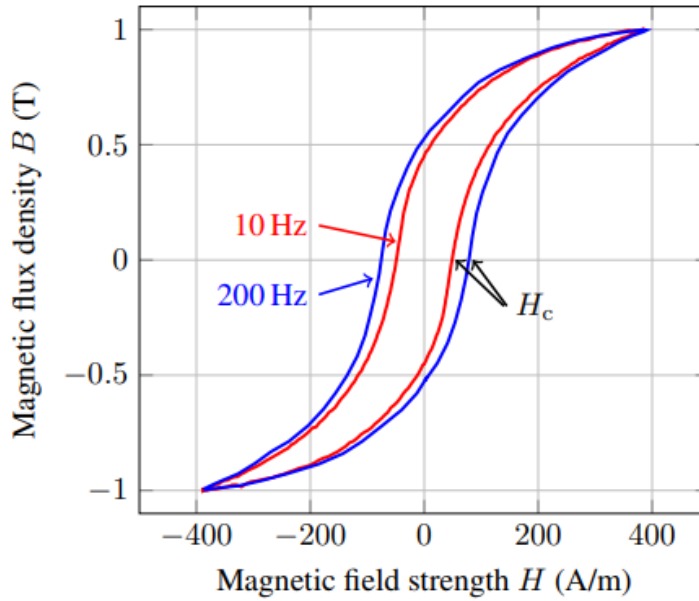


Fig. 3.10 B-H hysteresis curve of a ferromagnetic material for two different frequencies [80].

Regarding the different methodologies available to estimate the iron losses several approaches have been developed to compute them. The first group of models is based on the Steinmetz equation [97] which has the form given by (3.25).

$$p_{FE} = C_{SE} f_{op}^{\alpha} \hat{B}^{\beta} \quad (3.25)$$

where  $p_{FE}$  are the specific iron losses in W/kg,  $\hat{B}$  the peak value of the flux density in the sheet and  $f_{op}$  the operating frequency. The three coefficients  $C_{SE}$ ,  $\alpha$ , and  $\beta$  are determined by fitting the loss model to experimental measurements. It should be noted that the classical Steinmetz equation (3.25) is valid only for sinusoidal flux density waveforms. Consequently, several modified formulations, such as the one presented in [98], have been proposed in recent decades to extend its applicability to non-sinusoidal flux density waveforms, such as those generated by power electronic converters.

In addition to the various extensions of the Steinmetz equation, alternative iron loss models decompose the total losses into physically based components. In [99], a separation of the iron losses into hysteresis, eddy currents and excess component is considered as shown in (3.26).

$$p_{FE} = C_{hyst} f \hat{B}^\alpha + C_{ec} f^2 \hat{B}^2 + C_{exc} f^{1.5} \hat{B}^{1.5} \quad (3.26)$$

In (3.26),  $C_{hyst}$ ,  $C_{ec}$  and  $C_{exc}$  are the hysteresis, eddy current and excess loss coefficients, respectively. Again,  $\alpha$  is determined based on data fitting with available experimental measurements.

### Permanent magnet losses

Permanent magnet losses arise primarily from eddy currents induced within the magnets due to time-varying magnetic fields. Although permanent magnets ideally generate a constant magnetic field, in practical machines they are subjected to both spatial and temporal variations, which lead to additional losses.

Spatial harmonics originate from distortions in the magnetic field distribution along the air-gap circumference. These distortions are mainly caused by slotting effects, winding configurations, magnet geometry, and air gap non-uniformities. This phenomenon is particularly pronounced in machines with fractional slot concentrated windings and in high pole-count designs.

Temporal harmonics, on the other hand, represent variations of the magnetic field with respect to time at a fixed spatial location. They arise when the supply currents or voltages deviate from ideal sinusoidal waveforms, resulting in magnetic fields containing multiple frequency components. These harmonics are typically introduced by power electronic converters and associated control strategies.

The interaction between spatial and temporal harmonics has a significant impact on magnet losses, torque quality, and overall machine performance. A comprehensive analysis of permanent magnet losses, including a comparison of different modeling approaches reported in the literature, is presented in [100].

### Mechanical losses

Mechanical losses in an AC machine arise from mechanical effects associated with its operation and can be broadly classified into two main components: friction losses and windage losses. Friction losses are primarily attributed to bearing friction and are proportional to the product of friction torque and rotational speed. Windage

losses, on the other hand, result from the interaction between the rotating parts of the machine and the surrounding air within the enclosure. These losses are strongly dependent on rotational speed and are typically assumed to scale with the cube of this parameter.

### 3.5.2 Heat transfer principles

Heat transfer in electric machines occurs through three fundamental mechanisms: conduction, convection, and radiation. All three are driven by temperature gradients, transferring heat from regions of higher temperature to lower temperature. In axial flux machines, the complex three-dimensional geometry and the presence of multiple heat transfer paths make an accurate representation of these mechanisms essential for reliable thermal analysis. The following sections briefly describe each heat transfer mode and its governing equations.

#### Conduction

Conduction is the transfer of heat through a material due to particle interactions. In solids, heat is transmitted mainly by lattice vibrations and free electrons, whereas in fluids it results from molecular collisions and diffusion. The conduction heat rate depends on the material thermal conductivity, the geometry of the heat path, and the temperature gradient, and is given by

$$\dot{Q}_{cond} = -k_{th}A_c \frac{dT}{dx} \quad (3.27)$$

where,  $k_{th}$  is the thermal conductivity of the material,  $A_c$  is the cross-sectional area normal to the plane of conduction and  $dT/dx$  represents the temperature gradient across the medium.

#### Convection

Convection describes the heat transfer between a solid surface and a moving fluid, combining the effects of conduction and fluid motion. It can be classified as forced convection, when the fluid motion is externally driven, or natural convection, when

it results from buoyancy forces caused by temperature differences. The convective heat transfer rate is expressed as

$$\dot{Q}_{conv} = hA_s(T_{surf} - T_{amb}) \quad (3.28)$$

where  $h$  is the convection heat transfer coefficient,  $A_s$  is the surface area through which convection heat transfer takes place,  $T_{surf}$  is the surface temperature, and  $T_{amb}$  is the temperature of the fluid sufficiently far from the surface (usually the ambient temperature for electric machine thermal analysis).

The convective heat transfer coefficient is not an intrinsic property of the fluid. Instead, it is an experimentally determined parameter that depends on several factors influencing convection, including surface geometry, flow regime, fluid properties, and bulk fluid velocity. Nevertheless, the convection coefficient can be estimated analytically using empirical correlations developed for specific geometries and flow conditions. These formulations will be discussed in detail in Chapter 6, where a thermal model for axial flux permanent magnet machines is presented.

### Radiation

Radiation is the transfer of heat by electromagnetic waves and, unlike conduction and convection, does not require a material medium. The net radiative heat transfer between two surfaces is:

$$\dot{Q}_{rad} = \varepsilon\sigma_b A_s(T_{surf}^4 - T_{abs}^4) \quad (3.29)$$

where  $\varepsilon$  is the emissivity of the surface, that measure how closely a surface approximates a blackbody.  $\sigma_b$  is the Stefan–Boltzmann constant and  $T_{abs}$  is the absolute temperature of the surrounded surface.

In electrical machines, radiation typically plays a secondary role compared to convection and conduction, and this effect is commonly neglected in thermal analyses. However it can become much more relevant in situations with high temperature operation, enclose or poorly ventilated systems or vacuum or low pressure environments where convection is limited.



### 3.5.3 State-of-the-art of thermal modeling for AFM

Achieving high specific power and power density in electric machines requires careful consideration of thermal behavior during the design phase. Accurate prediction of the temperature distribution is particularly critical for permanent magnet machines, as excessive temperatures may lead to partial or irreversible demagnetization of the magnets, as well as accelerated aging of the winding insulation, ultimately degrading machine performance.

The thermal behavior of electric machines is traditionally evaluated using numerical methods such as finite element analysis and computational fluid dynamics (CFD). Although these approaches provide high-fidelity results, they involve significant computational cost and time, which limits their applicability in iterative sizing and optimization procedures typically required during early-stage design. To overcome this limitation, analytical methods can be employed to enable rapid thermal evaluation, though with a moderate reduction in accuracy. Among these, lumped parameter thermal networks (LPTNs) are widely used, as they extend circuit-based modeling concepts to represent heat transfer phenomena.

#### Lumped parameter thermal network

Lumped Parameter Thermal Networks are widely used for thermal analysis of electrical machines due to their computational efficiency and simplicity. In this approach, the machine is represented as a network of discrete thermal nodes connected by thermal resistances and capacitances, analogous to an electrical circuit.

In an LPTN model,

- temperature is equivalent to voltage;
- heat flow is equivalent the circuit current;
- thermal resistances model the opposition to heat transfer;
- thermal capacitance represents heat storage.

The governing equation to obtain the temperature distribution in the network is given by (3.30) shown below.

$$\mathbf{Q} = \mathbf{G}\mathbf{T} + \mathbf{C}\frac{d\mathbf{T}}{dt} \quad (3.30)$$

where  $\mathbf{Q}$  is the heat source vector representing the power losses within the machine,  $\mathbf{G}$  is the conductance matrix describing the thermal resistance interconnections of the network, and  $\mathbf{T}$  is the temperature vector containing the nodal temperatures of interest. Furthermore,  $\mathbf{C}$  denotes the thermal capacitance matrix, while  $d\mathbf{T}/dt$  represents the temporal variation of the temperature within the network.

For a thermal network with  $n + 1$  nodes, the thermal conductance matrix can be constructed as shown in (3.31).

$$\mathbf{G} = \begin{bmatrix} \sum_{i=1}^n \frac{1}{R_{1,i}} & \frac{-1}{R_{1,2}} & \cdots & \frac{-1}{R_{1,n}} \\ \frac{-1}{R_{1,2}} & \sum_{i=1}^n \frac{1}{R_{2,i}} & \cdots & \frac{-1}{R_{2,n}} \\ \vdots & \vdots & \ddots & \vdots \\ \frac{-1}{R_{n,1}} & \frac{-1}{R_{n,2}} & \cdots & \sum_{i=1}^n \frac{1}{R_{n,i}} \end{bmatrix} \quad (3.31)$$

Where  $R_{i,j}$  is the thermal resistance between nodes  $i$  and  $j$ . The general thermal conductivity resistance  $R_{th}$  is proportional to the thickness of the body  $l_{body}$ , and inversely proportional to thermal conductivity of the material  $k_{th}$  and the surface area  $A_s$  normal to the heat flow direction as defined by (3.32).

$$R_{th} = \frac{l_{body}}{k_{th}A_s} \quad (3.32)$$

For the capacitance matrix, it is build as follows,

$$\mathbf{G} = \begin{bmatrix} C_1 & 0 & \cdots & 0 \\ 0 & C_2 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & C_n \end{bmatrix} \quad (3.33)$$

With,  $C_i$  being the thermal capacitance connected to node  $i$ , that can be computed with (3.34) as shown below.

$$C_{th} = V_i \cdot \rho_i \cdot c_i \quad (3.34)$$

with  $V_i$  as the volume of the body,  $\rho_i$  the material density and  $c_i$  the specific heat capacity of the material.

### Finite element analysis

The Finite Element analysis method is a numerical technique widely used to solve heat transfer problems in electrical machines by discretizing the domain into smaller elements. In thermal analysis, FEM is primarily employed to model heat conduction within solid components, making it particularly suitable for evaluating temperature distributions in stator cores, windings, and permanent magnets. Unlike CFD, FEM does not directly solve fluid flow. Instead, convection is modeled using boundary conditions, where the convective heat transfer coefficient must be obtained beforehand by the means of analytical estimations or through experimental measurements.

### Computational fluid dynamics

Computational fluid dynamics enables the numerical analysis of fluid flow, mass transfer, and heat transfer phenomena, providing accurate results even for complex geometries. In contrast to lumped parameter thermal networks, CFD offers spatially resolved solutions, allowing the capture of local effects and detailed flow behavior. For the accurate modeling of convective heat transfer, particularly in axial flux permanent magnet machines, CFD is therefore a valuable tool.

The flow and thermal phenomena within AFPM machines are governed by the continuity, momentum (Navier–Stokes), and energy conservation equations. These governing equations are defined in (3.35), (3.36), and (3.37), respectively.

$$\nabla \cdot \mathbf{u} = 0 \quad (3.35)$$

$$\rho_{fluid} \left( \frac{\partial \mathbf{u}}{\partial t} + \mathbf{u} \cdot \nabla \mathbf{u} \right) = -\nabla p_{field} + \mu_{fluid} \nabla^2 \mathbf{u} \quad (3.36)$$

$$\rho c_p \left( \frac{\partial T}{\partial t} + \mathbf{u} \cdot \nabla T \right) = Q_{th} + k \nabla^2 T \quad (3.37)$$

where,

- $\mathbf{u}$  is the velocity vector field of the fluid;
- $\rho_{fluid}$  is the fluid density;
- $p_{field}$  is the pressure field;
- $\mu_{fluid}$  is the dynamic viscosity;
- $c_p$  is the thermal specific heat capacity;
- $T$  is the temperature field;
- $Q_{th}$  is the volumetric heat source.

### 3.5.4 Cooling methods for Axial flux machines

Electrical machine cooling can be classified according to the type of cooling fluid employed, the dominant heat transfer mechanism, and the machine components targeted for heat removal. The primary components requiring effective cooling in an electric motor are the stator core, the slot windings, and the end-winding regions. Fig. 3.11 illustrates different cooling methods that can be implemented in electrical machines.

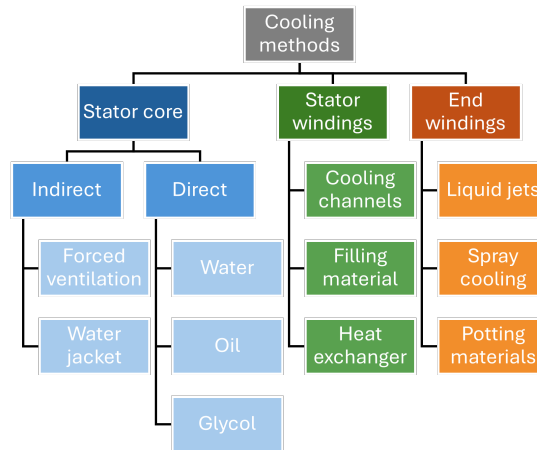


Fig. 3.11 Cooling techniques for electrical machines [101].

In the literature, several approaches have been developed to enhance the cooling of different regions in AFPM machines. In addition to conventional forced ventilation

systems commonly used in industrial machines, recent research has focused on innovative air-cooling techniques that promote heat extraction through the stator core.

In [102], a cooling system based on air jets is proposed, where airflow is directed through the stator and impinges on the rotor surface to improve heat transfer. In [103], an encapsulated air-cooling system is introduced, in which air is drawn through dedicated ventilation channels, forced across the stator core, and directed toward the rotor disk and permanent magnets before being expelled to the ambient. Furthermore, [104] presents a study on different fan blade configurations aimed at enhancing heat removal from the permanent magnets in a double-rotor axial flux machine.

With regard to liquid cooling approaches, [105] presents several techniques, including water jackets surrounding the stator and submersible cooling systems, in which the machine operates within a liquid medium such as water, oil, or a mixture of fluids. Additional methods include wet rotor and wet stator cooling, where a cooling fluid flows through the air gap and over the shaft and end-winding regions, enhancing heat extraction from both rotor and stator components.

For winding cooling, more targeted techniques have also been proposed. Spray cooling involves directing the cooling fluid through channels within the shaft and spraying it onto the inner surfaces of the end windings, often aided by centrifugal forces. Another approach is direct conductor cooling, such as slot water jackets, where the cooling fluid flows through the stator slots. In such configurations, internal ducts may be integrated within the slots to contain the fluid, or the fluid may flow directly between the conductors.

Finally, alternative cooling strategies have been proposed that, rather than enhancing heat removal through convection, focus on improving heat conduction by introducing additional thermal paths from hot spots to the ambient. In [106], a novel approach is presented in which copper bars are used as conductive heat transfer paths between the stator teeth and liquid cooling reservoirs located in the end shields, resulting in reduced stator temperatures. Similarly, in [107], a potting silicone gel is introduced in the gaps between the end windings and the machine casing of permanent magnet synchronous machines, enhancing thermal conduction and improving overall thermal management.

## 3.6 Chapter conclusion

This chapter has provided a comprehensive overview of axial flux permanent magnet machines, including a comparison with radial flux topologies that highlights their potential performance advantages in specific applications. Key aspects of winding configurations were discussed, with particular focus on distributed and fractional-slot concentrated windings, as well as single- and double-layer arrangements. In their synchronous PM topology, due to their suitability for high pole counts, AFMs are well suited to FSCW configurations, offering reduced end-winding length, lower copper losses and material usage, as well as enhanced fault tolerance and modularity features that are highly desirable for aircraft propulsion systems.

In addition, various modeling methodologies for axial flux machines were reviewed. While three-dimensional simulations are generally preferred for accurately capturing electromagnetic phenomena, they are computationally intensive. Although reduced-order and two-dimensional approaches exist, the inherently iterative nature of electric machine design, particularly when combined with optimization procedures, motivates the development of analytical tools that can provide sufficiently accurate results with significantly lower computational effort.

Finally, a review of loss mechanisms, heat transfer phenomena, thermal modeling approaches, and cooling strategies for electric machines was presented, establishing the foundation for the methodologies developed in the following chapters.

# Chapter 4

## Sizing tool for axial flux machines

The electromagnetic sizing of electric machines is often guided by the designer's expertise, prior experience, and available design databases. Subsequently, Finite Element analysis is typically employed to iteratively refine the machine geometry until the desired performance targets are achieved. However, this approach may require multiple design iterations, leading to a time-consuming process, particularly in the case of axial flux machines.

Indeed, the design and modeling of AFM topologies using FEA is more challenging than for conventional radial flux machines, owing to their inherently three-dimensional magnetic field distribution. Consequently, the development of a reliable analytical framework is essential to determine an initial geometry that closely approximates the final design, thereby significantly reducing the overall design time.

### 4.1 Lamination core sizing formulations

This chapter presents an electromagnetic sizing procedure for the preliminary design of AFPM machines. The proposed algorithm is inspired by the methodology introduced in [108], with its equations adapted to suit the axial flux topology. In addition, further formulations are incorporated to account DW and FSCW. The proposed approach is entirely based on analytical expressions and relies on a progressive increase of the machine diameters until all input constraints are satisfied and the desired performance is achieved (see fig. 4.1).

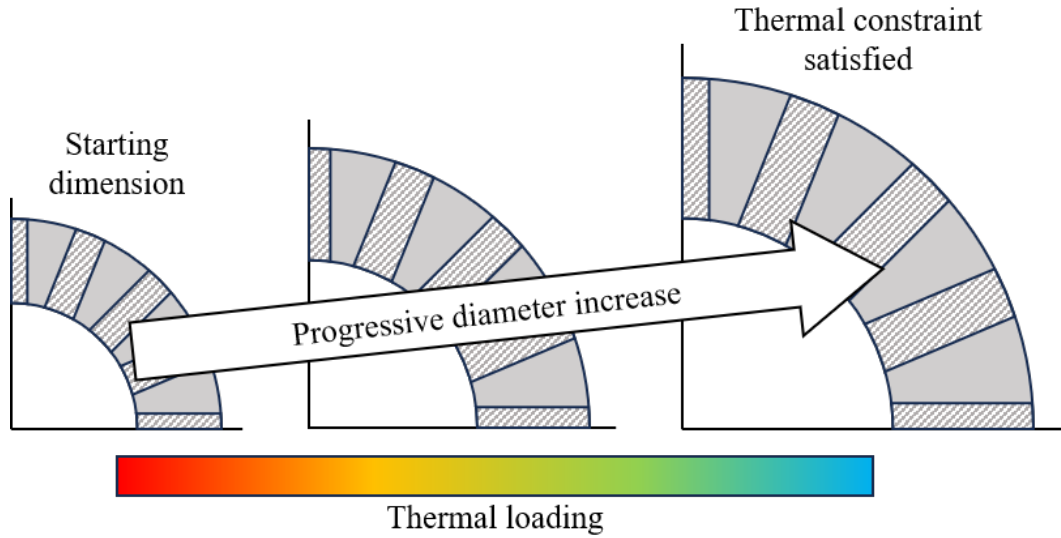


Fig. 4.1 Schematic illustration of the iterative sizing procedure for the AFM. The outer diameter is progressively increased until the thermal constraints are satisfied.

Furthermore, the method is not limited to single stator–single rotor configurations and can be readily extended to multi-stage axial flux machine arrangements. In particular, the lamination sizing procedure remains applicable to double-stage topologies (i.e., double stator or double rotor) with different pole-facing polarities in yoke-less designs. It should be noted that, for torus-type axial flux machines, the proposed procedure must be adapted to account for their distinct winding geometry, which is beyond the scope of this thesis. In particular, the underlying assumptions regarding the magnetic flux paths and the modeling of the windings differ from those adopted in this work.

#### 4.1.1 User inputs and design procedure

The flowchart of the proposed sizing procedure is presented in Fig. 4.2. The required input data include the target performance specifications, material properties, electromagnetic loading indicators, as well as selected geometric and winding parameters. Table 4.1 summarizes the inputs required to begin the sizing procedure.

Among the performance specifications, the torque is dictated by the application requirements and represents the primary design objective. The rotational speed (and consequently the electrical frequency) is used in the evaluation of the motor



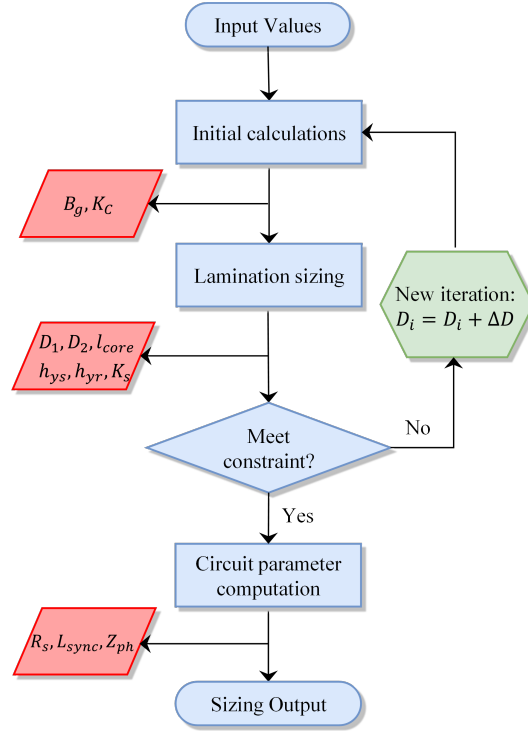


Fig. 4.2 Sizing algorithm flowchart.

equivalent circuit parameters. Similarly, the phase voltage is required to determine the number of conductors in series per phase.

The material specifications correspond to the designer's selection for the active components of the machine, including the magnetic laminations, windings, and permanent magnets. The electromagnetic loading indicators depend on factors such as the insulation class and the cooling system adopted. Several references in the literature provide typical ranges for different machine topologies and configurations, which can serve as guidelines for the initial selection [89], [90].

The aspect ratio,  $\lambda_{AFM}$ , is defined as the ratio between the inner and outer diameters of the stator. This parameter influences both the geometric shape of the machine, ranging from a more disc-like to a more ring-like structure, and its torque capability. For example, [109] suggests that  $\lambda_{AFM} = 1/\sqrt{3}$  maximizes the output torque. In the present procedure, however, this parameter is treated as a design degree of freedom, to be selected by the user according to dimensional constraints and allowable overall size.

Table 4.1 Sizing input values.

| Parameter                                   | Symbol                   | Unit              |
|---------------------------------------------|--------------------------|-------------------|
| <b>Target performance</b>                   |                          |                   |
| Torque                                      | $T_r$                    | Nm                |
| Speed                                       | $n_r$                    | rpm               |
| Phase voltage                               | $V_r$                    | V                 |
| <b>Material specifications</b>              |                          |                   |
| Lamination stack coefficient                | $k_{fe}$                 |                   |
| Permanent magnet remanence                  | $B_r$                    | T                 |
| Permanent magnet relative permeability      | $\mu_r$                  |                   |
| <b>Stress loading indexes</b>               |                          |                   |
| Max. teeth flux density                     | $B_{ts}$                 | T                 |
| Max. stator yoke flux density               | $B_{ys}$                 | T                 |
| Max. rotor yoke flux density                | $B_{yr}$                 | T                 |
| Max. linear current density                 | $K_{s,max}$              | A/cm              |
| Slot current density                        | $J_{rms}$                | A/mm <sup>2</sup> |
| <b>Geometric and winding considerations</b> |                          |                   |
| Aspect ratio                                | $\lambda_{AFM}$          |                   |
| Number of poles                             | $p$                      |                   |
| Number of slots                             | $N_s$                    |                   |
| Shaft diameter                              | $D_{sh}$                 | mm                |
| Air gap thickness                           | $h_{gap}$                | mm                |
| PM height                                   | $h_{PM}$                 | mm                |
| PM to pole ratio                            | $\alpha_{PM}$            |                   |
| Slot enclosure dimensions                   | $w_{so}, h_{ls}, h_{es}$ | mm                |
| Slot fill factor                            | $k_{cu}$                 |                   |
| PM radial length                            | $l_{PM}$                 | mm                |
| Winding pitch reduction as slot number*     | $n_{sr}$                 |                   |

\* Only for DW arrangements.

The pole–slot combination affects the air gap MMF distribution, the back-EMF waveform, and the noise and vibration characteristics during operation [110]. The shaft diameter is related to the required torque and can be determined using standard mechanical design formulations [111]. Within the proposed algorithm, this value is used as the initial inner diameter for the magnetic core sizing. Other geometric parameters, such as magnet dimensions, air gap length, and slot geometry, can be initially selected based on prior experience or available design data.

The developed sizing procedure is valid under the assumptions listed below:

- constant permeability of the magnetic materials;
- semi-open squared-shaped slots;
- single-stage and two-stage axial flux topologies.

The algorithm should be intended as a useful preliminary design tool to obtain a geometry that can be refined at later stages with more detailed methodologies.

#### 4.1.2 Magnetic core sizing

The main dimensions of the axial flux machine considered in the proposed sizing procedure are illustrated in Fig. 4.3. In particular, Figs. 4.3a and 4.3b present the radial dimensions of the rotor and stator, respectively, while the axial dimensions are shown in Fig. 4.3c.

The algorithm is based on an iterative inner diameter increase until the stopping criteria is met. The most inner diameter  $D_{im}$  is set to be equal to the shaft diameter for the first iteration. This dimension is smaller than the lamination inner diameter  $D_1$ , since it must account for the end-winding encumbrance in the stator. According to Fig. 4.3b, the angle  $\beta_{ax}$  is defined based on the winding arrangement as:

$$\beta_{ax} = \begin{cases} \frac{\pi}{2p} & \text{for } DW \\ \frac{\pi}{N_s} & \text{for } FSCW \end{cases} \quad (4.1)$$

The diameter  $D_1$  of the stator lamination can be obtained from the diameter  $D_{im}$  and the angle  $\beta_{ax}$  as follows:

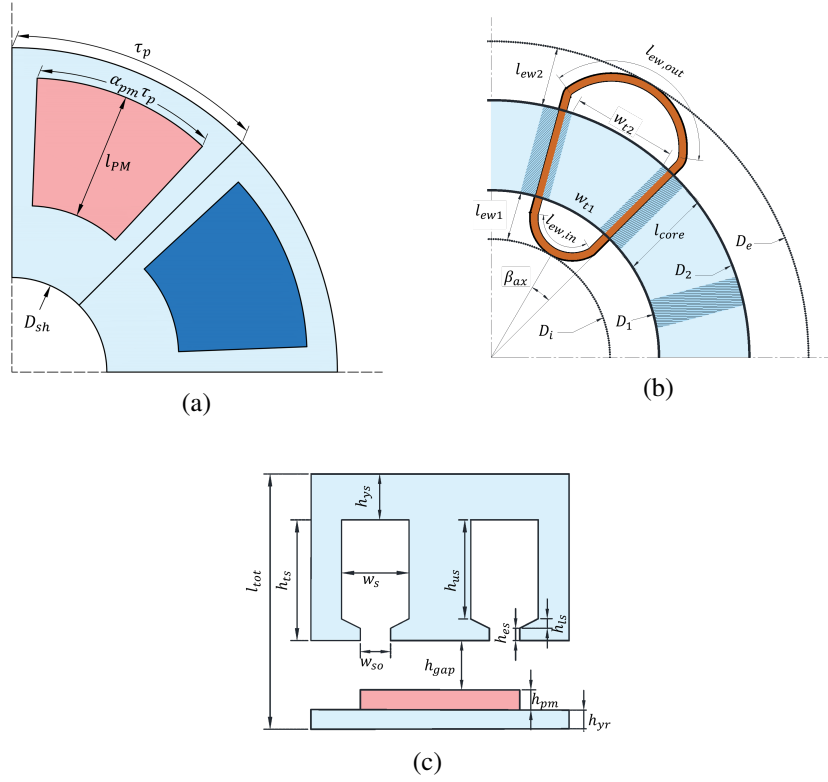


Fig. 4.3 Main dimensions of the AFM: (a) rotor structure, (b) stator and winding coil, (c) core, slot and magnet axial dimensions.

$$D_1 = \begin{cases} \frac{D_{im}}{\frac{1}{\cos \beta_{ax}} - \tan \beta_{ax}} & \text{if } p < 6 \\ \frac{D_{im}}{\cos \beta_{ax} - \sin \beta_{ax}} & \text{if } p \geq 6 \end{cases} \quad (4.2)$$

The outer diameter of the lamination  $D_2$  is imposed by the aspect ratio as shown in (4.3):

$$D_2 = \frac{D_1}{\lambda_{AFM}} \quad (4.3)$$

Then it is possible to calculate the outer-most diameter  $D_e$  with (4.4), while (4.5) gives the lamination stack length  $L_{core}$ .

$$D_e = D_2 \cdot (\sin \beta_{ax} + \cos \beta_{ax}) \quad (4.4)$$

$$L_{core} = \frac{D_2 - D_1}{2} \quad (4.5)$$

In axial flux machines, the magnetic flux is inherently non-uniform along the radial direction. Consequently, the flux per pole ( $\phi_{pole}$ ) can be calculated using (4.6), where  $L_{PM}$  denotes the radial length of the magnet, which can be approximated by  $L_{core}$ ,  $D_{ave} = (D_2 + D_1)/2$  represents the average stator diameter and the air gap flux density ( $B_g$ ) can be obtained using (3.10).

$$\phi_{pole} = \frac{\pi D_{ave} \alpha_{PM} l_{PM}}{2p} \cdot B_g \quad (4.6)$$

The stator and rotor yoke flux densities are higher at the outer diameter. Therefore, the yoke height is calculated at this location to satisfy the maximum lamination stress indexes specified by the designer. Assuming no leakage flux between the magnets, the lamination yoke heights for the stator ( $h_{ys}$ ) and rotor ( $h_{yr}$ ) are computed using (4.7) and (4.8), respectively. If the rotor yoke is not laminated, the lamination stacking factor  $k_{fe} = 1$  in (4.8).

$$h_{ys} = \frac{\pi D_2 \alpha_{PM} B_g / 2p}{2B_{ys} k_{fe}} \quad (4.7)$$

$$h_{yr} = \frac{\pi D_2 \alpha_{PM} B_g / 2p}{2B_{yr} k_{fe}} \quad (4.8)$$

Equations (4.7) and (4.8) are valid for single-stage configurations; for multi-stage machines, the yoke heights may be reduced or even neglected, depending on the selected configuration.

In axial flux machines, the teeth are more heavily stressed at the inner diameter  $D_1$ . Therefore, the inner tooth width  $w_{t1}$  is given by:

$$w_{t1} = \frac{B_g \pi D_1 / N_s}{B_{ts} k_{fe}} \quad (4.9)$$

Since only the squared shape slot is being considered in this work, the dimensions of the slot width  $w_s$  and upper tooth width  $w_{t2}$  can be computed as:

$$w_s = \frac{\pi D_1}{N_s} - w_{t1} \quad (4.10)$$

$$w_s = \frac{\pi D_2}{N_s} - w_s \quad (4.11)$$

As detailed in Chapter 3, the produced electromagnetic torque comes from the interaction between the air gap flux density and the current distribution across the stator slots. The average electromagnetic torque of the sized AFM can be calculated as in (3.19), considering  $k_{w1}$  as the fundamental winding factor,  $J_{slot}$  as the slot current density and  $A_{cu}$  as the copper area in the slot.

Therefore, equation (3.19) is used to determine the value of  $A_{cu}$  required to achieve the desired electromagnetic torque  $T_{em}$  as shown in (4.12), which is related to the output torque  $T_r$  through a factor accounting for friction losses.

$$A_{cu} = \frac{16T_{em}}{\sqrt{2} \cdot k_{w1} A_{cu} (D_2^2 - D_1^2) B_{g1} J_{slot}} \quad (4.12)$$

At this stage, by considering the conductor fill factor  $k_{cu}$  within the slot, the useful slot area  $A_{us}$  can be determined and the useful slot height  $h_{us}$  is obtained by dividing  $A_{us}$  by the slot width as depicted in (4.13) and (4.14) respectively.

$$A_{us} = \frac{A_{cu}}{k_{cu}} \quad (4.13)$$

$$h_{us} = \frac{A_{us}}{w_s} \quad (4.14)$$

Finally, the linear current density  $K_s$  is expressed as:

$$K_s = k_{w1} J_{slot} \left( \frac{N_s A_{cu}}{\pi D_1} \right) \quad (4.15)$$

It should be noted that  $K_s$  is evaluated at the inner diameter of the motor,  $D_1$ . Moreover, equation (3.19) is valid for single-stage AF machines; for double-stage configurations, the expected electromagnetic torque is approximately doubled [70].

### 4.1.3 Iterative procedure exit condition

The exit condition of the iterative design process is critical to obtaining a suitable electromagnetic machine design. For each external rotor diameter satisfying the maximum rotor yoke flux density constraint, the stator design routine determines a slot geometry that meets the required torque. However, at low rotor diameters, the air gap flux density and flux per pole are reduced. In high-torque designs, this may lead to impractical solutions with tapered slots (see Fig. 4.4).

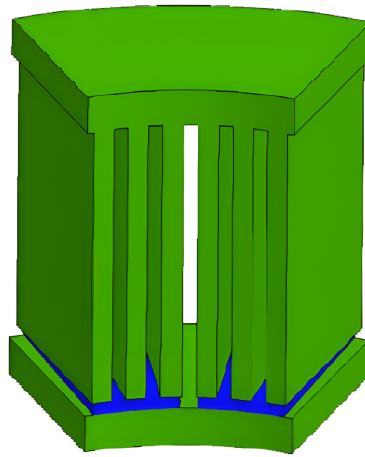


Fig. 4.4 Example of tapered slots in AFM [112].

To prevent this, the stopping criterion of the iterative procedure is based on the maximum allowable product of the linear current density  $K_s$  and the current density  $J_{slot}$ , defined as input constraints. This product is directly related to the thermal behavior of the winding, as slot over-temperature depends on Joule losses and heat dissipation. Therefore, it serves as an effective indicator to ensure both feasible geometry and acceptable thermal performance. Some reference values for both  $K_s$  and  $J_{slot}$  can be found on [89] and [90] depending on the cooling mechanism, topology and rated power of the machine to be designed.

At the initial stage of the iterative design process, the slot geometry tends to be tapered, requiring a high slot current, and consequently a high value of  $K_s$ , to meet the target torque. As the rotor diameter increases, the air gap flux density and the flux per pole generated by the magnets also increase. This reduces the required slot current, leading to wider and shallower slot geometries. As a result, the linear current density  $K_s$  decreases.

To achieve a balanced torque contribution from both the magnet flux and the stator current, the exit condition of the design procedure is defined by comparing, at each iteration, the product  $K_s \cdot J_{slot}$  with the constraint imposed by the designer. This product is adopted as the stopping criterion for the sizing algorithm, as expressed below:

$$K_s \cdot J_{slot} \leq K_{s,max} \cdot J_{slot} \quad (4.16)$$

Since the motor aspect ratio is fixed from the input, the left-hand side of (4.16) decreases with each step. Once the inequality is satisfied, the process is terminated, and the resulting dimensions are taken as the final sizing output.

## 4.2 Equivalent circuit parameter computation

Once the lamination sizing process is completed, the equivalent circuit parameters and the number of conductors in series per phase,  $Z_{ph}$ , can be determined. Figure 4.5 illustrates the equivalent phase circuit of an AF synchronous motor together with the corresponding phasor diagram.

In this representation,  $V_r$  is the rated phase voltage imposed as a design input,  $R_s$  is the phase winding resistance, and  $L_{sync}$  is the synchronous inductance. The phasor  $E_0$  denotes the no-load back-EMF, which is proportional to the mechanical speed and to the flux linkage produced by the permanent magnets, while  $I_s$  represents the current required to achieve the rated torque. These parameters are used to analytically determine  $Z_{ph}$  based on the imposed phase voltage.

From the phasor diagram in Fig. 4.5, and assuming that the phase current is entirely aligned with the q-axis, the rated voltage at an electrical angular speed  $\omega_e = 2\pi f_{op}$  can be expressed as in (4.17), where bold symbols denote phasor quantities.

$$\bar{\mathbf{V}}_r = \bar{\mathbf{E}}_0 + R_s \bar{\mathbf{I}}_s + j\omega_e L_{sync} \bar{\mathbf{I}}_s \quad (4.17)$$



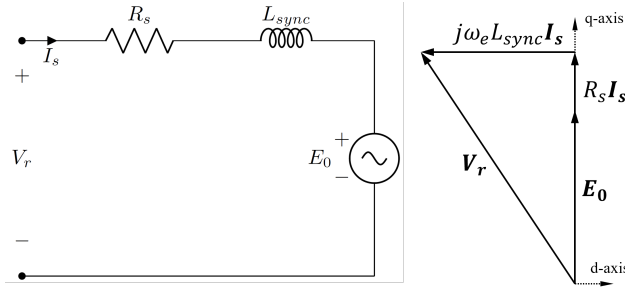


Fig. 4.5 Phase equivalent circuit for a no-salient PM synchronous motor (left) and the phasor diagram (right side).

### 4.2.1 Phase resistance

Considering a stator winding with  $Z_{ph}$  conductors in series per phase, the phase resistance  $R_s$  can be computed using (4.18), where  $\rho_{cu}$  is the winding resistivity at the operating temperature,  $q_s$  is the number of slots per pole per phase, and  $l_{turn,ave}$  is the average winding turn length.

$$R_s = \rho_{cu} \frac{Z_{ph}^2 l_{turn,ave}}{2 \cdot q_s \cdot p \cdot A_{cu}} \quad (4.18)$$

The value of  $l_{turn,ave}$  is computed as twice the sum of the slot length and the end-winding length  $l_{ew,ave}$  evaluated along its average path, as in (4.19). The calculation of this quantity depends on the winding configuration. For the end-winding, the average lengths for DW and FSCW are estimated using (4.20).

$$l_{turn,ave} = 2(L_{core} + l_{ew,ave}) \quad (4.19)$$

$$l_{ew,ave} = \begin{cases} \left(1 - \frac{n_{sr}}{3q}\right) \cdot \pi \left(\frac{D_1 + D_2}{4p}\right) & \text{for DW} \\ \frac{\pi w_s}{2} + \frac{\pi(w_{t1} + w_{t2})}{4} & \text{for FSCW} \end{cases} \quad (4.20)$$

### 4.2.2 Inductance

In rotating electrical machines, the total magnetic flux can be decomposed into two components: the main flux ( $\phi_m$ ) and the leakage flux ( $\phi_\sigma$ ). The main flux traverses

the air gap, linking the stator and rotor to generate the flux linkage in the windings responsible for electromechanical energy conversion. In contrast, the leakage flux remains largely confined to the stator windings, the rotor permanent magnets, or other local magnetic paths, without effectively crossing the air gap. Although it does not contribute directly to torque production, leakage flux gives rise to stator and rotor leakage flux linkages, increasing the amount of magnetic material or magnetizing current required compared to the ideal case without leakage.

The calculation of synchronous inductance in AF motors is a challenging task, as it requires a compromise between analytical simplicity and accuracy. Several methodologies for estimating this parameter are available in the literature. In this section, the approaches presented in [113] and [114] are adapted to account for the axial flux geometry. The synchronous inductance  $L_{sync}$  is expressed as the sum of separate contributions, as given in (4.21).

$$L_{sync} = L_m + L_{gap} + L_{slot} + L_{end} + L_{tt} \quad (4.21)$$

Where the components of the inductance are described below:

- $L_m$  is related to the useful flux component;
- $L_{gap}$  is related to the air gap leakage flux;
- $L_{slot}$  is related to the slot leakage flux;
- $L_{ew}$  is related to the end-winding component of the leakage flux;
- $L_{tt}$  related to the tooth tip leakage flux.

In the following, each of the components will be discussed along with its formulations.

### **Distributed windings**

When a three-phase symmetrical system of sinusoidal currents with amplitude  $I_s$  supplies a three-phase stator winding, a rotating air gap MMF distribution is produced. The coil inductance is generally defined as the ratio between the flux linkage of the coil and the current flowing through it that generates such flux. In the definition

of magnetizing inductance, the linked flux is that produced by the currents in all winding coils. Therefore, the concept of magnetizing inductance adopted in this section inherently accounts for the mutual coupling between phases, enabling the evaluation of the flux linkage of a given phase as a function of the corresponding current.

By neglecting the influence of the iron, the entire MMF distribution is assumed to drop across the air gap. Under this assumption, the amplitude of the  $h^{th}$ -order MMF harmonic is given by:

$$A_h = \frac{3}{2} \frac{Z_{ph} k_{w,h}}{h \cdot \pi \cdot pp} I_s \quad (4.22)$$

As a consequence, the magnetic flux density amplitude for the  $h^{th}$  harmonic order results:

$$B_{h(\alpha,t)} = \frac{3}{2} \frac{Z_{ph} k_{w,h}}{h \cdot \pi \cdot pp} \frac{\mu_0}{h'_{gap}} I_s \quad (4.23)$$

The main flux linked with the stator winding is given by the sum of all the harmonic contributions as shown in (4.24).

$$\Lambda_{main} = \sum_h \Lambda_h \quad (4.24)$$

where:

$$\Lambda_h = \frac{3}{2} \left( \frac{Z_{ph} k_{w,h}}{h \cdot \pi \cdot pp} \right)^2 \frac{\mu_0 \pi D_{ave} L_{core}}{2h'_{gap}} I_s \quad (4.25)$$

The main magnetizing inductance is therefore given by:

$$L_{m,main} = \sum_h \frac{3}{2} \left( \frac{Z_{ph} k_{w,h}}{h \cdot \pi \cdot pp} \right)^2 \frac{\mu_0 \pi D_{ave} L_{core}}{2h'_{gap}} \quad (4.26)$$

The magnetizing inductance expressed in (4.26) accounts for all air gap harmonic flux components. However, as discussed in the previous sections, only the fundamental flux component contributes to electromechanical energy conversion. The

remaining components, corresponding to higher-order harmonics, are conventionally treated as leakage flux. Accordingly, the air gap leakage inductance can be expressed as:

$$L_{gap} = L_{m,main} - L_{m,fund} \quad (4.27)$$

where,  $L_{m,fund}$  is the fundamental magnetizing inductance obtained by (4.26) substituting the harmonic order  $h = 1$ :

$$L_{m,fund} = \frac{3}{2} \left( \frac{Z_{ph} k_{w,h}}{\pi \cdot p p} \right)^2 \frac{\mu_0 \pi D_{ave} L_{core}}{2 h'_{gap}} \quad (4.28)$$

To evaluate the slot leakage inductance it is useful to refer to the magnetic energy stored in the slot defined by (4.29) below,

$$E_s = \frac{1}{2} \mu_0 \lambda_{slot} L_{core} I_{slot}^2 \quad (4.29)$$

where  $\lambda_{slot}$  is the slot leakage coefficient, and  $I_{sl}$  the total current in the slot. From 4.3c a rectangular slot shape is depicted, considering  $N_{cs}$  conductor in the slot. The magnetic energy stored into the slot can therefore be written in the following form:

$$E_s = \frac{1}{2} \mu_0 N_{cs}^2 [\lambda_1 + \lambda_2 + \lambda_{12}] I_{slot}^2 L_{core} \quad (4.30)$$

where the slot coefficient values can be computed as:

$$\lambda_1 = \frac{1}{3} \frac{h_{us}/2}{w_s} + \frac{h_{us}/2}{w_s} + \frac{h_{ls}}{\frac{w_s + w_{so}}{2}} + \frac{h_{es}}{w_{so}} \quad (4.31)$$

$$\lambda_2 = \frac{1}{3} \frac{h_{us}/2}{w_s} + \frac{h_{ls}}{\frac{w_s + w_{so}}{2}} + \frac{h_{es}}{w_{so}} \quad (4.32)$$

$$\lambda_{12} = \frac{h_{us}/2}{w_s} + 2 \frac{h_{ls}}{\frac{w_s + w_{so}}{2}} + 2 \frac{h_{es}}{w_{so}} \quad (4.33)$$

Considering the total magnetic energy stored in the slot written in the following form:

$$E_s = \frac{1}{2} L_{slot} I_s^2 \quad (4.34)$$

Then, for one phase of a distributed winding it is possible to evaluate the slot leakage inductance with (4.35):

$$L_{slot} = \mu_0 \frac{Z_{ph}^2}{4pq_s} \left[ \lambda_1 + \lambda_2 + \left( 2 - \frac{n_{sr}}{q_s} \right) \lambda_{12} \right] L_{core} \quad (4.35)$$

The end-winding leakage flux generally circulates around the end-winding conductors through air, although it may also pass through steel clamping rings, parts of the machine frame, or the end laminations of the stator. This inductance component is inherently a three-dimensional phenomenon, therefore, its accurate evaluation requires detailed 3-D FEA modeling, which entails significant computational effort. Additionally, conductive solid parts, where eddy currents may arise, influence the magnetic flux paths and consequently affect the leakage inductance. In contrast, other leakage inductance components can be estimated with reasonable accuracy using 2-D FEA models. For a three-phase distributed winding, it can be estimated using the empirical expression reported in (4.36) [115].

$$L_{ew} = (0.44 \div 0.52) \cdot \mu_0 \cdot p \cdot \left( \frac{Z_{ph}}{p} \right)^2 \cdot L_{ew,ave} \quad (4.36)$$

It should be noted that, in (4.36), the coefficient within the range 0.44–0.52 is an empirical end-winding geometry factor that accounts for the shape and arrangement of the end-winding region. For preliminary machine design, the midpoint value can be adopted as a nominal estimate. For an existing machine or prototype, however, the coefficient should be selected or calibrated according to the actual end-winding geometry to improve the accuracy of the inductance estimation.

### Fractional slot concentrated windings

For FSCW, a similar procedure to the one used for distributed windings can be applied to determine the magnetizing and leakage inductance components. The main

stator flux linkage (i.e., the flux linkage produced by the working harmonic of the stator single-phase current) is generated by each phase stator current  $I_s$  and the main inductance  $L_{sp}$ , while the other phases are not excited as presented in (4.37).

$$\Lambda_{sp} = k_{w,pp} \left( \frac{Z_{ph}}{2} \right) \frac{2}{\pi} \tau_p L_{core} B_{sp} \quad (4.37)$$

where,  $B_{sp}$  is the air gap magnetic flux density associated to that single phase stator current, which can be computed as,

$$B_{sp} = \frac{\mu_0 4 q_s (Z_{ph}/2)}{h'_{gap} \pi c_{ly} q_s p p} k_{w,pp} I_s \quad (4.38)$$

where,  $c_{ly}$  represents the number of winding layers in the slots. Therefore, the single-phase main inductance can be expressed using the peak values of the sinusoidal flux linkage and the phase current as:

$$L_{sp} = \frac{2}{\pi} \tau_p L_{core} \frac{\mu_0 4 q_s m}{h'_{gap} \pi N_s} \left( k_{w,pp} \frac{Z_{ph}}{2} \right) \quad (4.39)$$

The magnetizing inductance  $L_m$  of a rotating field winding is obtained by multiplying (4.39) by a factor  $m/2$ , which accounts for the mutual inductance between the phases, as the resulting current linkage wave is produced collectively by all  $m$  phases of the machine, i.e.,

$$L_m = \tau_p L_{core} \frac{\mu_0 4 q_s}{h'_{gap} N_s} \left( \frac{m}{\pi} k_{w,pp} \frac{Z_{ph}}{2} \right)^2 \quad (4.40)$$

It should be noted that, in this case, the main harmonic for the FSCW arrangement corresponds to the  $pp^{th}$  frequency (i.e., the harmonic associated with the pole pair of the AFM); therefore, the winding factor related to this harmonic is considered.

Due to the discrete nature of tooth-coil (TC) windings, the current linkage generated by such configurations results in a highly non-sinusoidal air gap waveform. This waveform contains a significant number of harmonic components, which give rise to harmonic magnetic fluxes [116]. These fluxes do not contribute to useful torque production; however, they increase the armature reaction in the windings

and, consequently, contribute to the leakage inductance. Accordingly, the air gap harmonic leakage inductance  $L_{gap}$  can be expressed as:

$$L_{gap} = \sigma_{\delta} L_m \quad (4.41)$$

where  $\sigma_{\delta}$  is the harmonic air gap leakage factor [116]:

$$\sigma_{\delta} = \sum_{\substack{h=1 \\ h \neq pp}}^{h=+\infty} \left( pp \cdot \frac{k_{w,h}}{h \cdot k_{w,pp}} \right)^2 \quad (4.42)$$

where  $k_{w,h}$  is the winding factor of the  $h^{th}$  spatial harmonic [116], [117]. The term  $h = pp$  corresponds to the synchronous harmonic and, therefore, to the magnetizing inductance component  $L_m$ . Table 4.2 reports the calculated air gap harmonic leakage factors  $\sigma_{\delta}$  for a few slot/pole combinations of FSCW PM machines. For these configurations, the number of slots per pole and phase  $q_s$  and the winding factors of the synchronous harmonic  $k_{w,h}$  component. A more comprehensive table with additional slot/pole combinations can be found in [114].

Table 4.2 Harmonic air gap leakage factor for FSCW PM machines [114].

| $N_c \backslash p$ |                   | 4     | 6     | 8     | 10    |
|--------------------|-------------------|-------|-------|-------|-------|
| 6                  | $q_s$             | 1/2   |       | 1/4   | 1/5   |
|                    | $k_{w,pp}$        | 0.866 |       | 0.866 | 0.5   |
|                    | $\sigma_{\delta}$ | 0.46  |       | 4.8   | 26    |
| 9                  | $q_s$             | 3/4   | 1/2   | 3/8   | 3/10  |
|                    | $k_{w,pp}$        |       | 0.866 | 0.945 | 0.945 |
|                    | $\sigma_{\delta}$ |       | 0.46  | 1.2   | 2.4   |
| 12                 | $q_s$             | 1     | 2/3   | 1/2   | 2/5   |
|                    | $k_{w,pp}$        |       |       | 0.866 | 0.933 |
|                    | $\sigma_{\delta}$ |       |       | 0.46  | 0.96  |

Unbalanced magnetic pull

Invalid configuration

Since the magnetizing inductance  $L_m$  results from the combined current linkages of all three phases, third-order harmonics are excluded from the analysis in a three-phase system. Moreover, Table 4.2 allows a comparison among different FSCW PM machine configurations. Combinations characterized by higher air gap harmonic leakage factors  $\sigma_{\delta}$  exhibit larger leakage inductances and typically wider field-

weakening regions, but lower maximum torque at nominal speed (assuming the same voltage limit). Additionally, a higher  $\sigma_\delta$  may indicate increased rotor losses due to asynchronous current linkage harmonics. Configurations with excessively high values of  $\sigma_\delta$  are generally not feasible, as their torque capability is significantly reduced by the high leakage inductance.

For the slot leakage inductance in FSCW, similarly to DW configurations, the current flowing in the slot conductors generates a magnetic flux, part of which crosses the air gap, while the remaining portion flows across the slot from one tooth to another. The latter component gives rise to the slot leakage inductance  $L_{slot}$ , which can be estimated as:

$$L_{slot} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 L_{core} \lambda_{slot} \quad (4.43)$$

where  $\lambda_{slot}$  is the slot leakage permeance factor [90]. The slot leakage permeance factor is calculated assuming that the leakage flux passes straight from one side of the slot to the other side crossing the center line of the slot in the orthogonal direction. The parameter  $\lambda_{slot}$  is calculated using the slot dimensions in Fig. 4.3c. The slot leakage permeance factor for a double-layer winding and horizontally divided layers can be calculated as [118]:

$$\lambda_{slot} = k_{1h} \left( \frac{h_{us} - h_{ly}}{3w_s} \right) + k_{2h} \left( \frac{h_{ns}}{w_s} + \frac{h_{es}}{w_{so}} + \left( \frac{h_{ls}}{w_s - w_{so}} \right) \ln \frac{w_s}{w_{so}} \right) + \frac{h_{ly}}{4w_{so}} \quad (4.44)$$

where the factors  $k_{1h}$  and  $k_{2h}$  take into account the mutual inductance when the coils of different phases are present in one slot [118]. Without making a significant error, 4.44) can be also used for vertically divided layers. The factors  $k_{1h}$  and  $k_{2h}$  are calculated as

$$k_{1h} = 1 - \frac{9}{16} \varepsilon \quad ; \quad k_{2h} = 1 - \frac{3}{4} \varepsilon \quad ; \quad \varepsilon = 1 - \frac{p}{N_c} \quad (4.45)$$

where  $\varepsilon$  represents the short pitching of the winding. The slot leakage inductance is modeled under the assumption that the leakage flux lines crossing the slot are straight. However, this assumption is not entirely accurate, particularly in the region close to the air gap, where the flux lines bend due to the proximity of the rotor iron



[119]. This fringing flux can be accounted for by introducing a tooth-tip leakage inductance model.

Several empirical and analytical approaches have been proposed in the literature to evaluate the end-winding leakage inductance. However, these methods often yield significantly different results and are predominantly developed for integer-slot distributed windings. For instance, an empirical analytical expression for calculating the end-winding leakage inductance  $L_{ew}$  for FSCW is presented in [90], i.e.,

$$L_{ew} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 q_s l_{ew,ave} \lambda_{ew} \quad (4.46)$$

where  $l_{ew,ave}$  is the average end-winding length (computed using (4.20)), and  $\lambda_{ew}$  is the end-winding leakage permeance factor. The permeance factor  $\lambda_{ew}$  is typically determined based on empirical values available in the literature for different distributed end-winding geometries [120].

However, for FSCW, the end-winding sections of a PMSM around each tooth can be approximated as halves of a solenoid. Under this assumption, the total end-winding inductance of a double-layer winding can be expressed as:

$$L_{ew} = \mu_0 \mu_{r,env} \frac{N_s}{m} \frac{(Z_{ph}/2)^2 \pi l_{ew,ave}^2}{c_{ly}^2 (q_s \cdot pp)^2 h_{us}} \quad (4.47)$$

The relative permeability  $\mu_{r,env}$  will depend on the surroundings of the end-windings.

The combined effect of slot and tooth-tip leakage inductances increases when the slot openings are narrow and the air gap is relatively long. Conversely, when the air gap is short and the slot openings are wide, as in PMSMs with open slots, their combined contribution decreases, as illustrated in Fig. 4.6. In this case, the flux within the slot tends to bend toward the rotor, reducing the slot leakage. This effect is captured by the tooth-tip leakage inductance  $L_{tt}$ . For large values of  $w_{so}/h_{gap}$  (open-slot configurations),  $L_{tt}$  may even assume negative values, thus requiring corrective adjustments in the slot leakage inductance calculation [114]. Accordingly,

$$L_{tt} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 L_{core} k_{2h} \lambda_{tt} \quad (4.48)$$

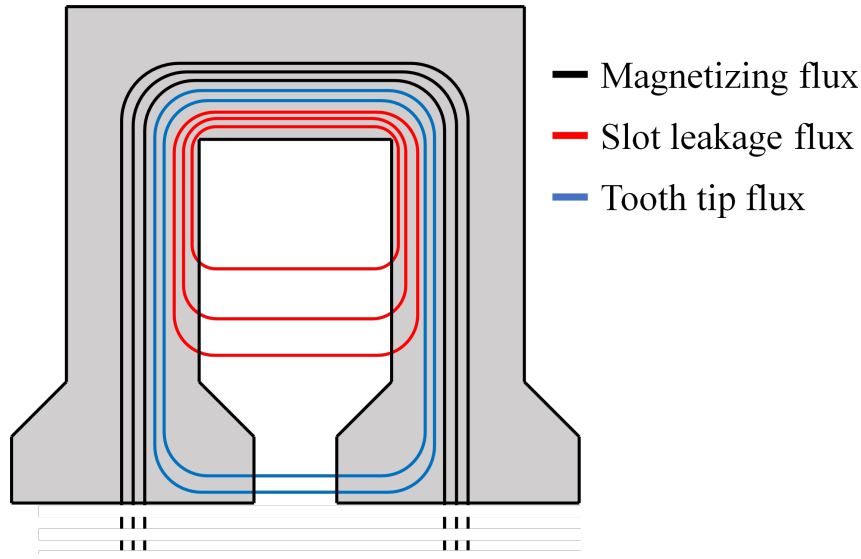


Fig. 4.6 Magnetizing and leakage flux in the stator.

where  $\lambda_{tt}$  is the tooth-tip leakage permeance factor, and  $k_{2h}$  is a factor that takes into account the presence of different phases in a slot (4.45). The permeance factor can be calculated according as:

$$\lambda_{tt} = \frac{1}{2\pi} \left[ \ln \left( \frac{h_{gap,PM}^2}{w_{so}^2} + \frac{1}{4} \right) + \left( \frac{h_{gap,PM}}{w_{so}} \right) \arctan \left( \frac{b_1}{2h_{gap,PM}} \right) \right] \quad (4.49)$$

For surface permanent magnet (SPM) configurations, the effective air gap length should account for both the physical air gap and the magnet thickness, the latter scaled by the relative permeability of the permanent magnet material, i.e.,  $h_{PM}/\mu_{r,PM}$ . In contrast, for interior permanent magnet (IPM) configurations, only the physical air gap length is considered in the tooth-tip leakage calculation.

For the reader's convenience, Table 4.3 summarizes all the formulations presented in this subsection for both winding arrangements considered.

Table 4.3 Inductance formulations for different winding arrangements.

| Inductance                       | Formulation                                                                                                                                  |
|----------------------------------|----------------------------------------------------------------------------------------------------------------------------------------------|
| <b>DW</b>                        |                                                                                                                                              |
| Synchronous ( $L_{sync}$ )       | $L_{sync} = L_m + L_{gap} + L_{slot} + L_{end} + L_{tt}$                                                                                     |
| Magnetizing ( $L_m$ )            | $L_m = \frac{3}{2} \left( \frac{Z_{ph} k_{w,h}}{\pi \cdot p p} \right)^2 \frac{\mu_0 \pi D_{ave} L_{core}}{2 h'_{gap}}$                      |
| Air gap leakage ( $L_{gap}$ )    | $L_{gap} = L_{m,main} - L_{m,fund}$                                                                                                          |
| Slot leakage ( $L_{slot}$ )      | $L_{slot} = \mu_0 \frac{Z_{ph}^2}{4 p Q} \left[ \lambda_1 + \lambda_2 + \left( 2 - \frac{n_{sr}}{q_s} \right) \lambda_{12} \right] L_{core}$ |
| End-winding leakage ( $L_{ew}$ ) | $L_{ew} = (0.44 \div 0.52) \cdot \mu_0 \cdot p \cdot \left( \frac{Z_{ph}}{p} \right)^2 \cdot L_{ewt}$                                        |
| Tooth-tip leakage ( $L_{tt}$ )   | $L_{tt} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 L_{core} k_{2h} \lambda_{tt}$                                               |
| <b>FSCW</b>                      |                                                                                                                                              |
| Synchronous ( $L_{sync}$ )       | $L_{sync} = L_m + L_{gap} + L_{slot} + L_{end} + L_{tt}$                                                                                     |
| Magnetizing ( $L_m$ )            | $L_m = \tau_p L_{core} \frac{\mu_0}{h'_{gap}} \frac{4 q_s}{N_s} \left( \frac{m}{\pi} k_{w,h} \frac{Z_{ph}}{2} \right)^2$                     |
| Air gap leakage ( $L_{gap}$ )    | $L_{gap} = \sigma_{gap} L_m$                                                                                                                 |
| Slot leakage ( $L_{slot}$ )      | $L_{slot} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 L_{core} \lambda_{slot}$                                                  |
| End-winding leakage ( $L_{ew}$ ) | $L_{ew} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 q_s l_{ew} \lambda_{ew}$                                                    |
| Tooth-tip leakage ( $L_{tt}$ )   | $L_{tt} = \frac{4m}{N_s} \mu_0 \left( \frac{Z_{ph}}{2} \right)^2 L_{core} k_{2h} \lambda_{tt}$                                               |

### 4.2.3 Conductors in series per phase

The no-load back-EMF can be computed using (4.50), with the operation frequency determined by the mechanical speed and the number of poles of the motor and where  $\phi_{pole1}$  is the amplitude of the fundamental component of the magnetic flux per pole.

$$E_0 = \frac{2\pi}{\sqrt{2}} f_{op} k_{w1} \phi_{pole1} \frac{Z_{ph}}{2} \quad (4.50)$$

The armature current  $I_s$  can be obtained dividing the slot current by the number of conductors in series in the slots.

$$I_s = (J_{rms} A_{cu}) \frac{N_s}{m Z_{ph}} \quad (4.51)$$

It can be noted that the above equations, as well as those related to the circuit parameters, depend on the value of the conductors in series per phase  $Z_{ph}$ . Therefore, these parameters can be expressed as in (4.52), where  $K_R$ ,  $K_L$ ,  $K_E$ , and  $K_I$  are obtained by rearranging the previous formulations.

$$\begin{aligned} R_s &= K_R \cdot Z_{ph}^2 & L_{sync} &= K_L \cdot Z_{ph}^2 \\ E_0 &= K_E \cdot Z_{ph} & I_s &= K_I / Z_{ph} \end{aligned} \quad (4.52)$$

Finally, using (4.17) and (4.52), the turns in series per phase can be computed as shown in (4.53), completing the whole sizing procedure.

$$Z_{ph} = \frac{V_r}{\sqrt{(K_e + K_R K_I)^2 + (\omega K_L K_I)^2}} \quad (4.53)$$

#### 4.2.4 Volume and weight computation for active materials

One crucial value to estimate is the specific power of the machine to assess its performance and to compare it with other candidates during the design workflow. For this purpose, the volumes and then, the mass of each part of the machine has to be calculated.

The volumes of the single-stage AFPM machine are evaluated by simple geometrical considerations starting from the main dimensions calculated in the electromagnetic iterative sizing process. The stator and rotor yokes occupied volumes are calculated by (4.54) and (4.55).

$$V_{ys} = \pi \left( \frac{D_2^2 - D_1^2}{4} \right) h_{ys} \quad (4.54)$$

$$V_{yr} = \pi \left( \frac{D_2^2 - D_1^2}{4} \right) h_{yr} \quad (4.55)$$

Similarly, the permanent magnet volume is calculated by (4.56).

$$V_{PM} = \pi \cdot \alpha_{PM} \left( \frac{D_2^2 - D_1^2}{4} \right) l_{PM} \quad (4.56)$$

The volume of the stator teeth are evaluated with the following:

$$V_{ts} = \left[ \pi \left( \frac{D_2^2 - D_1^2}{4} \right) h_{ts} \right] - V_{slots} \quad (4.57)$$

where  $V_{slots}$  is the total slots volumes evaluated by (4.60).

$$V_{slots} = A_{ts} N_c L_{core} \quad (4.58)$$

Taking into account the average turn length evaluated by (4.19), the stator winding volume is given by:

$$V_{cu} = A_{cu} N_c L_{turn.ave} \quad (4.59)$$

The total volume of the machine active materials is therefore obtained adding the previous quantities:

$$V_{tot} = V_{ys} + V_{yr} + V_{PM} + V_{ts} + V_{slots} + V_{cu} \quad (4.60)$$

Finally, to obtain the active material mass, the total iron volume, the conductor material and permanent magnets volumes are multiplied by the respective material density. This can be computed with (4.61), (4.62) and (4.65) as shown below:

$$m_{fe} = \rho_{fe} \cdot (V_{ys} + V_{yr} + V_{ts}) \quad (4.61)$$

$$m_{PM} = \rho_{PM} \cdot V_{PM} \quad (4.62)$$

$$m_{cu} = \rho_{cu} \cdot V_{cu} \quad (4.63)$$

The total active materials weight of the machine is obtained adding the previous quantities:

$$m_{tot} = m_{fe} + m_{PM} + m_{cu} \quad (4.64)$$

Finally, the specific power ( $SP$ ) of the machine (only considering the active materials) can be computed as follows

$$SP_{mot} = \frac{m_{tot}}{P_{mot}} \quad (4.65)$$

with  $P_{mot}$  being the rated motor power considered at the beginning of the sizing procedure.

### 4.3 Validation of the methodology

In order to validate the procedure, a comparison with both finite element simulations and experimental tests is carried out.

#### 4.3.1 FEA validation

The lamination sized with the proposed methodology has been 3D modeled and simulated on a commercial finite element software. Two different motors, referred to as M1 and M2, are used as case studies to validate the approach. These motors are characterized as follows:

- M1 is a 48-slot, 8-pole, double-stator (2S-1R) motor with DW.
- M2 is a 12-slot, 10-pole, single-stage (1S-1R) motor with FSCW.

The general dimensions and input parameters of the two machines are provided in Table 4.4. Figure 4.7 and Fig. 4.8 present the 3D-FEM models of the M1 and M2 machines, respectively. Comparative results, including air gap and core flux densities, back-EMF, torque, and inductance values, are summarized in Table 4.5 and Table 4.6. The FEA results show good agreement with the predictions obtained from the proposed sizing procedure.

In detail, the deviations in flux density are within 10%, while the differences in back-EMF and torque values are below 5%. The larger discrepancies observed in the

Table 4.4 Case studies design data.

| Parameter           | M1     | M2     |
|---------------------|--------|--------|
| Topology            | 2S-1R  | 1S-1R  |
| Slot number         | 48     | 12     |
| Pole number         | 8      | 10     |
| Inner diameter      | 120 mm | 128 mm |
| Outer diameter      | 192 mm | 221 mm |
| Air gap thickness   | 1 mm   | 1 mm   |
| Winding arrangement | DW     | FSCW   |
| PM height           | 4 mm   | 5 mm   |
| PM remanence        | 1.1 T  | 1 T    |

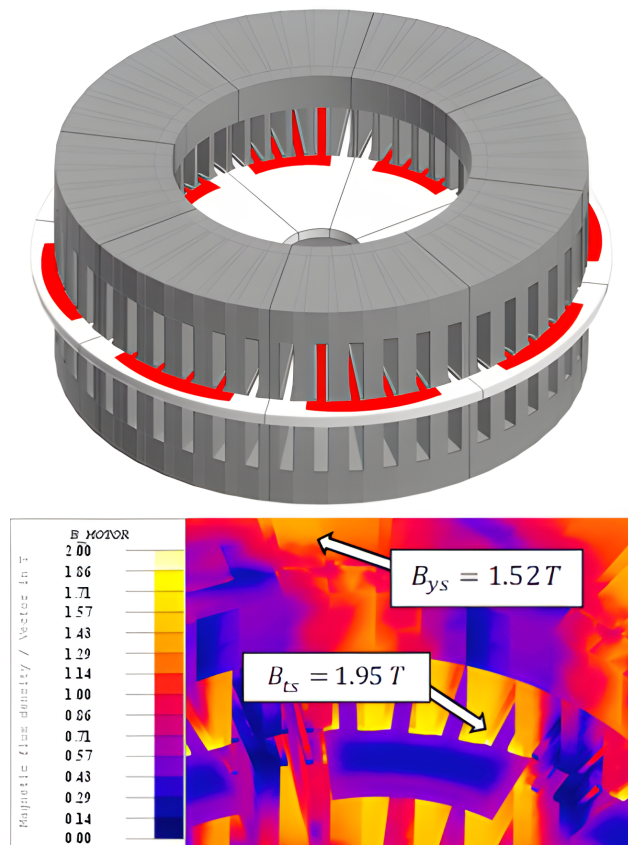


Fig. 4.7 3D model of the M1 AF motor (at the top) and magnetic flux density distribution (at the bottom).

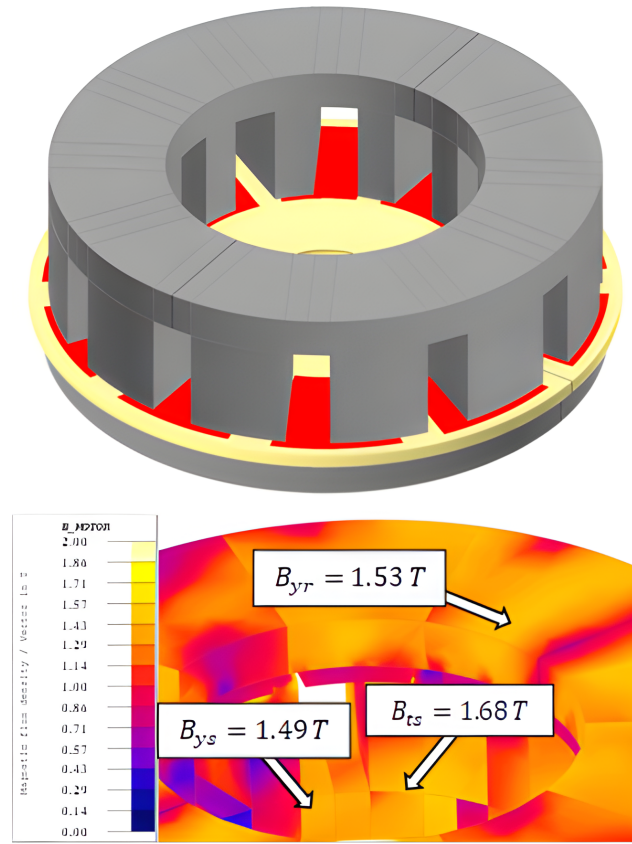


Fig. 4.8 3D model of the M2 AF motor (at the top) and magnetic flux density distribution (at the bottom).

Table 4.5 M1 FEA validation results.

| Parameter                  | Tool  | FEA   | $\Delta\%$ |
|----------------------------|-------|-------|------------|
| $B_{gap}$ , (T)            | 0.61  | 0.58  | 5.0%       |
| $B_{ts}$ , (T)             | 1.85  | 1.95  | -5.1%      |
| $B_{ys}$ , (T)             | 1.40  | 1.52  | -7.9%      |
| back-EMF (V)               | 103.9 | 100.5 | 3.4%       |
| Torque, Nm                 | 35    | 33.81 | 3.5%       |
| Synchronous inductance, mH | 96    | 102   | -5.9%      |



Table 4.6 M2 FEA validation results.

| Parameter                  | Tool | FEA  | $\Delta\%$ |
|----------------------------|------|------|------------|
| $B_{gap}$ , (T)            | 0.82 | 0.85 | -3.5%      |
| $B_{ts}$ , (T)             | 1.80 | 1.68 | 7.1%       |
| $B_{ys}$ , (T)             | 1.40 | 1.49 | -6.4%      |
| $B_{yr}$ , (T)             | 1.40 | 1.53 | -8.5%      |
| back-EMF (V)               | 95.9 | 92.8 | 3.5%       |
| Torque, Nm                 | 40   | 40.4 | -1.0%      |
| Synchronous inductance, mH | 24   | 21   | -14%       |

core flux densities can be primarily attributed to the accurate modeling of magnetic saturation in the FEA simulations, which is not accounted for in the analytical expressions used in the sizing algorithm. Additionally, variations in the selection of reference values within the FEA model, which may differ from point to point, contribute to these differences. In terms of performance, specifically back-EMF and torque, the results show closer agreement with the predictions from the proposed procedure, since they are less sensitive to the magnetic behavior assumptions of the lamination material (particularly the back-EMF). For the inductance value prediction, higher differences can be expected from analytical computations as the analytical approach relies heavily on empirical formulations and geometric simplifications introduced to ease the computational process. Finally, Fig. 4.9 and Fig. 4.10 depict the back-EMF for M1 and M2 from 3D simulations, respectively, while the torque of both motors during load conditions is shown in Fig. 4.11.

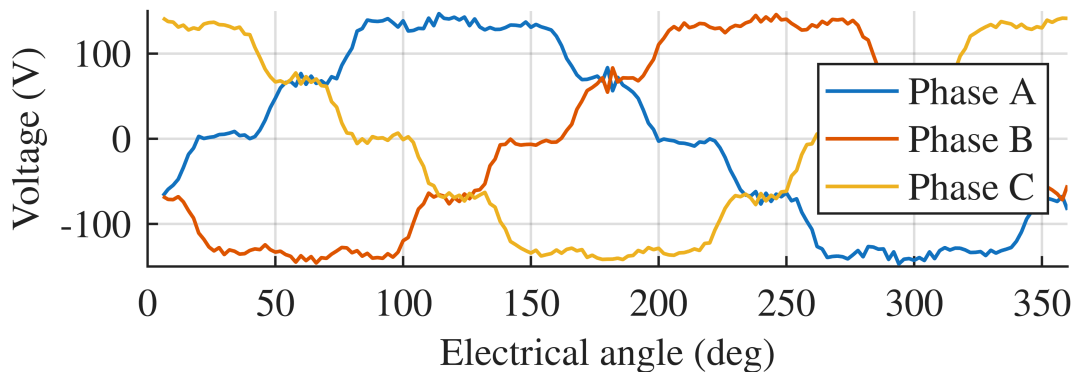


Fig. 4.9 Back-EMF simulation for M1.

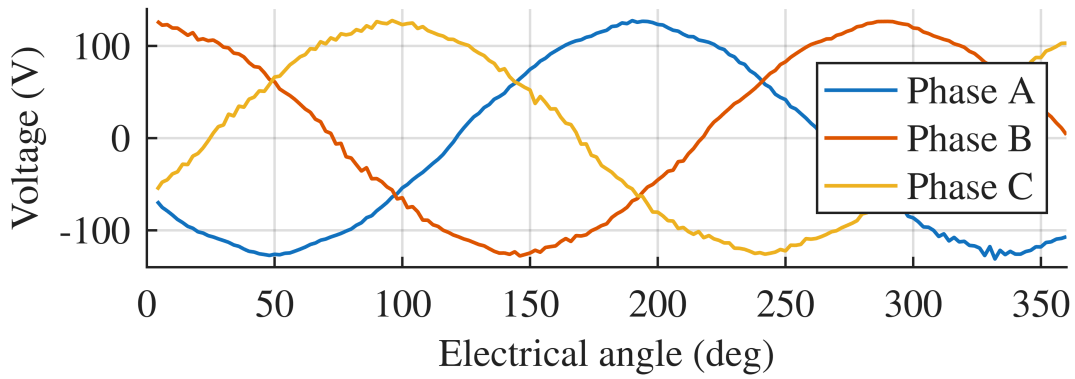


Fig. 4.10 Back-EMF simulation for M2.

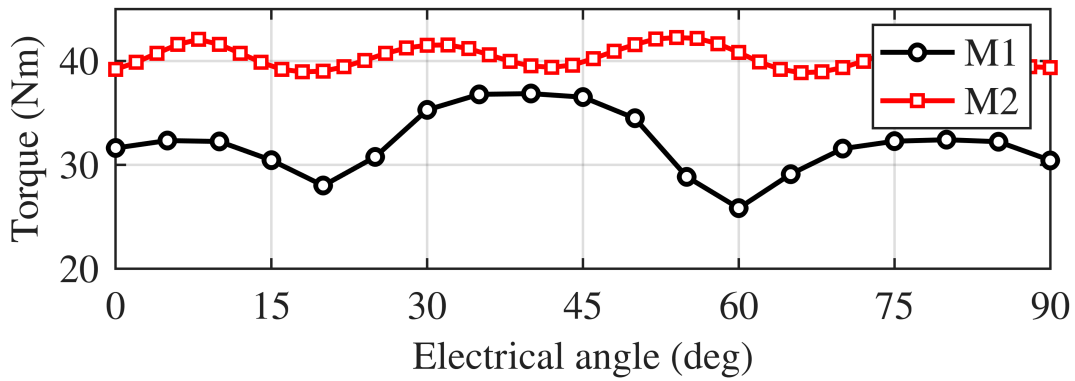


Fig. 4.11 3D FEA torque simulation for both study cases.

### 4.3.2 Experimental validation

To further validate the proposed procedure, a reverse engineering approach is adopted. Experimental tests are conducted on two motors, and the measured torque values are used as target inputs for the sizing algorithm. The resulting predicted dimensions are then compared with the actual machine dimensions.

The first prototype tested is a double-stage, 12-slot/10-pole axial flux PM motor with FSCW. The second experimental setup features a single-stage, 48-slot/8-pole motor with distributed windings. The specifications of both machines, hereafter referred to as P1 and P2, respectively, are summarized in Table 4.7.

The experimental setups shown in Fig. 4.12 and Fig. 4.13 are tested by operating both machines as generators connected to a resistive load. For P1, an induction motor was used to drive the machine and for P2, a DC motor was used as the prime mover, measuring the shaft torque for both prototypes.

Table 4.7 Prototypes specification data.

| Parameter           | P1           | P2           |
|---------------------|--------------|--------------|
| Topology            | Double-stage | Single-stage |
| Slot number         | 12           | 48           |
| Pole number         | 10           | 8            |
| Inner diameter      | 155 mm       | 170 mm       |
| Outer diameter      | 286 mm       | 282 mm       |
| Air gap thickness   | 6 mm         | 1.5 mm       |
| Winding type        | FSCW         | DW           |
| PM height           | 20 mm        | 4 mm         |
| PM material         | SmCo         | NdFeB        |
| Lamination material | M400-50A     | M235-35A     |

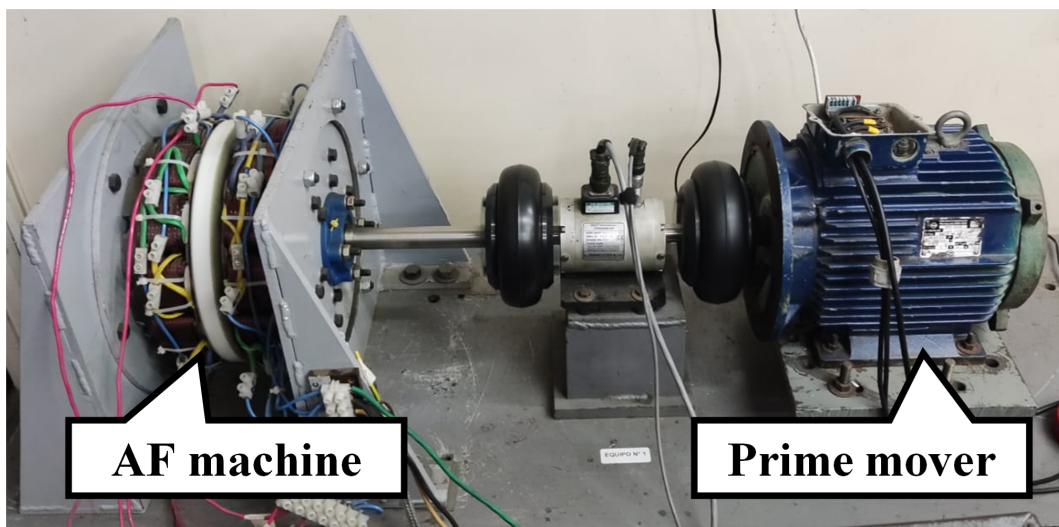


Fig. 4.12 Experimental setup for P1.

The validation process involves determining the input parameters listed in Table 4.1 through a combination of direct measurements and design specifications. For the stress loading indices, the slot current density is calculated based on the current measured during testing as well as the maximum linear current density. The maximum magnetic loading indices are obtained via FEA simulations. Subsequently, using the

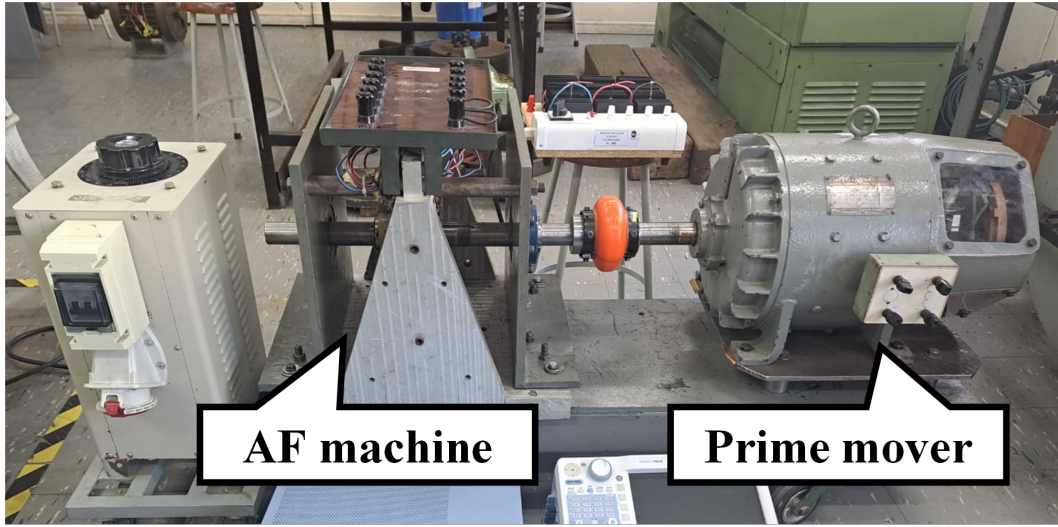


Fig. 4.13 Experimental setup for P2.

torque measurements from both prototypes, the sizing procedure is carried out, and the resulting outputs are compared to the dimensions of both machines.

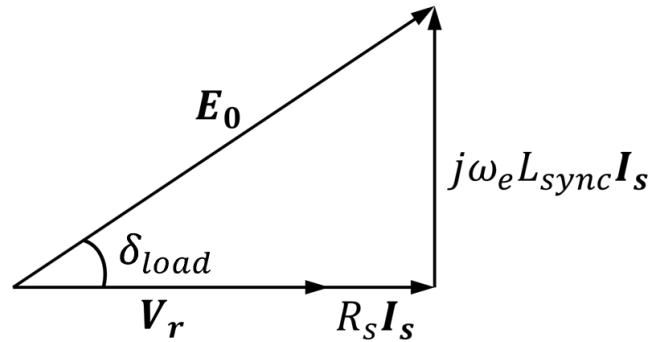


Fig. 4.14 Phasor diagram of an AFPM in generator mode connected to a resistive load.

An additional consideration before comparing the measured and expected dimensions is the need to adjust the experimental torque values. Since the axial flux machine operates as a generator, the load angle will impact on the machine's output torque. Therefore, determining the electrical circuit parameters, specifically, the phase resistance  $R_s$  and the phase synchronous inductance  $L_{sync}$ , is essential. In the phasor diagram shown in Fig. 4.14,  $\delta_{load}$  is the unknown load angle. The phase inductance was experimentally determined following the test procedure described in [121], while the phase resistance was obtained through a simple DC test. Based on the phasor diagram, the load angle was subsequently computed as shown in (4.66).

$$\delta_{load} = \arctan \left( \frac{I_s \omega_e L_{sync}}{V_r + I_s R_s} \right) \quad (4.66)$$

Table 4.8 and Table 4.9 present the predicted and measured values for prototypes P1 and P2, respectively. According to these results, the comparison shows differences within 5% between the predicted dimensions from the sizing tool and the measurements from both prototypes with the only exception of the inner tooth width which present a deviation around 12%. As mentioned in the previous section, this larger discrepancy can be primarily attributed to the assumptions regarding the iron material in the sizing procedure. In particular, the inner region of the stator tooth operates under higher saturation (see Fig. 4.15), leading to greater deviations. Additionally, although the relative error appears significant, the absolute error remains below 3 mm, which is not considered critical. These outputs further validate the effectiveness of the proposed methodology and demonstrate its suitability as a reliable early-stage design tool for axial flux permanent magnet motors.

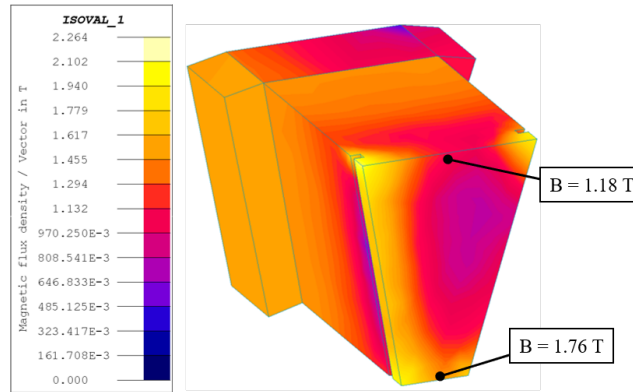


Fig. 4.15 Magnetic flux density distribution in one tooth.

## 4.4 Chapter conclusion

This chapter presented a self-consistent algorithm for the preliminary sizing of permanent magnet axial flux motors. The proposed procedure follows an iterative approach, wherein the machine dimensions are incrementally adjusted at each step to satisfy the imposed electromagnetic loading constraints. The formulations presented in this work are based on analytical formulations considering several material stresses indexes provided by the user. Owing to the simplicity and clarity of the equations,

Table 4.8 Tool results comparison with P1.

| Parameter             | Tool (mm) | Measured (mm) | $\Delta\%$ |
|-----------------------|-----------|---------------|------------|
| Stator outer diameter | 281.8     | 286           | -1.5%      |
| Stator inner diameter | 152.7     | 155           | -1.5%      |
| Slot height           | 59.8      | 60            | -0.3%      |
| Slot width            | 22.5      | 23            | -2.2%      |
| Tooth inner width     | 17.5      | 20            | 12.5%      |
| Tooth outer width     | 51.3      | 52            | -1.3%      |
| Stator yoke height    | 16.8      | 16.5          | 1.8%       |

Table 4.9 Tool results comparison with P2.

| Parameter             | Tool (mm) | Measured (mm) | $\Delta\%$ |
|-----------------------|-----------|---------------|------------|
| Stator outer diameter | 278.9     | 282           | -1.1%      |
| Stator inner diameter | 168.1     | 170           | -1.1%      |
| Slot height           | 19.5      | 19.8          | -1.5%      |
| Slot width            | 6.3       | 6.6           | -4.5%      |
| Tooth inner width     | 4.7       | 4.6           | 2.2%       |
| Tooth outer width     | 12        | 12            | 0%         |
| Stator yoke height    | 38        | 40            | -5.0%      |
| Rotor yoke height     | 29.6      | 30            | -1.3%      |

the procedure is accessible and suitable for implementation even by less-experienced designers. The proposed sizing procedure was validated through 3D finite element simulations conducted on two selected case studies. Additionally, two prototypes were tested, and experimental data were used to further assess the accuracy and reliability of the procedure through a reverse engineering method. The observed discrepancies between predicted and measured values remained within acceptable limits, proving the consistency of the analytical-based approach since the preliminary electromagnetic sizing stage.

One of the main advantages of the presented procedure is its straightforward integration with more sophisticated design and analysis tools. For instance, this

methodology can be coupled with optimization algorithms, magnetic equivalent circuit procedures, thermal analysis tools, mechanical stress analysis and finite element analysis methodologies to further refine the sizing procedure. This flexibility enables the development of more accurate and robust designs prior to the prototyping phase, thereby reducing development time and improving overall design efficiency.

## **Chapter 5**

# **Electromagnetic analysis tool for axial flux machines**

As discussed in the introduction, the design of electric machines requires a multi-physics approach, encompassing electromagnetic, thermal, and mechanical analyses to accurately evaluate performance. Electromagnetic analysis, in particular, is typically carried out using finite element analysis tools. However, FEA is associated with high computational costs and long simulation times. Consequently, analytical methods are often considered a valuable alternative, as they enable faster computations and facilitate integration with other modeling tools and optimization frameworks. Among these methods, the magnetic equivalent circuit (MEC) approach offers an effective compromise between accuracy and computational efficiency.

The MEC methodology discretizes the electric machine into distinct regions, each represented by a reluctance (or permeance) element to model its magnetic behavior. By applying principles analogous to electrical circuit theory, MEC models can effectively predict machine performance. Moreover, this approach allows the inclusion of phenomena such as magnetic leakage, rotor motion, and different winding configurations, making it a versatile tool for electromagnetic analysis.

In this context, this chapter presents a magnetic equivalent circuit model for PM axial flux machines, capable of accommodating various configurations under both no-load and load operating conditions [122]. The proposed MEC accounts for nonlinear magnetic behavior, rotor motion, and permanent magnet subdivision to improve resolution and accuracy. It is based on a quasi-3D representation, enabling





The MEC model also accounts for leakage flux paths, represented by slot leakage reluctances ( $R_{ss}$ ) and permanent magnet leakage reluctances ( $R_{r\sigma}$  and  $R_{PM\sigma}$ ). A preliminary assessment of these effects indicates an improvement in electromagnetic torque prediction accuracy of approximately 2–3% for the considered case studies. Furthermore, the inclusion of these additional reluctances has a negligible impact on computational cost, as their number remains small compared to that of the air gap reluctances.

In the proposed model, no specific reluctance element is assigned to the stator shoe region. This simplification is adopted to facilitate integration into design workflows and to reduce computational time by limiting the network complexity. The effects of saturation in the stator shoe can, if necessary, be evaluated in subsequent stages of the design process.

The proposed circuit is a mesh-based model, in which Kirchhoff's voltage law is applied to determine the unknown magnetic fluxes in each loop based on the imposed magnetomotive forces and magnetic reluctances [123]. Since the reluctances associated with ferromagnetic regions are nonlinear elements, the resulting system of equations is nonlinear and must be solved numerically to evaluate machine performance.

To solve the MEC, each reluctance in the network must be computed. The general expression for the magnetic reluctance  $R_m$  is given in (5.1), where  $l$  is the length of the magnetic flux path,  $S$  is its cross-sectional area,  $\mu_r$  is the relative magnetic permeability, and  $\mu_0$  is the permeability of free space. This expression simplifies for prismatic or cylindrical geometries, where both the cross-sectional area and the relative permeability remain constant along the path length  $l$ .

$$R_m = \int \frac{dl}{\mu_0 \mu_r(l) S(l)} \Rightarrow R_m = \frac{l}{\mu_0 \mu_r S}, \quad \text{if } \begin{cases} \frac{dS}{dl} = 0 \\ \frac{d\mu_0}{dl} = 0 \end{cases} \quad (5.1)$$

The following formulations of the reluctances are based on the dimensions reported in Fig. 4.3 for the considered diameter  $D_i$ .

### 5.1.1 Stator and rotor iron reluctance

The  $j$ -th stator tooth reluctance can be computed as in (5.2), where  $w_t$  is the tooth width at a given diameter,  $\mu_{r,st,j}$  is the relative permeability of the flux tube iron and  $L_{core}$  is the radial length of the stator core.

$$R_{st,j} = \frac{h_{ts}}{\mu_0 \mu_{r,st,j} L_{core} w_t} \quad (5.2)$$

In (5.2),  $\mu_{r,st,j}$  is function of the magnetic flux passing through the reluctance  $R_{st,j}$  as defined in (5.3).

$$\mu_{r,st,j} = f(\Phi_{s,j-1} - \Phi_{s,j}) \quad (5.3)$$

As shown in (5.3), the permeability of the stator tooth is determined by the resultant flux between adjacent loops, as illustrated in Fig. 5.1. This resultant flux corresponds to the actual flux flowing through the element, leading to iron saturation and, consequently, a relative permeability defined by the material's BH curve.

Similarly, the  $j$ -th stator yoke reluctance and the  $j$ -th rotor yoke reluctance can be calculated using (5.4) and (5.5), respectively. Here,  $N_c$  denotes the number of stator slots;  $\mu_{r,sy,j}$  and  $\mu_{r,ry,j}$  represent the relative permeability of the  $j$ -th flux tube in the stator and rotor yokes, respectively;  $h_{ys}$  and  $h_{yr}$  are the stator and rotor yoke thicknesses;  $p$  is the number of poles; and  $L_{core}$  is the radial length of the iron core.

$$R_{sy,j} = \frac{\pi D_i / N_c}{\mu_0 \cdot \mu_{r,sy,j} \cdot L_{core} \cdot h_{ys}} \quad (5.4)$$

$$R_{ry,j} = \frac{\pi D_i / (p)}{\mu_0 \cdot \mu_{r,ry,j} \cdot L_{core} \cdot h_{yr}} \quad (5.5)$$

In this case, the flux passing through the reluctances directly corresponds to the loop flux (see Fig. 5.1). Therefore:

$$\mu_{r,sy,j} = f(\Phi_{s,j}) \quad ; \quad \mu_{r,ry,j} = f(\Phi_{rot,j}) \quad (5.6)$$

### 5.1.2 Slot leakage reluctance

To account for the effects of slot leakage flux, several formulations are available in the literature, many of which have demonstrated good agreement with experimental results. In this work, the permeance expressions proposed in [89] and [90] are adopted to compute the slot leakage reluctance for distributed windings and fractional-slot concentrated windings, respectively. Accordingly, the following expression is used to evaluate the slot permeance for both winding configurations.

$$\begin{cases} \lambda_{\sigma s} = \frac{h_{us}}{w_{t2}} \left[ \frac{\left(\frac{w_{t1}}{w_{t2}}\right)^2 - \left(\frac{w_{t1}}{4w_{t2}}\right)^4 - \ln\left(\frac{w_{t1}}{w_{t2}}\right) - \frac{3}{4}}{\left(1 - \frac{w_{t1}}{w_{t2}}\right)\left(1 - \left(\frac{w_{t1}}{w_{t2}}\right)^2\right)^2} \right], & \text{for } DW \\ \lambda_{\sigma s} = k_{1h} \left(\frac{h_{us} - h_{ly}}{3w_s}\right) + k_{2h}K + \frac{h_{ly}}{4w_{so}}, & \text{for } FSCW \end{cases} \quad (5.7)$$

In (5.7),  $w_{t1}$  and  $w_{t2}$  are the inner and outer tooth width. While  $h_{ly}$  is the distance between two winding layers,  $k_{1h}$  and  $k_{2h}$  are the coefficients that consider the number of winding layers, calculated as in 4.45, while  $K$  is defined as follows:

$$K = \frac{h_{ns}}{w_s} + \frac{h_{es}}{w_{so}} + \left(\frac{h_{ls}}{w_s - w_{so}}\right) \ln \frac{w_s}{w_{so}} \quad (5.8)$$

### 5.1.3 Air gap reluctance

The relative motion between the rotor and stator teeth results in a continuous variation of the MEC topology at each operating point. To accurately capture this behavior within the reluctance network, an appropriate air gap reluctance model must be adopted. In this study, four distinct operating scenarios are identified, as illustrated in Fig. 5.2.

The number of sections into which the air gap is subdivided is a user-defined parameter that directly influences both the model accuracy and the computational effort. As shown in Fig. 5.2, Case A corresponds to the condition in which the air gap section fully faces both the stator tooth and the permanent magnet. Case B represents the situation where the air gap section faces the permanent magnet and only a portion of the stator tooth, resulting in an increased reluctance. In Case C, the air gap section faces the permanent magnet and the stator slot entirely. Finally, Case

D corresponds to the condition in which the air gap section does not include any portion of the permanent magnet, making the associated magnetic flux negligible.

The detailed formulations for the first three cases are provided in (5.9), (5.10), and (5.11).

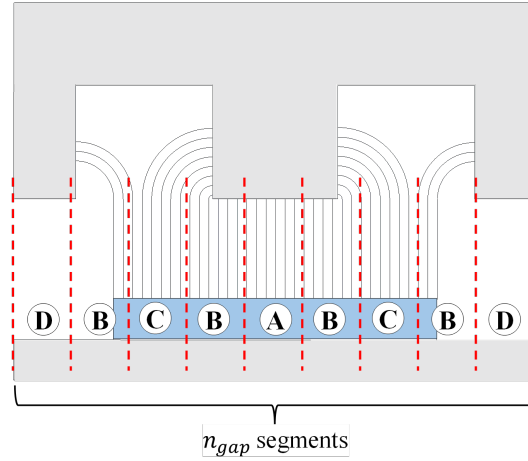


Fig. 5.2 Air gap subdivision principle.

$$R_{gap,j} = \begin{cases} R_{gap,min} \\ k_{gap}R_{gap,min} + (1 - k_{gap})R_{fringe} \\ R_{fringe} \end{cases} \quad (5.9)$$

$$R_{gap,min} = \frac{h_{gap}}{\mu_0 S_{gap}} \quad (5.10)$$

$$R_{fringe} = \frac{1}{\mu_0 S_{gap}} \left( h_{gap} + \frac{\pi}{4} l_{fringe} \right) \quad (5.11)$$

where  $R_{gap,min}$  is the minimum air gap reluctance,  $k_{gap}$  defines the portion of an air gap section facing a stator tooth,  $h_{gap}$  is the air gap thickness, and  $S_{gap}$  is the area of the air gap section. The term  $R_{fringe}$  represents the fringe flux reluctance, modeled as an additional flux path with a length equal to a quarter of a circumference and a radius  $l_{fringe}$ , defined as the distance between a tooth edge and the air gap section under consideration.

### 5.1.4 Permanent magnet reluctance

Each permanent magnet segment is modeled as a magnetomotive force source connected in series with a magnetic reluctance. The amplitude of this MMF can be calculated as in (5.12), where  $B_r$  is the remanent flux density at the operating temperature,  $h_{PM}$  is the magnet axial thickness, and  $\mu_{r,PM}$  is the relative permeability of the permanent magnet material.

$$F_{PM} = \frac{B_r h_{PM}}{\mu_0 \mu_{r,PM}} \quad (5.12)$$

It is worth noting that the sign of the PM MMF, is given by the polarity of the magnet and should be assign accordingly in the MEC. The reluctance of each PM segment can be computed as:

$$R_{PM} = \frac{h_{PM}}{\mu_0 \cdot \mu_{r,PM} \cdot \tau_p \alpha_{PM} L_{PM}} \quad (5.13)$$

where,  $L_{PM}$  is the magnet length in the radial direction.

### 5.1.5 Rotor leakage reluctances

The developed MEC also accounts for flux leakage occurring in the regions between adjacent permanent magnets. In particular, this includes leakage flux passing through the air gap between two adjacent PMs, as well as flux circulating between a PM and the rotor yoke without linking to the stator iron, as described in [124]. Fig. 5.3 illustrates these situations.

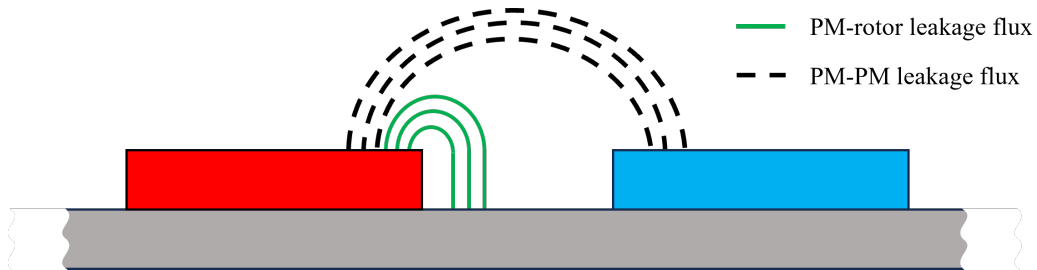


Fig. 5.3 Rotor leak fluxes considered in the MEC.

To model these effects, two types of reluctances are introduced in the MEC between the permanent magnets. The formulations for these reluctances are derived from [124]. Although some assumptions related to no iron saturation and negligible magnetic field contribution from the armature windings are considered, the authors in [124] achieve good agreement with finite element simulation results.. These reluctances can be computed as:

$$R_{r\sigma} = \left[ \mu_0 \frac{L_{core}}{\pi} \ln \left( 1 + \frac{\pi h'_{gap}}{h_{PM}} \right) \right]^{-1} \quad (5.14)$$

$$R_{PM\sigma} = \left[ \mu_0 \frac{L_{core}}{\pi} \ln \left( 1 + \frac{\pi h'_{gap}}{\tau_p(1 - \alpha_{PM})} \right) \right]^{-1} \quad (5.15)$$

where,  $R_{r\sigma}$  is the PM-rotor leakage reluctance,  $R_{PM\sigma}$  is the PM-PM leakage reluctance, and  $h'_{gap}$  is the effective air gap that considers the slotting effect.

### 5.1.6 Multi-stage configuration

The MEC shown in Fig. 5.1 is developed for a single-stage configuration. However, the circuit topology can be readily adapted to the various multi-stage machine configurations presented in Fig. 3.2. This adaptation requires only minor modifications to the baseline model.

Figure 5.4a illustrates the modified circuit topology for the 2S1R-NN configuration, in which the rotor yoke reluctances are removed, as the PM-to-rotor leakage paths can be neglected under the assumption that the PM holder is made of non-ferromagnetic material. Figure 5.4b shows the circuit for the 1S2R-NS configuration, where the stator yoke reluctances are omitted compared to the baseline model.

For the 2S1R-NS and 1S2R-NN configurations, the performance can be directly obtained by doubling the electromagnetic torque values of the baseline MEC.

## 5.2 The system of non-linear equations

To solve the proposed MEC, Kirchhoff's voltage law is applied to each loop, resulting in the system of nonlinear equations expressed in matrix form in (5.16).

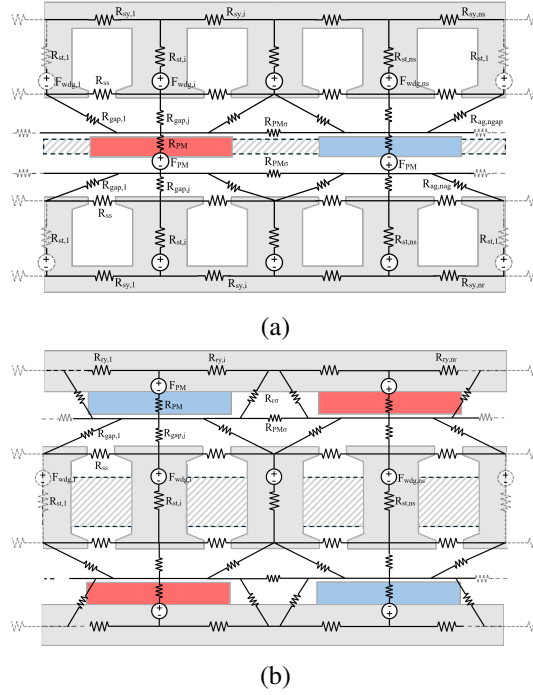


Fig. 5.4 Magnetic equivalent circuit for: (a) 2S1R-NN and (b) 1S2R-NS.

$$\mathbf{A_r} \cdot \Phi = \mathbf{F} \quad (5.16)$$

Here,  $\mathbf{A_r}$  represents the reluctance matrix associated with the network,  $\Phi$  is the vector of the loop magnetic fluxes, and  $\mathbf{F}$  denotes the vector of magnetomotive force sources.

The matrix  $\mathbf{A_r}$  is constructed from the previously computed reluctances and arranged according to the network topology. Its partitioning into sub-matrices is shown in (5.17), where the diagonal matrices  $\mathbf{A_s}$ ,  $\mathbf{A_{gap}}$ ,  $\mathbf{A_{PM}}$ , and  $\mathbf{A_{rot}}$  correspond to the reluctances of the stator, air gap, permanent magnets (including leakage paths), and rotor, respectively. The off-diagonal terms represent the interactions between these machine components.

$$\mathbf{A_r} = \begin{bmatrix} \mathbf{A_s} & \mathbf{A_{s,gap}} & \mathbf{A_{s,PM}} & \mathbf{0} \\ \mathbf{A'_{s,gap}} & \mathbf{A_{gap}} & \mathbf{A_{gap,PM}} & \mathbf{A_{gap,rot}} \\ \mathbf{A'_{s,PM}} & \mathbf{A'_{gap,PM}} & \mathbf{A_{PM}} & \mathbf{A_{PM,rot}} \\ \mathbf{0} & \mathbf{A'_{gap,rot}} & \mathbf{A'_{PM,rot}} & \mathbf{A_{rot}} \end{bmatrix} \quad (5.17)$$



The matrix is assembled according to circuit theory principles for mesh-based network analysis. In particular, each diagonal element at position  $(i, i)$  represents the sum of all reluctances associated with the  $i$ -th loop. Conversely, each off-diagonal element at position  $(i, j)$  is equal to the negative of the reluctances shared between the  $i$ -th and  $j$ -th loops.

Special attention must be given to off-diagonal elements associated with boundary reluctances, for which symmetry conditions must be enforced. In particular, if the machine periodicity is odd, the sign of the boundary reluctance must be reversed.

The vector  $\mathbf{F}$  in (5.16) represents the magnetomotive force associated with each magnetic flux loop. In the proposed MEC, two sources of MMF are considered: the stator windings and the permanent magnets. The MMF contribution of the permanent magnets is calculated using (5.12) and assigned to the corresponding loops. The MMF generated by the windings is instead determined using (5.18), which accounts for the total magnetomotive force produced by the stator currents.

$$\mathbf{F}_{wdg} = \mathbf{i}_{abc} \cdot \mathbf{N}_{wdg} \quad (5.18)$$

In (5.18),  $\mathbf{F}_{wdg}$  denotes the MMF vector associated with the windings,  $\mathbf{i}_{abc}$  is the vector of instantaneous phase currents, and  $\mathbf{N}_{wdg}$  is the winding matrix that describes the distribution of turns within the slots.

The stator winding matrix  $\mathbf{N}_{wdg}$  is constructed based on the specific stator winding configuration, such as distributed or concentrated, and single or double-layer arrangements. It ultimately depends on the number of slots per pole and per phase  $q_s$ , as well as the number of turns in series per phase per slot  $Z_c$ . An example is provided in (5.19) for a 10-pole, 12-slot axial flux machine with single-layer FSCW.

$$\mathbf{N}_{wdg} = \begin{bmatrix} Z_c & -Z_c & 0 & 0 & 0 & 0 & -Z_c & Z_c & 0 & 0 & 0 & 0 \\ 0 & 0 & -Z_c & Z_c & 0 & 0 & 0 & 0 & Z_c & -Z_c & 0 & 0 \\ 0 & 0 & 0 & 0 & Z_c & -Z_c & 0 & 0 & 0 & 0 & -Z_c & Z_c \end{bmatrix} \quad (5.19)$$

Finally, the vector  $\Phi$  contains the magnetic fluxes associated with each loop and represents the unknown of the system. It is important to note that these loop fluxes do not always correspond directly to the physical fluxes in the machine. For instance, the reluctances in the stator yoke are directly traversed by the loop fluxes, whereas, in the stator teeth, the actual flux is given by the difference between the fluxes of adjacent loops, as expressed in (5.3).

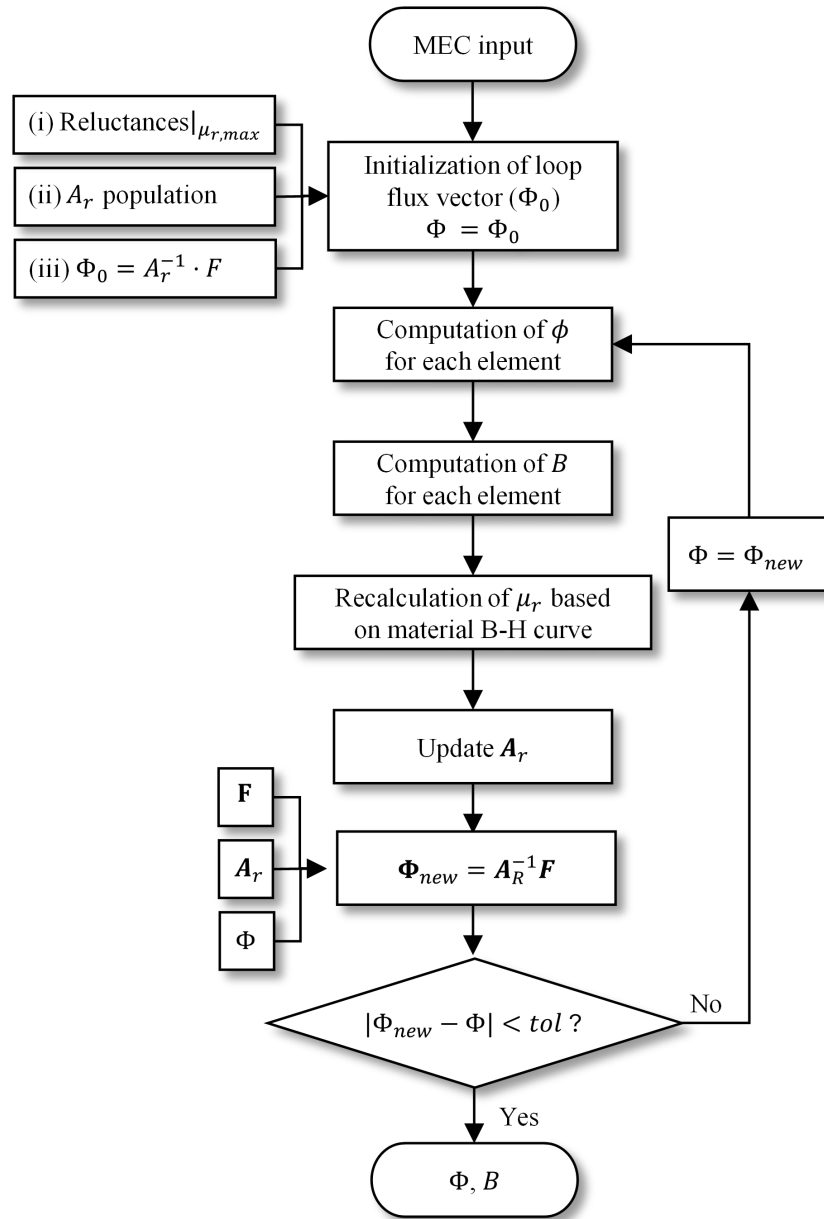


Fig. 5.5 Procedure workflow.

### 5.2.1 Numerical solution procedure

The MEC solution is obtained numerically using an iterative method suited for large, sparse, and symmetric systems of the form  $A \cdot x = b$ . The complete solution procedure is illustrated in Fig. 5.5.

The process begins with an initial estimate of the loop flux vector  $\Phi_0$ . This initial guess is obtained by: (i) computing the network reluctances under unsaturated conditions (i.e., assuming maximum magnetic permeability), (ii) assembling the reluctance matrix  $A_r$ , and (iii) solving (5.16) using its inverse.

The loop flux vector is then used to determine the flux in each circuit branch and, consequently, the magnetic flux density associated with each discretized element of the network. These magnetic flux density values are used to compute the corresponding magnetic permeabilities based on the BH curve of the material. With the updated permeability values, the reluctances are recalculated and the reluctance matrix  $A_r$  is consequently updated. This procedure is therefore applied iteratively until a convergence value defined beforehand is achieved.

### 5.2.2 Mechanical rotation

The proposed MEC accounts for mechanical rotation by updating the reluctance network at each rotor position, requiring the procedure outlined in Fig. 5.5 to be repeated for every considered rotor step. To enable this, the machine is modeled as an infinitely extended stator by replicating the slot pattern multiple times, thereby emulating continuous rotation. Appropriate boundary conditions, such as periodicity, are applied to accurately represent the magnetic field behavior within the domain.

At each position, the procedure determines the connectivity between stator teeth and air gap sections based on their relative alignment. Using this information, the mesh loops are redefined and the reluctance matrix is reconstructed accordingly.

This approach allows the algorithm to handle arbitrary slot/pole combinations, winding arrangements, and slot geometries, as it relies on both the relative positioning between stator teeth and permanent magnets and the periodicity of the machine, and it is not tailored for some specific topologies. Furthermore, it enables the evaluation of the extent to which each air gap section faces a stator tooth, which is particularly important for accurately modeling fringe flux effects since the approach presented in Fig. 5.2 is applied at each mechanical step.

Finally, a sufficiently fine discretization is recommended to accurately capture the magnetic interactions between individual magnet segments and the stator iron.

## 5.3 Electromagnetic performance computation

This section provides the formulations used to estimate the performance of the AFM predicted by the magnetic equivalent circuit developed in this chapter. For the no-load operation the MMF produced by the stator currents are set to zero, while for load condition, the dq-axis currents are provided as inputs.

### 5.3.1 Magnetic flux density

Once the system of equations from the MEC has been solved for each considered mechanical position, the magnetic flux density on each element  $B_{m,i}$  can be obtained by dividing the corresponding flux by the surface area of the flux tube, as shown below:

$$B_{m,i} = \frac{\phi_{m,i}}{A_{flux}} \quad (5.20)$$

where  $\phi_{m,i}$  is the magnetic flux flowing through a given MEC element, obtained by solving (5.16), and  $A_{flux}$  is the corresponding cross-sectional area, which depends on the geometry of the flux tube under consideration. By collecting the results at each mechanical step, the magnetic flux and flux density over one period can be reconstructed, allowing the computation of peak and RMS values.

### 5.3.2 No-load air gap flux density and voltage

The MEC can be used to simulate the machine under no-load conditions. For a given rotor position, the air gap flux density can be obtained by dividing the flux in each air gap element by the corresponding flux tube area. The fundamental component of the no-load magnetic flux density can then be used to compute the fundamental component of the no-load voltage  $E_1$  as follows:

$$E_1 = \frac{2\pi}{\sqrt{2}} f N_{turns} k_{w,1} (\alpha_{PM} B_{g1} \tau_p L_{PM}) \quad (5.21)$$

Where  $N_{turns}$  is the number of turns in series per phase, and  $k_{w,1}$  is the fundamental winding factor and  $B_{g1}$  is the peak value of the fundamental component of the air gap flux density.

### 5.3.3 Electromagnetic torque

The analytical computation of the electromagnetic torque is based on the dq-axes formulation given in (5.22):

$$T_{em} = \frac{3}{2} pp (\lambda_d i_q - \lambda_q i_d) \quad (5.22)$$

where  $T_{em}$  is the electromagnetic torque,  $\lambda_d$  and  $\lambda_q$  are the flux linkages along the d- and q-axes, respectively, and  $i_d$  and  $i_q$  are the corresponding d- and q-axis currents, obtained from the stator phase currents through reference frame transformations. The flux linkages in the windings can be calculated using (5.23) [125]:

$$\lambda_{abc} = pp \cdot \mathbf{N}_{wdg} \cdot \Phi_{sy} \quad (5.23)$$

Here,  $\lambda_{abc}$  is the linked flux in the  $abc$  reference frame and  $\Phi_{sy}$  is the stator yoke flux vector.

## 5.4 Validation of the methodology

In order to validate the proposed reluctance network, a comparison with both finite element simulations and experimental tests is carried out.

### 5.4.1 FEA validation

For the FEA validation, two single-stage axial flux motor models with different pole/slot combinations and periodicities are considered. Figure 5.6 presents the 3D geometries of the case studies (M1 and M2, respectively), while Table 5.1 summarizes their main dimensions.

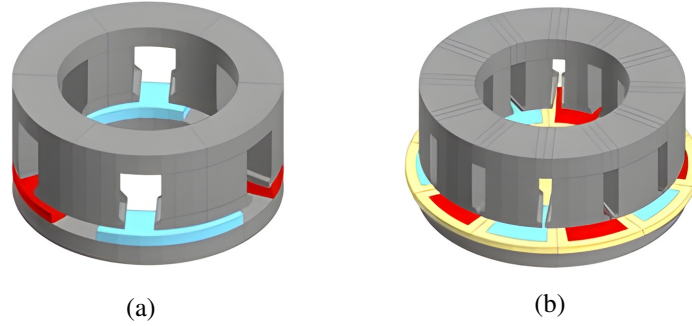


Fig. 5.6 3D sketches of the study cases: (a) M1, (b) M2.

Table 5.1 Main specifications for the study cases.

| Parameter             | M1     | M2     |
|-----------------------|--------|--------|
| Slot number           | 6      | 12     |
| Pole number           | 4      | 10     |
| Stator Inner diameter | 120 mm | 155 mm |
| Stator Outer diameter | 180 mm | 286 mm |
| Air gap thickness     | 1 mm   | 6 mm   |
| PM remanence          | 0.80 T | 1.05 T |

The results are evaluated in terms of no-load magnetic flux density in the air gap, back electromotive force, and electromagnetic torque under load conditions. For the MEC analysis, a fine discretization of the air gap is adopted, with at least 50 elements per pole.

Figure 5.7 shows the air gap magnetic flux density obtained using both the proposed MEC and 3D FEA simulations. The largest discrepancies are observed in regions where slotting effects are more pronounced, with differences reaching up to 10%. These deviations are primarily attributed to the neglect of tooth-tip geometry in the MEC and to the analytical evaluation of fringing flux, which relies on assumptions regarding the length of the flux lines entering the sides of the teeth (see Fig. 5.2, Cases B and C).

The fundamental component of the air gap magnetic flux density obtained from the MEC is then used to compute the fundamental component of the back electromotive force, as given in (5.21). A comparison with the fundamental component of the

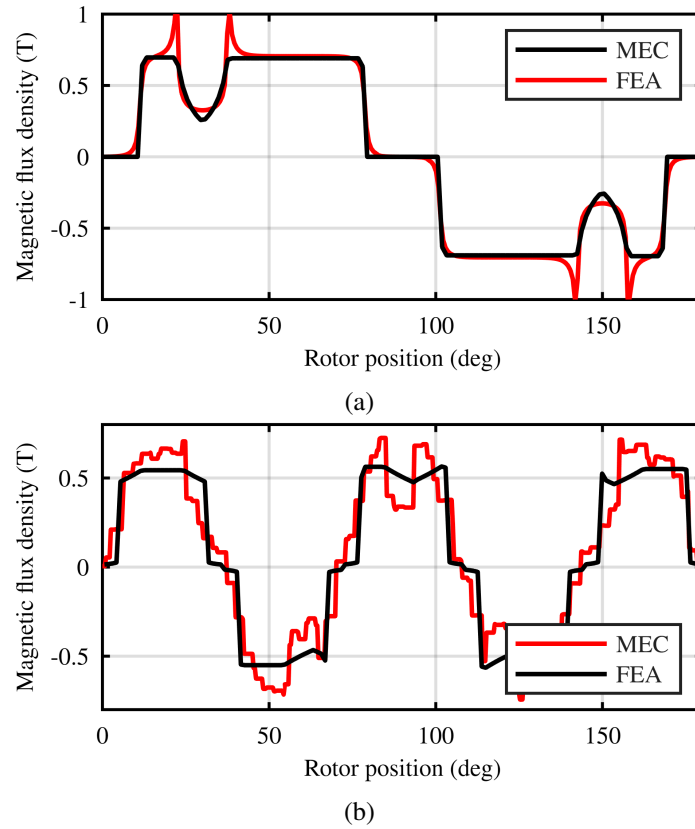


Fig. 5.7 Air gap flux density comparison between MEC and FEA for: (a) M1, and (b) M2.

Table 5.2 No-load and load results comparison for the analyzed study cases.

|          | M1      |         |             | M2      |         |             |
|----------|---------|---------|-------------|---------|---------|-------------|
|          | MEC     | FEA     | $\Delta \%$ | MEC     | FEA     | $\Delta \%$ |
| $B_g$    | 0.65 T  | 0.66 T  | 1.5%        | 0.61    | 0.61    | 0.1%        |
| $E_1$    | 129.3 V | 130.9 V | 1.2%        | 107.8 V | 107.8   | 0.1%        |
| $T_{em}$ | 24.7 Nm | 25.0 Nm | 1.1%        | 40.6 Nm | 40.3 Nm | 1%          |

back electromotive force obtained from FEA is reported in Table 5.2, showing an average error of less than 1.5% for both cases.

Figure 5.8 compares the MEC and FEA models under load conditions in terms of electromagnetic torque for both motors. While the torque ripple predicted by the MEC differs from that obtained with the 3D FEA model, primarily due to the limited number of air gap reluctance elements considered in the network, the average torque values show good agreement between the two approaches.

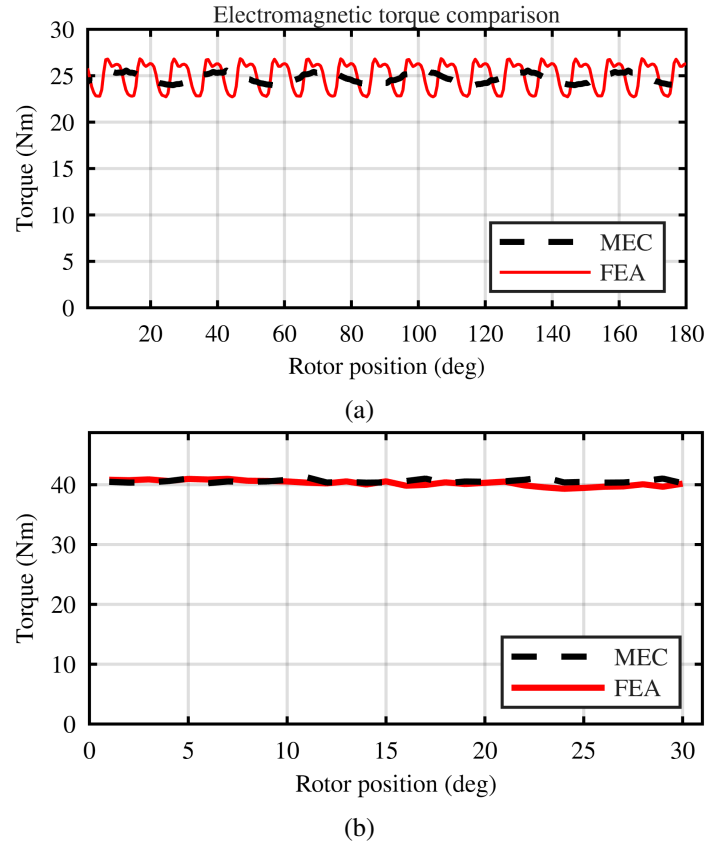


Fig. 5.8 Electromagnetic torque comparison between MEC and FEA for: (a) M1, and (b) M2.

This is confirmed by the results reported in Table 5.2, which indicate that the maximum error in the average electromagnetic torque for the two case studies is on the order of 1%. These findings demonstrate that the developed MEC is suitable for integration into fast design and optimization procedures. Accordingly, the model accuracy is considered satisfactory when the mean torque and the fundamental components of the electromotive force and air gap magnetic flux density are in good agreement with the validation results.

One limitation of the proposed MEC model lies in its reduced capability to accurately capture secondary effects such as torque ripple and air gap flux density harmonics. Although these phenomena are typically analyzed using FEA within the design workflow of electrical machines, they could be partially addressed within the MEC framework by increasing the number of air gap reluctance elements per pole,



including the reluctances of the tooth tips and increasing fringe flux computation accuracy.

An additional comparison under high-saturation conditions was carried out for M2 to assess the performance of the model under overload operation. In this case, an operating point exceeding twice the rated current was considered. The results show a deviation of approximately 2% of the torque, indicating that the MEC maintains a high level of accuracy even under such conditions.

It is also worth highlighting that the MEC-based results are obtained with a significantly reduced computation time compared to FEA. For the M2 case study, the computation time decreases from 5270 seconds for the 3D FEA model to 13 seconds for the MEC, making the proposed approach approximately 400 times faster. Both simulations were performed on the same workstation, equipped with 128 GB of RAM and 20 CPU cores.

## 5.4.2 Experimental validation

To further validate the proposed reluctance network, the MEC results are compared with experimental data obtained from an AFPM prototype. The specifications of the tested machine correspond to those of M2 reported in Table 5.1.

Figure 5.9 illustrates the experimental setup, in which the prototype operates as a generator connected to a resistive load. The machine is driven by a prime mover and equipped with a torque meter.

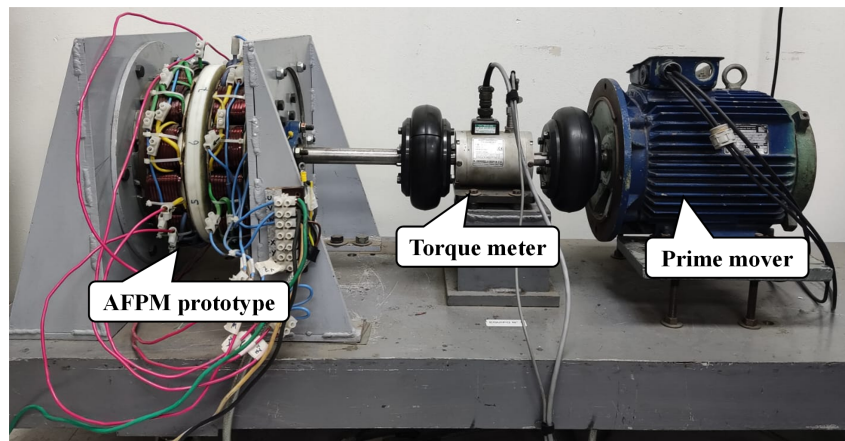


Fig. 5.9 Experimental set-up for validation.

To replicate the test conditions in the MEC, the load angle  $\delta_{load}$ , shown in the phasor diagram in Fig. 5.10, must be determined. This requires the evaluation of the machine circuit parameters, namely the phase resistance  $R_s$  and the synchronous inductance  $L_{sync}$ .

In the phasor diagram,  $\overline{V_s}$  represents the terminal voltage,  $\overline{E}$  is the open-circuit electromotive force aligned with the q-axis,  $\overline{I_s}$  is the phase current, and  $\delta$  is the load angle to be determined. Since the machine is connected to a resistive load, the power factor is unity. Additionally, as the permanent magnets are surface-mounted, the saliency is negligible, and the inductances along the d- and q-axes can be considered equal. The phase inductance is obtained as described in Section 4.3.2, following the test procedure presented in [121], while the phase resistance is measured through a DC test.

Based on the phasor diagram, the load angle is then computed using (4.66). The resulting parameters and the corresponding load angle are reported in Table 5.3. Finally, it is worth noting that the developed MEC follows the motor convention; therefore, the computed angle must be shifted by  $180^\circ$  to obtain a positive torque value.

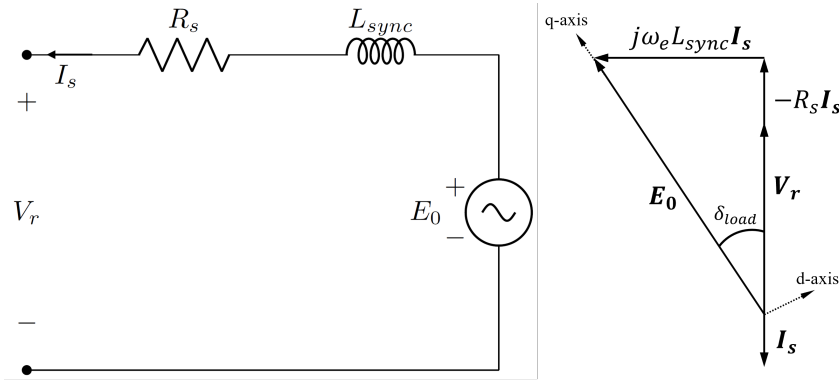


Fig. 5.10 Phasor diagram of the AFM working as generator on a resistive load.

Table 5.3 Winding parameters and load angle.

|                                  |                   |
|----------------------------------|-------------------|
| Phase resistance ( $R_s$ )       | 0.18 ( $\Omega$ ) |
| Synchronous inductance ( $L_s$ ) | 13.60 (mH)        |
| Load angle ( $\delta$ )          | 11.17 (deg)       |

With regard to the materials used in the MEC model, the BH curve provided in the lamination datasheet has been implemented for both the stator and rotor cores. The remanent flux density of the permanent magnets has been determined through calibration based on no-load voltage measurements. It is worth noting that the permanent magnets are made of SmCo. Owing to their high thermal stability, their remanent flux density can be assumed constant between the no-load condition and the evaluated operating point. However, for permanent magnet materials with higher temperature sensitivity, the remanent flux density should be adjusted according to the operating temperature.. Figure 5.11 shows the comparison between the measured and calibrated MEC no-load voltage.

The resulting calibrated model is then used to predict the electromagnetic torque corresponding to the load angle obtained from the experimental test. Figure 5.12 presents the comparison between measured and MEC-predicted torque, showing minor discrepancies, with an average error below 3%.

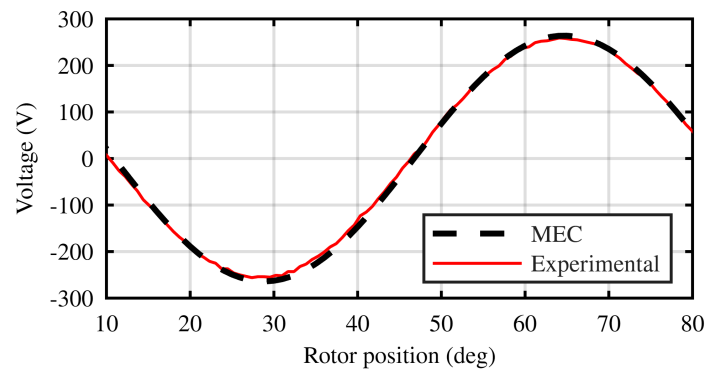


Fig. 5.11 No-load voltage comparison.

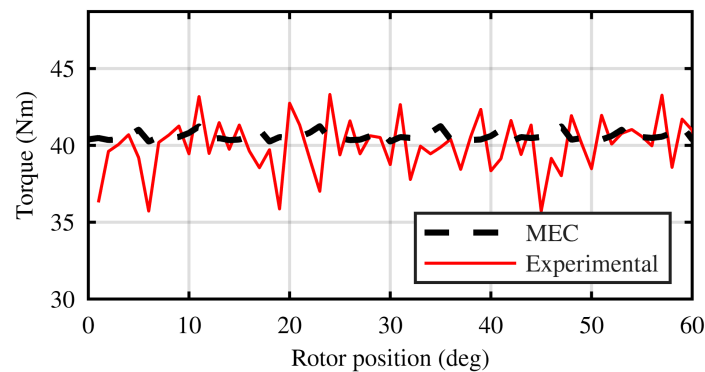


Fig. 5.12 Torque comparison.

The discrepancies observed in Fig. 5.12 can be primarily attributed to two factors. First, the MEC does not capture the torque ripple with the same level of accuracy as FEA models, owing to the lower geometric discretization adopted in the reluctance network. Second, the experimental measurements are affected by noise introduced by both the torque sensor and the data acquisition system.

## 5.5 Chapter conclusion

This chapter has presented a magnetic equivalent circuit for the analysis of axial flux permanent magnet synchronous machines. The proposed methodology effectively incorporates nonlinear iron behavior, rotor motion, and leakage effects, enabling accurate prediction of key performance quantities such as magnetic flux density, back electromotive force, and electromagnetic torque. Furthermore, the approach can be adapted to a wide range of axial flux machine configurations, from single-stage topologies to double-stage arrangements with multiple stators or rotors.

Given its accuracy, computational efficiency, and adaptability, the proposed MEC represents a valuable tool for the rapid and preliminary design of axial flux machines. Owing to its analytical formulation, the model can be readily integrated into optimization algorithms, sizing procedures, and multiphysics frameworks, thereby enabling a more automated and efficient design process.

## Chapter 6

# Lumped parameter thermal network development for axial flux machines

Determining the temperature distribution within an electric machine is of critical importance, as it defines the operational limits and directly affects material properties. In permanent magnet machines, the most thermally critical components are the winding insulation and the permanent magnets.

With regard to the windings, the insulation system determines the maximum allowable operating temperature according to its thermal class, as defined by international standards such as IEC 60034. In addition, the electrical resistance of the windings is temperature-dependent. In particular, the variation of the winding resistance  $R_s$  with temperature is defined by (6.1).

$$R_s = R_{s,0}[1 + \alpha_{cu}(T_{op} - T_0)] \quad (6.1)$$

where  $R_{s,0}$  is a known resistance value at a reference temperature,  $\alpha_{cu}$  is the temperature coefficient of resistivity of the conductor material,  $T_{op}$  is the operating temperature, and  $T_0$  is the reference temperature.

With regard to permanent magnets, these materials are highly sensitive to temperature variations, which may lead to partial or irreversible demagnetization, directly affecting machine performance. Similar to winding resistance, the remanent flux density  $B_r$  and coercive field  $H_c$  of the magnets vary with temperature, as expressed below.

$$B_r = B_{r0}[1 + \alpha_B(T_{op} - T_0)] \quad (6.2)$$

$$H_c = H_{c0}[1 + \beta_H(T_{op} - T_0)] \quad (6.3)$$

where  $B_{r0}$ ,  $H_{c0}$  are the known remanence and the PM coercivity at the reference temperature respectively.  $\alpha_B$  and  $\beta_H$  are temperature coefficients that define the variation of remanence and coercivity with temperature, respectively. For reference, Table 6.1 reports the typical values for thermal properties that influence the magnetic performance of the most common permanent magnet materials used in electric machines.

Table 6.1 Properties of PM materials.

| Material | Max. temperature* | $\alpha_B$ | $\beta_H$ |
|----------|-------------------|------------|-----------|
| Ferrite  | 200 °C            | -0.20      | 0.27      |
| SmCo     | 350 °C            | -0.03      | -0.20     |
| NdFeB    | 100 °C            | -0.10      | 0.40      |

\* Depends on the PM grade.

From Table 6.1, it can be observed that ferrite and NdFeB materials are particularly sensitive to operating temperature, making them less suitable for applications where overload conditions may occur frequently. In contrast, SmCo magnets exhibit more stable properties with respect to temperature, making them more suitable for high specific power applications and scenarios where fault tolerance is critical.

## 6.1 Thermal model of axial flux PM machine

This section presents the lumped parameter thermal network proposed for permanent magnet axial flux machines. As discussed in Chapter 3, in order to accelerate the machine design process and facilitate the integration of thermal analysis with other design stages, a fast and simplified methodology is highly desirable.

The model is formulated based on the analogy with electrical circuit theory, where thermal resistances, thermal capacitances, and heat sources correspond to electrical resistances, capacitances, and current or voltage sources, respectively. Moreover, due to the inherent geometry of axial flux machines, an accurate representation of heat transfer paths requires a modeling approach with network elements in multiple directions.

To construct the thermal network, the following simplifying assumptions are introduced:

- the heat flow in radial and axial directions is independent from each other;
- heat generation within each machine component is considered to be concentrated;
- circumferential heat flow is not considered;
- radiation is not taken into account.

Under these assumptions, the dominant heat transfer mechanisms for the axial flux machine are conduction and convection.

The proposed LPTN is shown in Fig. 6.1, where axial conduction thermal resistances are indicated in blue, radial conduction resistances in red, and convection paths in yellow. The heat sources considered in the model include iron and permanent magnet losses, evaluated through finite element analysis, and Joule losses in the windings, calculated analytically based on resistance and operating current values. Bearing and windage losses are not explicitly included; however, their contribution can be incorporated by assigning the corresponding heat sources to the appropriate nodes of the thermal network.

### **6.1.1 Conductive thermal resistances**

Based on the reference machine dimensions shown in Fig. 6.2, the conductive thermal resistances are determined by subdividing the machine into its primary components involved in the heat transfer process.





$$R_{sy} = \frac{h_{ys}}{\frac{\pi(D_2^2 - D_1^2)}{4} k_{lam}} \quad (6.4)$$

$$R_{st} = \frac{h_{ts}}{N_c \frac{(w_{t1} + w_{t2})}{2} L_{core} k_{lam}} \quad (6.5)$$

where,  $h_{ys}$  is the stator yoke thickness,  $D_1$  and  $D_2$  are the inner and outer diameter of the lamination,  $w_{t1}$  and  $w_{t2}$  are the inner and outer tooth width,  $L_{core}$  is the lamination stack length,  $N_c$  is the number of slots and  $k_{lam}$  is the lamination thermal conductivity.

### Stator winding thermal model

The stator winding is modeled as a composite structure consisting of multiple materials. Although an equivalent thermal conductivity can be derived analytically, factors such as conductor arrangement and the condition of the winding and insulation significantly influence its effective value. As an alternative, the equivalent thermal resistance between the winding and the stator iron, and thus the effective thermal conductivity of the winding bulk, can be determined experimentally through short-time thermal transient tests, as described in [126] and [127]. These studies report dedicated experimental campaigns to evaluate the equivalent thermal conductivity of the winding insulation system, denoted as  $k_{ins}$ .

In this work, empirical values from these studies are adopted for the thermal conductivity of the winding insulation. Accordingly, the winding thermal resistance is computed as given in (6.6), where  $l_{ins}$  represents the equivalent insulation thickness and  $l_{slot}$  is the slot perimeter as depicted in Fig. 6.3.

$$R_{th,wdg} = \frac{l_{ins}}{N_c l_{slot} L_{core} k_{ins}} \quad (6.6)$$

The equivalent insulation thickness can be computed from the slot dimensions and the slot fill factor  $k_{cu}$  as given by (6.7).

$$l_{ins} = \frac{w_s h_{us} (1 - k_{cu})}{l_{slot}} \quad (6.7)$$

In (6.7),  $h_{us}$  is the slot height.

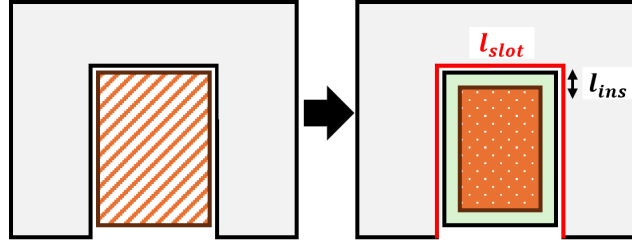


Fig. 6.3 Representation of the equivalent insulation thickness [126].

### Rotor thermal model

Similarly to the stator yoke, the rotor iron can be modeled as a hollow cylinder with a thickness  $h_{yr}$ . The corresponding resistance can thus be defined as in (6.8).

$$R_{th,ry} = \frac{h_{yr}}{\frac{\pi(D_{PM2}^2 - D_{PM1}^2)}{4} k_{lam}} \quad (6.8)$$

For the permanent magnet, the proposed model includes an axial thermal resistance  $R_{th,PM}$ , whose formulation is:

$$R_{th,PM} = \frac{h_{PM}}{\frac{\pi(D_{PM2}^2 - D_{PM1}^2)}{4} \alpha_{PM} k_{PM}} \quad (6.9)$$

In (6.9),  $\alpha_{PM}$  is the PM to pole ratio,  $h_{PM}$  is the magnet thickness,  $D_{PM1}$  and  $D_{PM2}$  are the inner and outer PM diameter respectively and  $k_{PM}$  is the thermal conductivity of the magnet material.

Finally, the rotor holder is modeled as a radial thermal resistance extending from the average diameter  $D_{ave}$  to the shaft diameter  $D_{sh}$  with a thermal conductivity  $k_{hold}$ . The corresponding thermal resistance is given by (6.10).

$$R_{th,hold} = \frac{\ln\left(\frac{D_{ave}}{D_{sh}}\right)}{2\pi h_{hold} k_{hold}} \quad (6.10)$$

### Shaft thermal model

In the proposed thermal network, the thermal conduction resistances of the endplates and housing are neglected due to their limited impact on the overall temperature

distribution. In contrast, the axial thermal resistance of the shaft,  $R_{th,sh}$ , is included to account for conduction heat transfer from the rotor to the ambient through the housing. The formulation for this resistance is presented in (6.11), where  $h_{sh}$  denotes the shaft axial length;  $D_{sh}$  is the shaft diameter; and  $k_{sh}$  represents the thermal conductivity of the shaft material

$$R_{th,sh} = \frac{h_{sh}}{\frac{\pi D_{sh}^2}{4} k_{sh}} \quad (6.11)$$

### 6.1.2 Convection modeling

Convection is considered at the external surfaces (e.g., the housing exposed to ambient air), at the end-windings in contact with the surrounding air, and within the air gap due to rotor rotation. Empirical correlations are used to evaluate relevant dimensionless numbers, from which the convective heat transfer coefficients are derived. This approach is widely adopted in the literature for the thermal analysis of electric machines across different topologies [128], [129]. The general expression for the convective heat transfer coefficient  $h_c$  is given in (6.12).

$$h_c = \frac{Nu_i k_{fluid}}{L_c} \quad (6.12)$$

Where,  $k_{fluid}$  is the thermal conductivity of fluid and  $L_c$  is the characteristic length of the surface.  $Nu_i$  is the Nusselt number whose formulation will depend on whether natural or forced convection is being considered.

#### Natural convection

Natural convection arises from temperature differences between the machine surface and the surrounding ambient, which induce fluid motion. The modeling of this phenomenon is based on empirical correlations developed for simplified geometries such as cylinders, plates, and disks. These correlations are expressed in terms of dimensionless numbers, whose formulations are provided in (6.13)-(6.16) [130].

$$Gr_i = \frac{g_{acc} \beta (T_{surf} - T_{amb}) L_{c,i}^3}{\nu^2} \quad (6.13)$$

$$Pr = \frac{\nu}{\alpha} \quad (6.14)$$

$$Ra_i = Gr_i \cdot Pr \quad (6.15)$$

$$Nu_i = \left( 0.6 + \frac{0.387 \cdot Ra_i^{1/6}}{[1 + (\frac{0.559}{Pr})^{9/16}]^{8/27}} \right)^2 \quad (6.16)$$

where  $g_{acc}$  denotes the gravitational acceleration;  $\beta$  is the thermal expansion coefficient;  $T_{surf}$  represents the surface temperature;  $T_{amb}$  is the ambient temperature;  $\nu$  is the kinematic viscosity of air and  $\alpha$  represents the thermal diffusivity.  $Gr_i$  is Grashof number that represents the ratio of the buoyancy force to the viscous force acting on the fluid. The Prandtl ( $Pr$ ) number determines whether momentum or heat diffuses faster in convective heat transfer. The Rayleigh number ( $Ra_i$ ) describes the relationship between buoyancy and viscosity within the fluid and  $Nu_i$  is the Nusselt number that represents the enhancement of heat transfer across fluid layers of the convection phenomena in comparison to pure conduction. In the proposed model, natural convection is considered between the housing and the ambient as well as between the end-winding and the internal air.

### Forced convection heat exchange

The proposed thermal network accounts for forced convection induced by rotor rotation, which generates airflow within the machine. This mechanism enhances heat transfer through fluid motion over the surfaces, with the corresponding convective heat transfer coefficient depending on both surface geometry and flow conditions.

As in the case of natural convection, these coefficients can be evaluated analytically using empirical correlations. However, due to the geometric complexity of the stator, particularly the presence of slotting, correlations derived for simplified geometries must be suitably adapted. In [129], analytical formulations and empirical correlations specific to disk-type machines, such as axial flux configurations, are presented. The expressions adopted in this work are summarized below.

$$Re = \frac{\Omega_{mot} D_2^2}{4\nu} \quad (6.17)$$

$$G = \frac{h_{gap}}{D_2/2} \quad (6.18)$$

Where  $\Omega_{mot}$  is the rotational speed in (rad/s) and  $h_{gap}$  is the air gap thickness. Equation (6.17) defines the Reynolds number  $Re$  that characterizes how laminar or turbulent the fluid is, impacting its cooling capabilities. Additionally, in (6.18)  $G$  is defined as the air gap ratio.

To evaluate the convective heat transfer coefficient, the Nusselt number must first be determined. In [129], empirical correlations are provided for the rotor-side Nusselt number,  $Nu_r$ , and the stator-side Nusselt number,  $Nu_s$ . The value of  $Nu_r$  depends on the flow regime being laminar or turbulent (where for a flat plate,  $Re = 5 \times 10^5$  is considered for the transition value between regimes) and the air gap ratio, with the corresponding formulations given in Table 6.2. For the stator side,  $Nu_s$  is calculated using the general expression defined in (6.19), as reported below.

$$Nu_s = A \cdot Re^B \quad (6.19)$$

where  $A$  and  $B$  are empirical coefficients reported in Table 6.3.

Table 6.2 Rotor heat transfer correlations [129].

| <b>G</b>  | <b>Regime</b> | <b>Correlation</b>                                                      |
|-----------|---------------|-------------------------------------------------------------------------|
| 0.01      | Lam.          | $Nu_r = 7.46 \cdot Re^{0.32}$                                           |
| 0.02-0.06 | Lam.          | $Nu_r = 0.5 \left(1 + \frac{5.47}{10000} e^{112G}\right) Re^{0.5}$      |
| 0.01      | Turb.         | $Nu_r = 0.044 \cdot Re^{0.75}$                                          |
| 0.02-0.06 | Turb.         | $Nu_r = 0.033 \left(12.57 \cdot e^{-33.18G}\right) Re^{0.6+25G^{12/7}}$ |

Finally, the convective heat transfer coefficients are obtained using (6.12). In the proposed LPTN, forced convection is considered for the stator teeth and end windings exposed to the air gap, as well as for the rotor surface facing the air gap and the rotor yoke in contact with the internal air. For simplicity, a common convective coefficient is assumed for the stator teeth and end-windings, while a

Table 6.3 Stator heat transfer correlations [129].

| <b>G</b> | <b>A</b> | <b>B</b> |
|----------|----------|----------|
| 0.0106   | 0.0790   | 0.640    |
| 0.0127   | 0.0888   | 0.633    |
| 0.0170   | 0.0406   | 0.682    |
| 0.0212   | 0.0315   | 0.691    |
| 0.0297   | 0.0347   | 0.679    |
| 0.0467   | 0.0234   | 0.712    |

second coefficient is adopted for both the rotor disk and the back of the rotor facing the end cap air.

### Convection resistance computation

Based on this procedure, the convective thermal resistances illustrated in Fig. 6.1 are determined as follows.

$$R_x = \frac{1}{h_x A_x} \quad (6.20)$$

The expression in (6.20) defines the convective thermal resistances as functions of the corresponding convective heat transfer coefficients and the associated heat exchange surface areas, depending on the subscript  $x$ . This subscript denotes:

- $ew, ia$  for natural convection of the portion of the end-winding exposed to the internal air within the housing;
- $hsg$  for natural convection between the housing and the ambient;
- $gap, s$  for forced convection between the stator and the air gap;
- $gap, r$  for forced convection between the rotor and the air gap;
- $gap, ew$  for forced convection of the end-winding surface facing the air gap;
- $rot$  for forced convection between the rotor yoke and the internal air;

The surface areas associated with each convective element can be readily determined from the machine geometry as detailed in section 4.2.4.

### Thermal capacity computation

To account for transient thermal effects, thermal capacitances are assigned to each node of the lumped parameter thermal network. Given the volume  $V_i$ , specific weight  $\rho_i$  and specific heat capacity  $c_i$  of each component, the corresponding thermal capacitance  $C_{th}$  can be computed using (6.21).

$$C_{th} = V_i \cdot \rho_i \cdot c_i \quad (6.21)$$

### Heat source components

The heat sources in the LPTN are associated with the losses generated in the corresponding machine components. In particular, the Joule losses in the windings can be calculated using (3.21).

The iron losses in both the stator and rotor, as well as the permanent magnet losses, are obtained through finite element simulations to ensure accurate evaluation under different operating conditions. Finally, a temperature source is connected to each convection resistance element to represent the ambient temperature surrounding the machine.

## 6.2 FEA Validation of the methodology

To validate the proposed thermal model, a 48-slot, 8-pole axial flux motor with distributed windings and a single-stage configuration is selected as a case study. The motor specifications are summarized in Table 6.4, while the materials used for each component, along with their corresponding thermal properties, are reported in Table 6.5. The computed convective heat transfer coefficients are presented in Table 6.6. Copper losses are calculated using (3.21), whereas stator iron losses and permanent magnet losses are obtained from electromagnetic finite element simulations.

Table 6.4 Study case machine data.

| Parameter             | Symbol        | Value   |
|-----------------------|---------------|---------|
| Slots                 | $N_c$         | 48      |
| Poles                 | $p$           | 8       |
| Inner diameter        | $D_1$         | 119 mm  |
| Outer diameter        | $D_2$         | 191 mm  |
| Core stack length     | $L_{core}$    | 36 mm   |
| Tooth inner width     | $w_{t1}$      | 2.6 mm  |
| Tooth outer width     | $w_{t2}$      | 7.6 mm  |
| Slot depth            | $h_{ts}$      | 24.5 mm |
| Slot width            | $w_s$         | 5.2 mm  |
| Stator yoke thickness | $h_{ys}$      | 14.6 mm |
| Air gap thickness     | $h_{gap}$     | 1 mm    |
| PM height             | $h_{PM}$      | 4 mm    |
| PM to pole ratio      | $\alpha_{PM}$ | 0.64    |
| Shaft diameter        | $D_{sh}$      | 20 mm   |
| Housing length        | $h_{sh}$      | 151 mm  |

Table 6.5 Study case material data [90], [126].

| Motor part            | Conductivity<br>(W/mK) | Specific heat<br>(J/kgK) | Density (kg/m3) |
|-----------------------|------------------------|--------------------------|-----------------|
| Stator/rotor iron     | 40                     | 500                      | 7650            |
| Winding<br>insulation | 0.1                    | 1000                     | 1200            |
| Permanent<br>magnet   | 9                      | 440                      | 7700            |
| Housing               | 44                     | 500                      | 7700            |
| Shaft                 | 47                     | 450                      | 7850            |
| Rotor holder          | 167                    | 896                      | 2710            |



Table 6.6 Convective coefficient values computed for the study case.

| Motor part               | Symbol      | Value ( $\text{W}/\text{m}^2\text{K}$ ) |
|--------------------------|-------------|-----------------------------------------|
| Stator iron - air gap    | $h_{gap,s}$ | 20.50                                   |
| rotor–air gap            | $h_{gap,r}$ | 37.36                                   |
| End-winding–internal air | $h_{ew,ia}$ | 15.50                                   |
| housing surface–ambient  | $h_{hsg}$   | 9.98                                    |

For validation purposes, a three-dimensional thermal finite element model is developed to simulate heat transfer within the machine. The model geometry and the resulting temperature distributions, presented in both isometric and cross-sectional views, are shown in Fig. 6.4. The steady-state temperatures at the nodes identified in Fig. 6.1 are reported in Table 6.7. A comparison between the predicted and simulated temperatures shows good agreement, with discrepancies below 5% for all nodes.

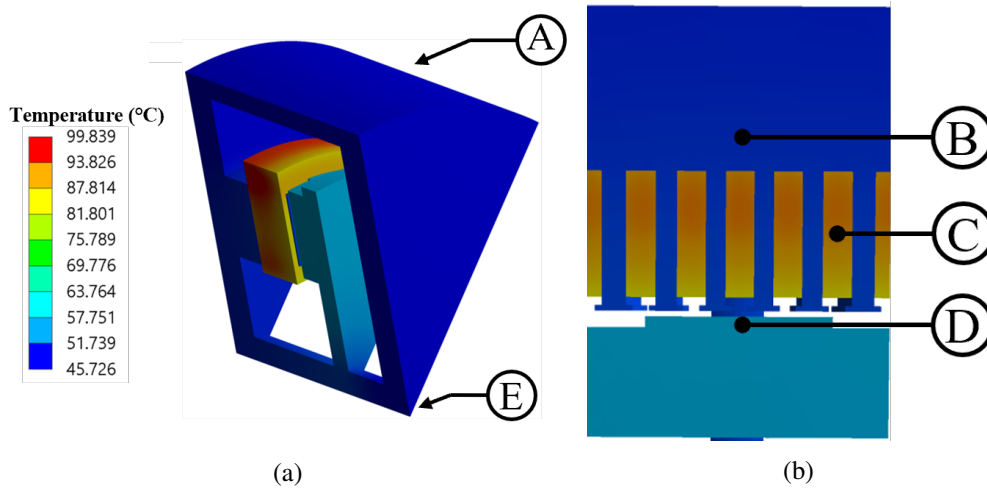


Fig. 6.4 FEA results: (a) isometric view, (b) cross section view.

In addition, transient simulations are carried out using both finite element analysis and the lumped parameter thermal network. Fig. 6.5 compares the temperature profiles obtained with the two approaches for all the nodes depicted in fig. 6.1. The results show good agreement over the simulated time interval, with a maximum deviation of 10%.

The larger discrepancies observed at earlier time instants can be attributed to the inherent limitations of the LPTN approach. In order to preserve model simplicity

Table 6.7 Temperature results comparison with FEA.

| Motor part (node)     | LPTN  | FEA   | $\Delta$ (%) |
|-----------------------|-------|-------|--------------|
| Housing (A)           | 63.31 | 63.79 | 1%           |
| Stator yoke (B)       | 64.08 | 65.25 | 2%           |
| Winding (C)           | 72.69 | 76.61 | 5%           |
| Permanent magnets (D) | 67.43 | 66.77 | -1%          |
| Shaft (E)             | 65.74 | 63.01 | -4%          |

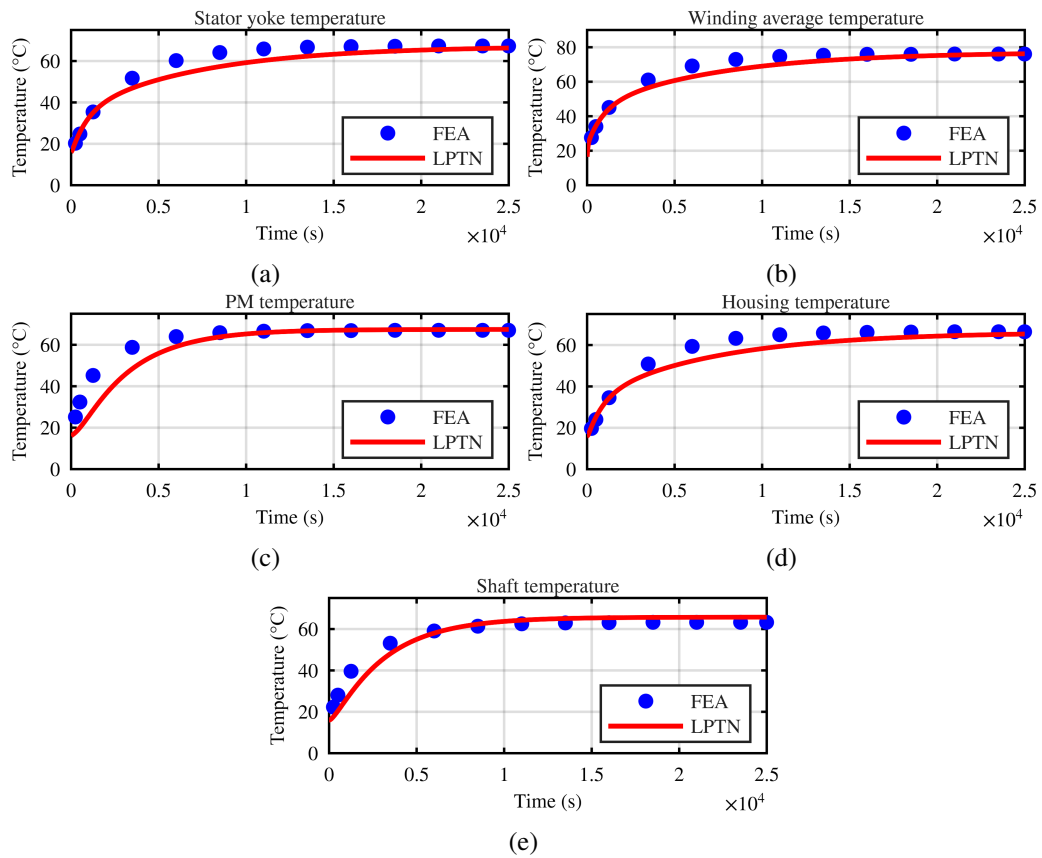


Fig. 6.5 Temperature over time comparison. (a) Stator yoke, (b) winding, (c) PM, (d) housing, (e) shaft.

and limit the number of nodes, relatively large thermal masses are aggregated and represented by a reduced number of thermal capacitance elements.

### 6.3 Experimental validation of the methodology

An additional validation is performed comparing with temperature measurements on the AFM prototype shown in Fig. 6.6, whose main specifications and operating conditions are summarized in Table 6.8. During the experimental tests, the temperatures of the three stator phases were recorded over time. When the proposed LPTN is applied without modifications, the predicted temperatures show significant discrepancies compared to the experimental results. This behavior is mainly attributed to the specific topology of the prototype, particularly the presence of ventilation holes in the housing near the end-winding regions. These openings promote airflow induced by rotor rotation, enhancing heat dissipation from the end-windings.

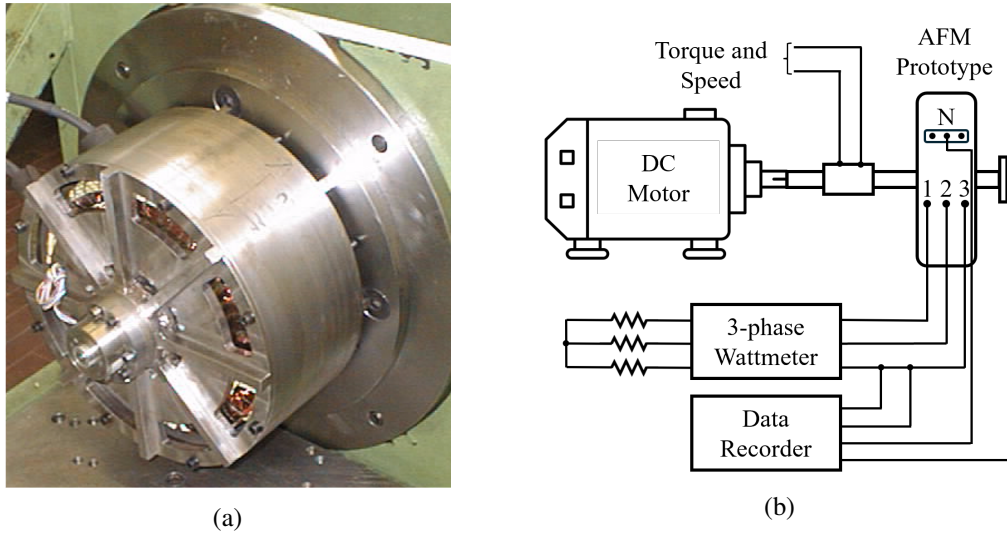


Fig. 6.6 (a) AFM prototype under testing, (b) experimental setup layout [131].

To account for this modified cooling scenario, additional convective thermal resistances, as illustrated in Fig. 6.7, are incorporated into the proposed LPTN. These represent: (i) forced convection in the end-windings due to airflow induced by rotor rotation, (ii) convection between the internal air flowing through the openings and the ambient, and (iii) natural convection at the rear portion of the end-windings exposed to the openings.

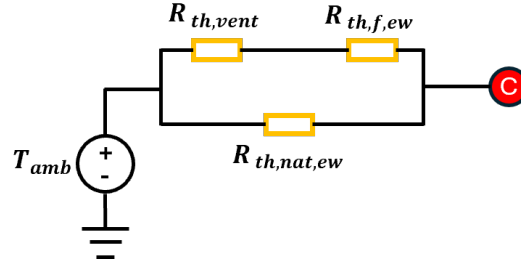


Fig. 6.7 Additional heat path to represent forced ventilation.

Table 6.8 Prototype machine data.

| Parameter                        | Value            |
|----------------------------------|------------------|
| Topology                         | Double stage AFM |
| Slots                            | 48               |
| Poles                            | 8                |
| Inner diameter                   | 119 mm           |
| Outer diameter                   | 191 mm           |
| Core stack length                | 36 mm            |
| Tooth inner width                | 2.6 mm           |
| Tooth outer width                | 7.6 mm           |
| Slot depth                       | 24.5 mm          |
| Slot width                       | 5.2 mm           |
| Stator yoke thickness            | 14.6 mm          |
| Air gap thickness                | 1 mm             |
| PM height                        | 4 mm             |
| PM to pole ratio                 | 0.64             |
| Housing inner diameter           | 270 mm           |
| Housing outer diameter           | 290 mm           |
| <b>Testing conditions</b>        |                  |
| RMS phase current                | 4.17 A           |
| Winding phase resistance (@75°C) | 5.62 $\Omega$    |
| Speed                            | 900 rpm          |
| Ambient temperature              | 16 °C            |

By approximating the end-winding region as a cylindrical surface, the corresponding Reynolds number ( $Re_{ew}$ ), Prandtl number ( $Pr_{ew}$ ) and Nusselt number ( $Nu_{ew}$ ) can be computed with (6.22), (6.23) and (6.24) respectively [130].

$$Re_{ew} = \frac{\rho_{air} V_{air} L_{ew}}{\mu_{air}} \quad (6.22)$$

$$Pr_{ew} = \frac{\mu_{air} c_{air}}{k_{air}} \quad (6.23)$$

$$Nu_{ew} = \frac{3}{10} + \frac{0.62 Re_{ew}^{1/2} Pr_{ew}^{1/3}}{[1 + (0.4/Pr_{ew}^{2/3})]^{1/4}} \left[ 1 + \left( \frac{Re_{ew}}{282000} \right)^{5/8} \right]^{4/5} \quad (6.24)$$

where,  $\rho_{air}$  is the air density,  $V_{air}$  is the air velocity at the end-winding section,  $L_{c,ew}$  is the characteristic length of the end-winding,  $\mu_{air}$  is the dynamic viscosity and  $c_{air}$  is the specific heat of the air. To estimate the air velocity, the rotor tip speed  $V_{tip}$  is first calculated using (6.25), based on the motor rotational speed  $n_{mot}$  expressed in rpm.

$$V_{tip} = \frac{\pi}{60} n_{mot} D_2 \quad (6.25)$$

To estimate the air velocity in the end-winding region, an empirical coefficient  $C$  is adopted based on literature data, in the absence of experimental measurements or CFD simulations [132–134]. Fig. 6.8 shows the fanning function  $C$  for axial flux machines. This function defines the relative air velocity at the end-winding region as a function of the rotor radius. Although these values may vary for different machines, they can be used as a reference for reasonable estimations. For the study presented in this section a value of 0.3 was choose.

Accordingly, the air velocity in the end-winding region is calculated using (6.26).

$$V_{air} = V_{tip} \cdot C \quad (6.26)$$

For this model, the velocity at the top of the end-winding region is considered to be 30% of  $V_{tip}$ . Once (6.22)–(6.24) are evaluated, the convective heat transfer

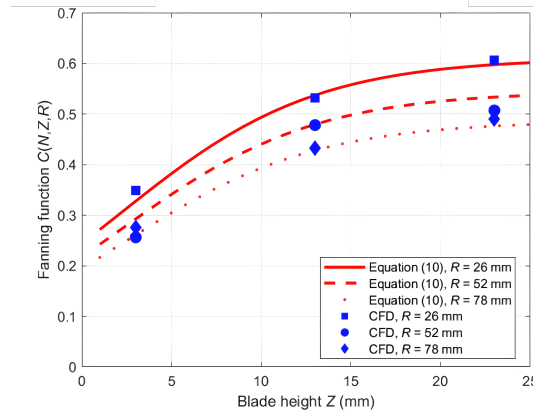


Fig. 6.8 Internal air velocity over the winding (left) and in the end-cap region (right) [133].

coefficient  $h_{f,ew}$  is determined using (6.12). The corresponding thermal resistance is then computed using (6.27), based on the upper end-winding surface area  $A_{ew,top}$ .

$$R_{f,ew} = \frac{1}{h_{f,ew}A_{ew,top}} \quad (6.27)$$

The heat exchange between the internal air and the external ambient through the ventilation openings is modeled by the thermal resistance  $R_{vent}$ , which is estimated using (6.28).

$$R_{vent} = \frac{1}{\rho_{air}A_{vent}V_{exit}c_{air}} \quad (6.28)$$

In (6.28),  $c_{air}$  is the specific heat of the air,  $A_{vent}$  denotes the total surface area of the ventilation openings, and  $V_{exit}$  represents the estimated air velocity at the outlet, also approximated based on typical values reported in the literature [132, 134]. Moreover, for this work  $V_{exit}$  is considering again the fanning function from the air velocity over the end-winding region. Finally, the thermal resistance associated with natural convection between the rear portion of the end-windings and the ambient  $R_{nat,ew}$  can be computed as follows.

$$R_{nat,ew} = \frac{1}{h_{nat,ew}A_{ew}} \quad (6.29)$$

In (6.29),  $A_{ew}$  is the area of the end-winding portion facing the ventilation hole and  $h_{nat,ew}$  is the natural convective coefficient.

In addition to the experimental tests carried out using the setup shown in Fig. 6.6b, an additional set of measurements was performed to determine both the thermal resistance between the housing and the ambient and the equivalent resistance between the winding and the motor iron. For the housing thermal resistance computation, a DC test was conducted by connecting all stator windings in series and recording the temperatures of the housing and each phase. The experimental setup is shown in Fig. 6.9, with all instruments detailed, while the tested machine is depicted in Fig. 6.10.

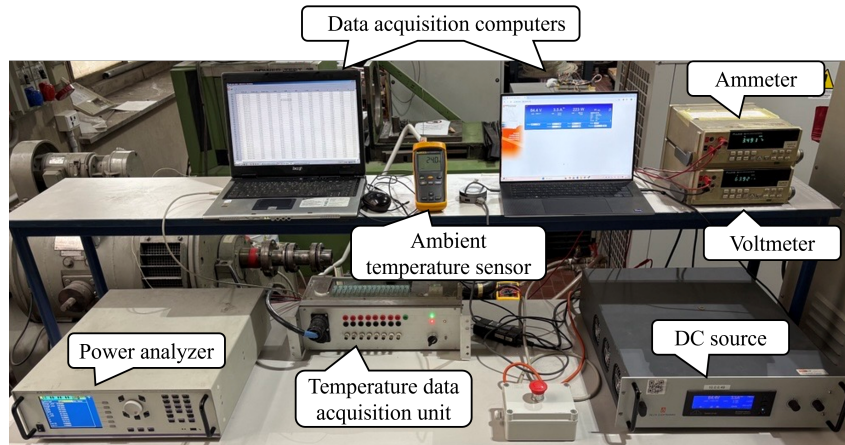


Fig. 6.9 DC thermal test setup.



Fig. 6.10 AFM under the DC thermal test.

The test was conducted until steady-state conditions were reached. According to the IEEE 115-2019 standard, steady state is achieved when the temperature rise remains below  $1^{\circ}\text{C}$  over a 30-minute interval [135]. The temperature recorded over time is presented in Fig. 6.11. Once the measurements were completed, the housing thermal resistance was computed using (6.30), as shown below.

$$R_{th,hsg} = \frac{T_{hsg} - T_{amb}}{V_{DC}I_{DC}} \quad (6.30)$$

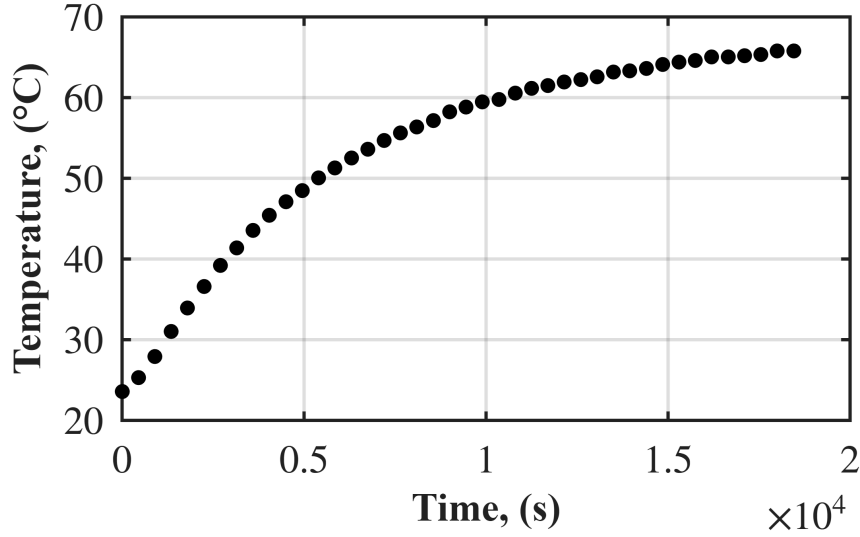


Fig. 6.11 Housing temperature measurements.

where  $T_{hsg}$  is the housing surface temperature, and  $V_{DC}$  and  $I_{DC}$  are the DC voltage and current applied during the thermal test, respectively. It is worth noting that the convection coefficient obtained from (6.30) shows only a 3% deviation compared to the value estimated using the methodology presented in Section 6.1.2.

To determine the equivalent thermal resistance between the winding and the iron, a short thermal transient test was performed as described in [136]. Using the setup shown in Fig. 6.9, the machine was supplied with DC current for a short duration (less than 10 minutes), during which Joule losses are the only active heat source.

From the measured voltage  $V_{DC}$  and current  $I_{DC}$ , the variation of the winding resistance and the thermal energy stored in the winding capacitance can be accurately determined. According to (6.31), the equivalent thermal capacitance  $C_{eq}$  is determined by fitting the time evolution of the thermal energy  $W_{wdg}$  as a function of the winding temperature  $T_{wdg}$  using linear regression.

$$C_{eq} = \frac{dW_{wdg}}{dT_{wdg}} \quad (6.31)$$



The thermal energy stored in the winding thermal capacitance is computed by means of the measured electrical quantities as the DC power  $P_{DC} = V_{DC} \cdot I_{DC}$  accumulated in time  $t$  as defined below.

$$W_{wdg} = P_{DC} \cdot t \quad (6.32)$$

The equivalent thermal resistance  $R_{wdg,eq}$  is determined by fitting the measured winding temperature evolution using a least-squares method. The fitting is based on the first-order differential equation describing a first-order thermal network, as defined in (6.33).

$$T_k = T_{k-1} + (T_{amb} + R_{wdg,eq}P_{k-1} - T_{k-1}) \left( 1 - e^{-\frac{t_k - t_{k-1}}{R_{wdg,eq}C_{eq}}} \right) \quad (6.33)$$

where the subscript  $k$  defines the current sample of the variable measured and  $t_k - t_{k-1}$  defines the sample time.

For the prototype presented in Fig. 6.12, the energy storage in the winding with its linear regression equation and the winding temperature evolution over time with the fitted curve after the equivalent thermal resistance was computed.

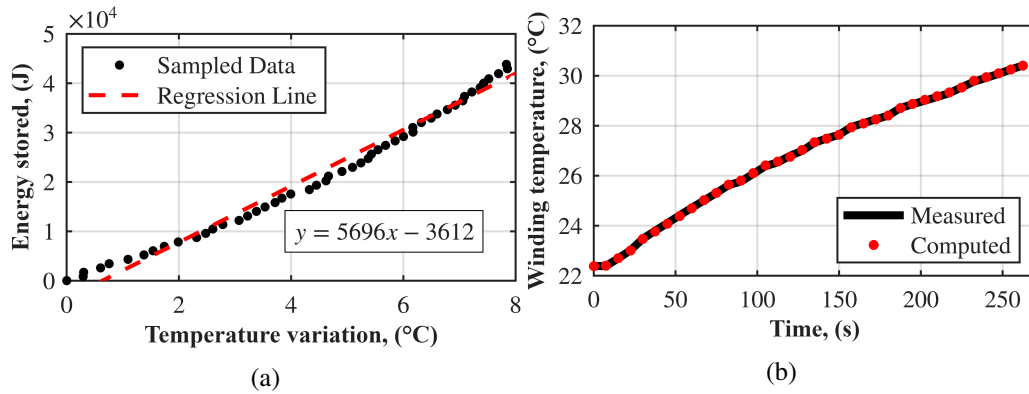


Fig. 6.12 Temperature over time comparison. (a) Energy stored in the winding, (b) Winding temperature over time.

Once all the parameters were computed, they were used for the thermal network developed and Figure 6.13 compares the measured temperature profiles with the predictions obtained from the proposed LPTN model. The results show good agreement between the experimental data and the model, with a maximum deviation of less than 5% relative to the measured values.

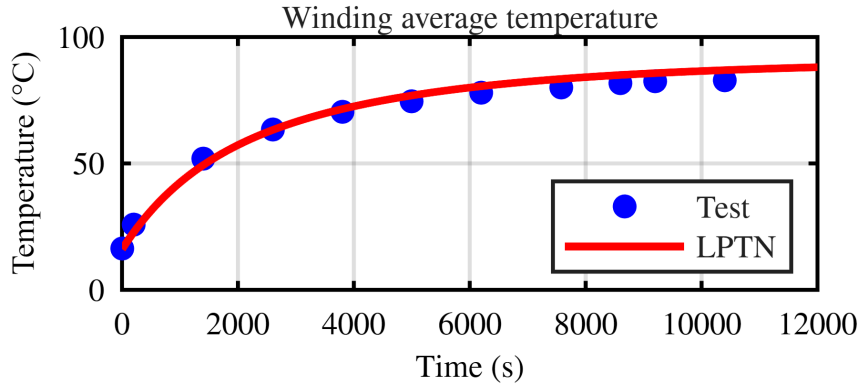


Fig. 6.13 Comparison between measurements and LPTN predictions.

## 6.4 LPTN sensitivity analysis

An additional aspect of interest is the impact of further simplifying the thermal network while maintaining acceptable accuracy. The winding representation constitutes the most critical part of the model, due to the high associated losses, the difficulty in accurately estimating bulk material properties, and the complexity of heat transfer paths to the surrounding air. This section therefore investigates different simplification approaches for modeling the windings within the LPTN.

First, the convective resistance  $R_{th,ew,ia}$ , corresponding to the heat exchange between the upper part of the winding and the internal air within the housing, is neglected (Case 1). In a second scenario (Case 2), both  $R_{th,ew,ia}$  and the convective resistance between the end-winding and the air gap,  $R_{th,gap,w}$ , are omitted. The resulting temperature distributions are compared with those obtained from the finite element thermal simulation and the original proposed thermal network (used as the benchmark). The comparison of the different approaches is presented in Fig. 6.14 for the winding temperature, as this represents a critical node in the thermal behavior of the machine.

The results indicate that neglecting the convective elements associated with the winding region in the LPTN leads to significant deviations in temperature predictions compared to the reference finite element simulations. In particular, when convection in the winding region is omitted, the average winding temperature is overestimated by up to 20%. This increase can be attributed to the low thermal conductivity between the windings and the stator iron, which becomes the only available heat dissipation path for Joule losses in Case 2.

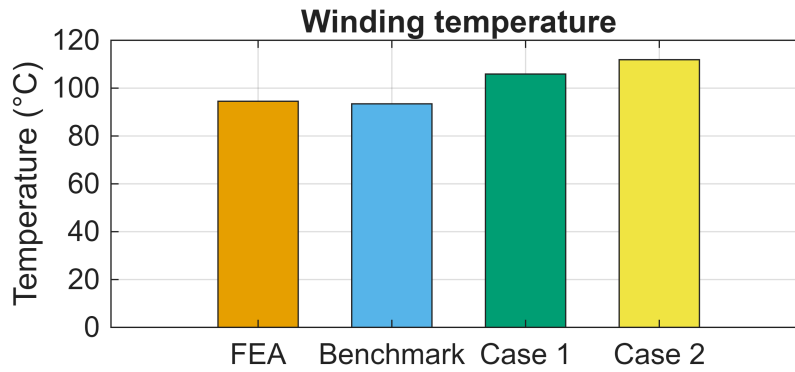


Fig. 6.14 Thermal network topology sensitivity analysis results.

In the same context, it is also important to quantify the impact of the equivalent thermal conductivity of the insulation system on the temperature distribution within the AFM. To this end, two additional scenarios are evaluated against both the FEA results and the original LPTN as reference cases. These scenarios consider a 50% increase (Case A) and a 50% decrease (Case B) in the insulation thermal conductivity. The resulting winding temperatures under these conditions are presented in 6.15.

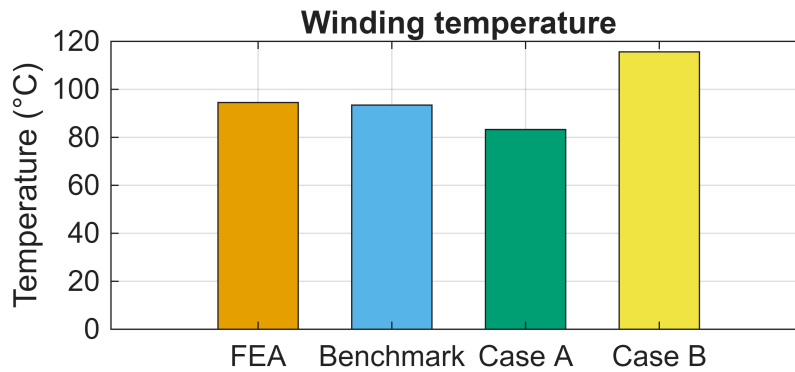


Fig. 6.15 Thermal network insulation conductivity sensitivity analysis results.

The results shown in 6.15 indicate that variations in the thermal conductivity of the insulation system have a significant impact on heat dissipation in the windings. In particular, Case A exhibits a temperature increase of 24%, while Case B shows a temperature reduction of approximately 11%.

This sensitivity analysis leads to two main conclusions regarding insulation modeling in LPTNs. First, for accurate thermal analysis, a precise estimation of the insulation thermal conductivity is essential, as it strongly influences the predicted temperature distribution. Second, from a design perspective, the selection

of insulation materials with higher thermal conductivity can significantly enhance the thermal performance of the machine.

## 6.5 Chapter conclusion

This chapter presented a simplified lumped parameter thermal network tailored for axial flux permanent magnet machines. The proposed model achieves a balance between accuracy and simplicity, avoiding the need for excessive model detail, which often limits the applicability of thermal methodologies in early-stage design. By incorporating representative equivalent thermal parameters from the literature, the methodology enables reliable temperature prediction with reduced modeling effort.

A generalized thermal network topology was first developed, including analytical formulations for all conduction and convection elements. Empirical correlations specific to disk-type geometries were integrated to accurately capture convective heat transfer mechanisms. Validation against three-dimensional finite element thermal simulations showed strong agreement, confirming the capability of the model to represent the dominant heat transfer paths. Additional experimental validation demonstrated good agreement between predicted and measured temperatures, further supporting the effectiveness of the proposed approach.

Finally, a sensitivity analysis highlighted the critical role of end-winding convection and insulation thermal conductivity in determining the temperature distribution. Neglecting or inaccurately estimating these effects was shown to result in significant deviations from finite element predictions, emphasizing their importance in both modeling and design.

# Chapter 7

## Conclusion and future work

### 7.1 Conclusion

This thesis has addressed the analysis and design of axial flux permanent magnet machines in the context of aircraft electrification, with a particular focus on developing efficient, accurate, and computationally affordable methodologies for early-stage design. The increasing electrification of aircraft systems, driven by the need to reduce emissions, improve efficiency, and lower operating costs, represents a fundamental shift in aerospace engineering. In this transition, electric machines play a pivotal role, especially in the evolution from More Electric Aircraft toward fully electric propulsion concepts.

From the analysis presented, it is evident that the electrification of propulsion systems constitutes one of the most promising pathways toward sustainable aviation. However, this transition imposes stringent requirements in terms of power density, efficiency, thermal management, and reliability. Within this framework, axial flux permanent magnet machines emerge as particularly suitable candidates due to their compact geometry, high torque density, and compatibility with direct-drive configurations. These characteristics make them especially attractive for applications where weight reduction and efficiency are critical, such as distributed propulsion and hybrid-electric aircraft architectures.

Building on this context, the core contribution of this thesis lies in the development of a set of complementary methodologies that enable a fast, reliable, and integrated design workflow for axial flux machines. Particular emphasis has been

placed on analytical and reduced-order approaches, which are essential to support iterative design and optimization processes while maintaining manageable computational requirements.

The first key contribution is the development of a self-consistent analytical sizing procedure for axial flux permanent magnet machines. The proposed method allows designers to obtain an initial machine geometry that satisfies electromagnetic constraints through an iterative yet computationally efficient process. Its validation against finite element simulations and experimental prototypes demonstrates that analytical approaches, when properly formulated, can provide reliable predictions even at early design stages. This is particularly relevant in the context of electric machines design workflows, where reducing development cycles and enabling rapid design exploration directly contributes to accelerating energy transition strategies.

The second major contribution is the formulation of a magnetic equivalent circuit model capable of capturing the electromagnetic behavior of axial flux machines with high fidelity. By incorporating nonlinear material properties, rotor motion, and leakage effects, the proposed MEC achieves a balance between accuracy and computational efficiency. Its flexibility allows the analysis of different machine configurations and operating conditions, making it a powerful tool for design space exploration and optimization. Compared to conventional finite element approaches, the significant reduction in computation time enables the evaluation of a large number of design candidates, which is essential for identifying optimal solutions in complex, multidisciplinary problems such as aircraft propulsion systems.

The third key contribution is the development of a lumped parameter thermal network tailored to axial flux machines. Thermal management is a critical aspect in high-performance electrical machines, particularly in aerospace applications where cooling capabilities are limited and reliability requirements are stringent. The proposed thermal model provides accurate temperature predictions while maintaining a simplified structure suitable for early-stage design. Its validation against both numerical and experimental results confirms its capability to capture the dominant heat transfer mechanisms. Moreover, the sensitivity analysis highlights the importance of accurately modeling end-winding convection and insulation properties.

Taken together, these three methodologies form a coherent and integrated design framework that addresses electromagnetic and thermal aspects in a consistent and computationally efficient manner. Their analytical nature enables seamless

integration into optimization algorithms and multiphysics workflows, supporting the development of advanced electric machine designs tailored to the demanding requirements of aircraft electrification.

In conclusion, this thesis demonstrates that analytical and reduced-order modeling approaches, when carefully developed and validated, can play a crucial role in the design of next-generation electric machines. By bridging the gap between accuracy and computational efficiency, these methods provide valuable tools for supporting the transition toward more sustainable and electrified aviation systems.

## 7.2 Future work

Despite the promising results achieved in this work, several opportunities remain for further development and extension of the proposed methodologies, particularly to address a broader range of operating conditions and design requirements.

The magnetic equivalent circuit model can be further enhanced by incorporating the capability to predict secondary electromagnetic effects, such as cogging torque and torque ripple. Moreover, the inclusion of electromagnetic force calculations for single-stage configurations would enable straightforward integration with mechanical and vibration analysis tools, supporting a more comprehensive multiphysics design framework.

With regard to the thermal model, additional experimental validation is recommended to further assess its robustness. In particular, future testing campaigns should include not only winding temperature measurements but also stator core and permanent magnet temperature data, allowing a more complete evaluation of the thermal behavior and heat transfer mechanisms within the machine.

Future developments may also extend the proposed methodologies to account for additional physical phenomena, such as advanced cooling strategies, mechanical stress and structural analysis, and fault-tolerant operation. Furthermore, their application to emerging machine topologies and hybrid propulsion architectures represents a promising direction for future research. In this context, conducting a comprehensive trade-off analysis based on a real application—such as the design of an electric motor for small aircraft propulsion—would provide valuable insights into the practical implementation of the proposed approaches.

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Finally, the integration of the developed models into fully automated optimization environments represents a key opportunity for accelerating the design process. Such integration would enable efficient exploration of large design spaces and support the development of high-performance, sustainable electric propulsion systems for future aircraft applications.



# List of publications

The following papers were published during the realization of the Ph.D program:

- N. Reyes, M. Jiménez, Á. Hoffer, C. Madariaga, C. Gallardo, J. A. Tapia, W. Jara, and M. Degano, “Structural Feasibility and Electromagnetic Analysis of an Ironless Rotor Axial Flux Permanent Magnet Synchronous Machine,” in Proc. 2023 IEEE Energy Conversion Congress and Exposition (ECCE), Nashville, TN, USA, Oct. 29–Nov. 2, 2023, pp. 4297–4301, doi: 10.1109/ECCE53617.2023.10362936.
- M. J. Molina, F. Graffeo, S. Vaschetto and A. Tenconi, "Application and Performance Trends of Electric Motors for Aircraft Propulsion," 2024 IEEE Transportation Electrification Conference and Expo (ITEC), Chicago, IL, USA, 2024, pp. 1-8, doi: 10.1109/ITEC60657.2024.10598961.
- A. Tenconi, F. Crescimbin, M. Tursini, A. Tani, M. Jiménez Molina, L. Vita, A. Credo, A. Bellini, L. Perilli, G. Fabri, G. A. Cagliari, and G. Sala, “Multi-Phase Fault Tolerant MW Range Synchronous Machine Drives for Hybrid-Electric Aircraft,” in Proc. 2024 IEEE Int. Conf. Electrical Systems for Aircraft, Railway, Ship Propulsion and Road Vehicles & Int. Transportation Electrification Conf. (ESARS-ITEC), Nov. 26–29, 2024, pp. 1–7, doi: 10.1109/ESARS-ITEC60450.2024.10819864.
- M. J. Molina, F. Graffeo, S. Vaschetto, A. Tenconi and A. Cavagnino, "Fast Procedure for the Electromagnetic Sizing of Axial Flux PM Motors," 2024 International Conference on Electrical Machines (ICEM), Torino, Italy, 2024, pp. 1-8, doi: 10.1109/ICEM60801.2024.10700359.
- F. Graffeo, S. Rubino, M. J. Molina, S. Vaschetto and A. Tenconi, "Efficiency Mapping of Electrically Excited Synchronous Motors with Different

- Control Strategies," 2025 IEEE Workshop on Electrical Machines Design, Control and Diagnosis (WEMDCD), Valletta, Malta, 2025, pp. 1-6, doi: 10.1109/WEMDCD61816.2025.11014137.
- M. Recabarren, C. Madariaga-Cifuentes, C. Gallardo, F. Graffeo, M. Jimenez, G. Perez-Guirriman, and J. A. Tapia, "Comparison of Optimized Electrically Excited Wound-Field and Permanent Magnet Synchronous Motors for Light EV Applications," in Proc. IECON 2025 – 51st Annu. Conf. IEEE Industrial Electronics Society, Madrid, Spain, 2025, pp. 1–6, doi: 10.1109/IECON58223.2025.11221476.
  - M. Jiménez, F. Graffeo, S. Vaschetto, A. Tenconi, and A. Cavagnino, "Magnetic Equivalent Circuit for Multistage Axial Flux Surface Permanent Magnet Machines," IEEE Open Journal of Industry Applications, vol. 7, pp. 327–337, 2026, doi: 10.1109/OJIA.2026.3665408.
  - A. Hoffer, F. Alarcón, M. Jiménez, D. Obando, B. Pavez, H. Young, J. Coloma, J. Mac-kay, F. Ruiz, and J. Rodríguez, "Study of Demagnetization in a Coreless Axial-Flux Permanent Magnet Synchronous Generator for Small Hybrid Vertical-Axis Wind Turbines," in Proc. IEEE Chilean Conf. Electrical, Electronics Engineering, Information and Communication Technologies (CHILECON), 2025, in press.
  - M. Jiménez, F. Graffeo, S. Vaschetto, A. Tenconi, and A. Cavagnino, "Simplified Thermal Network for Early-Stage Design of Axial Flux Permanent Magnet Machines," submitted to Int. Conf. Electrical Machines (ICEM), 2026.

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# Appendix A

## Motor Survey Data

This appendix reports the data collected and used to generate the analysis presented in Fig. 2.6 in subsection 2.4.

The following remarks apply to the data reported in the table:

- In the topology column, a distinction is made for PMSMs between inner rotor (IR) and outer rotor (OR) configurations;
- The liquid cooling category may include both water and oil cooling, depending on the specific case;
- The total mass is calculated considering the motor mass reported in the references, including the cooling system when such information is available;
- In the specific power column, values highlighted in red indicate quantities computed from the available data, whereas values in black correspond to those explicitly reported in the references;
- The results shown in Fig. 2.6 include only cases for which sufficient data is available; therefore, some entries may appear in certain plots and not in others.

Table A.1 List of study cases included in the electric machine survey

| Study Case                 | Topology  | Rated (kW) | Power | Rated (rpm) | speed | Cooling type | Total (kg) | Mass | Specific (kW/kg) | Power | Year | Reference |
|----------------------------|-----------|------------|-------|-------------|-------|--------------|------------|------|------------------|-------|------|-----------|
| Skypark                    | PMSM - OR | 60         |       | 2500        |       | LIQUID       | 30         |      | 2.00             |       | 2009 | [28]      |
| ALICE                      | PMSM      | 350        |       | 2089        |       | LIQUID       | 128        |      | 2.73             |       | 2022 | [29]      |
| NASA X-57 Maxwell          | -         | 72         |       | -           |       | AIR          | 53         |      | 1.36             |       | 2016 | [30]      |
| ACS-Itaipu Sora-E          | AFPM      | 37         |       | 6309        |       | LIQUID       | 7.6        |      | 4.87             |       | 2015 | [31]      |
| Spirit of Innovation       | AFPM      | 200        |       | 2200        |       | AIR          | 0          |      | 5.00             |       | 2021 | [32]      |
| Bye Aerospace eFlyer 2     | -         | 100        |       | 3000        |       | -            | 0          |      | 3.50             |       | 2018 | [33]      |
| Project DragonFly          | -         | 60         |       | 7162        |       | -            | 18         |      | 3.33             |       | 2019 | [34]      |
| Airlander 10               | PMSM      | 500        |       | 2000        |       | -            | 55.6       |      | 9.00             |       | 2022 | [35]      |
| E-Fan X                    | -         | 2000       |       | 6500        |       | LIQUID       | 261        |      | 7.66             |       | 2017 | [36]      |
| RRP70D, e-Flyer2           | -         | 70         |       | 2600        |       | AIR          | 26         |      | 5.22             |       | 2020 | [37]      |
| RRP200D                    | -         | 206        |       | 1300        |       |              | 49         |      | 5.22             |       | 2020 | [37]      |
| HPDM-30                    | PMSM      | 33         |       | 19945       |       | LIQUID       | 4.1        |      | 8.00             |       | 2023 | [38]      |
| HPDM-250                   | PMSM      | 200        |       | 19999       |       | LIQUID       | -          |      | 14.29            |       | 2023 | [38]      |
| HPDM-3000                  | PMSM      | 2800       |       | 1671        |       | -            | -          |      | 12.73            |       | 2022 | [38]      |
| CIAG-250-50                | -         | 20.1       |       | 1600        |       | AIR          | -          |      | 1.20             |       | 2023 | [39]      |
| SP260D0 Siemens            | PMSM      | 260        |       | 2500        |       | LIQUID       | -          |      | 5.20             |       | 2015 | [40]      |
| EMRAX 348 (liquid cooling) | AFPM      | 178        |       | 3999        |       | LIQUID       | 43.9       |      | 4.05             |       | 2016 | [41]      |
| EMRAX 348 (air cooling)    | AFPM      | 168        |       | 4011        |       | AIR          | 43.1       |      | 3.90             |       | 2016 | [41]      |
| MAGNIX 250                 | -         | 280        |       | 1900        |       | -            | -          |      | 5.50             |       | 2018 | [42]      |
| Solar Plane                | PM-DC     | 6          |       | 800         |       | AIR          | 18         |      | 0.50             |       | 2005 | [43]      |
| 5-phase Motor              | PMSM      | 73.5       |       | -           |       | AIR          | 45.9       |      | 1.60             |       | 2016 | [44]      |
| HPD Motor                  | PMSM - OR | 45         |       | 3500        |       | LIQUID       | 18         |      | 5.00             |       | 2023 | [45]      |
| Nottingham Motor           | PMSM - IR | 500        |       | 3000        |       | LIQUID       | 116        |      | 4.80             |       | 2019 | [46]      |
| SCEPTOR                    | PMSM - OR | 60         |       | 2400        |       | AIR          | 35         |      | 2.73             |       | 2016 | [47]      |
| Air-core                   | PMSM - OR | 1000       |       | 18000       |       | AIR          | 70.94      |      | 14.00            |       | 2016 | [48]      |
| HTS Motor                  | WFSM-HTS  | 1400       |       | 6800        |       | LIQUID       | 81.9       |      | 17.40            |       | 2019 | [49]      |
| High performance PMSM      | PMSM - IR | 20         |       | 3000        |       | AIR          | 9.09       |      | 2.20             |       | 2018 | [50]      |
| High Speed Wisconsin       | PMSM - IR | 1000       |       | 20000       |       | -            | 42.2       |      | 23.70            |       | 2019 | [51]      |
| MIT Motor                  | PMSM - IR | 1000       |       | 12500       |       | AIR          | 57.4       |      | 17.00            |       | 2023 | [52]      |
| Slot-less Motor            | PMSM - OR | 300        |       | 3600        |       | AIR          | 54         |      | 5.56             |       | 2021 | [53]      |