

Sizing Electrical Machines Using OpenOffice Calc

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Abstract

Sizing an electrical machine can simply be achieved by introducing the concepts of « electric loading » and « magnetic loading ». The aim is to predict the main dimensions of the active parts such as windings, magnetic circuits, slots and magnets.

We have developed a computational tool under the open source spreadsheet application “OpenOffice Calc”. This choice was encouraged by the fact that a collaborative work needed by the IMOTHEP project allows different users to easily share and integrate new data. This brings some comfort in carrying out repetitive calculations for machine optimizations.

1. Introduction

Sizing an electrical machine can simply be achieved by introducing the concepts of « electric loading » and « magnetic loading ». The aim is to predict the main dimensions of the active parts such as windings, magnetic circuits, slots and magnets [1-3]. This analytical sizing requires less computational effort compared to numerical computations while the precision remains acceptable.

In the frame of IMOTHEP project (<https://www.imothep-project.eu/>), several configurations of hybrid-electric propulsion aircrafts with different electrical power requirements are considered. It was then necessary to design electric generators coupled to a turbomachine whose data are given by other partners in the project.

For this purpose, we have developed a computational tool under the open source spreadsheet application “OpenOffice Calc”. This choice was encouraged by the fact that a collaborative work on spreadsheets allows different users to easily share and integrate new data. This brings some comfort in carrying out the needed repetitive calculations. Furthermore, Calc’s solver allows performing in a very simple way single objective constrained optimizations problems using metaheuristic algorithms.

In this paper, we have conducted many optimizations whose objective was the reduction of the machine mass to increase power to weight density of surface mounted permanent magnets (PMSM) machines. The developed computational tool integrates the following features related to the PMSM design:

- Electromagnetic sizing by means of analytical models based on the resolution of 2D magnetostatic problems in the air gap of the PMSM
- Winding temperature evaluation based on a lumped parameter thermal model

These features can be evaluated separately or coupled in a multi-physics way to ensure a robust optimized design of the machine. Only few minutes are required to get the results.

2. Sizing equations of ac electrical machines

For multiphase (3-phase is most common) alternating current (ac) machines, the principle of operation imposes a p pole-pair air-gap rotating magnetic fields created by sinusoidal currents (armature) or revolving magnetic poles (due to dc currents or magnets in inductor). In steady state, the torque is constant over the time if the normal flux density (b_n) and the tangential field strength (h_t) in the air-gap are sinusoidal waves with respect to the angular position.

The starting point for the sizing is the expression of the electromagnetic power transmitted throughout the air-gap. For a machine rotating at the speed Ω in rad/s, the transmitted power (or electromagnetic power) is expressed by:

$$P_t = 2\Omega \cdot k_w \cos(\psi) \cdot (BH) \cdot \pi R^2 L \quad (1)$$

With:

k_w : winding factor (fundamental wave with p pole-pairs)

ψ : phase shift between the tangential field (h_t) and the normal flux density (b_n) in the air-gap (or between the phase emf E and current I).

R [m]: armature bore radius

L [m]: axial length

B [T]: air-gap rms normal (radial) flux density (magnetic loading)

H [A/m]: air-gap rms tangential magnetic field (electric loading)

H is the “electric loading”. It is related to the RMS current I in the conductor by:

$$H = \frac{N_t I}{2\pi R} \quad (2)$$

Where N_t represents the total number of conductors of the armature winding.

2.1 Main steps for sizing ac machines

The machine specifications usually impose the mechanical power and the speed Ω . Hence, the torque is $\Gamma = P_t/\Omega$. Most of the time, the voltage is also specified.

Only 3-phase winding is considered here.

1st step: determination of R and L

Evaluate the product R^2L after choosing B and H :

B is limited by the iron saturation; its value is usually in the interval [0.3-0.8] T

H is limited by the winding heating and the cooling capabilities; for forced air cooling $H=30-80$ kA/m

Initial reasonable values for k_w and $\cos(\psi)$ are needed; $k_w = 0.9 - 0.95$, $\cos(\psi) = 0.8 - 1$ depending on the studied machine and its power.

From relation (1), calculate R^2L . Then for the considered number of pole-pairs p , a suitable ratio L/R can be affected [1]-[2] which allows the determination of R and L .

Note that relation (2) allows the computation of the total amperes-turns $N_t I$ of the machine.

2nd step: winding, number of turns, current and conductor cross-section area

It is preferable to adopt an integer number q of slots/pole/phase. Moreover, with q even, a coil pitch to pole pitch ratio of $5/6$ is always feasible (2 layer winding) to reduce 5th and 7th air-gap mmf harmonics.

The number of slots of the machine is $N_s = 6pq$.

The number of coils/phase is $N_{ph}=N_s/6$ for a double layer winding and $N_{ph}=N_s/3$ for a single layer winding.

The phase-phase voltage U being specified, the phase voltage is $V = U$ for delta connection or $V = U/\sqrt{3}$ for star connection. The emf RMS value is:

$$E = 2Nk_w\Omega RLB \quad (3)$$

Where N is the number of turns/phase, $N=N_t/6$.

An initial estimate of the ratio $KE=E/V$ allows to calculate E and deduce N .

The voltage drop is generally low in induction motors ($KE = 0.96 - 0.98$) and PMs motors ($KE = 0.9 - 0.95$).

The number of turns per coil is then $N_c = N/N_{ph}$ (consider the nearest integer value for N_c and recalculate N accordingly).

Finally, a slot will consist of $N_{cs} = N_c$ conductors for a single layer winding or $N_{cs} = 2N_c$ conductors for a double layer winding.

Furthermore, to keep the same emf E with the new N and k_w , we need to adjust the value of L in (3).

The current I in the conductor is calculated using (2).

After choosing the current density J , one can determine the conductor cross section area by: $S_c = I/J$.

If S_c is too high, it is necessary to subdivide the conductor into several parallel wires.

If the number of parallel paths of the winding is 'a', the emf at the machine terminals becomes E/a and the current a.I. However, it is preferable to use all the turns in series ($a=1$) because parallel paths could result in circulating currents (unbalance due to construction tolerances for instance).

3rd step: lamination dimensions (tooth, slot and yoke) and air-gap

According to the $B(H)$ curve of the laminations, one can choose the desired peak flux density B_{dmax} in the tooth to avoid excessive magnetic saturation. The tooth width l_d is determined by assuming that the air-gap flux under a tooth pitch (calculated using a constant B) circulates entirely in the tooth, which leads to :

$$l_d = \frac{2\sqrt{2}\pi R}{N_s} \frac{B}{B_{dmax}} \quad (4)$$

For a straight tooth, l_d is constant along the tooth height otherwise we consider the worst case where l_d is taken in the lowest tooth area (viz. at bore radius R for external stator).

To determine the tooth (or slot) height h_d , we first calculate the slot cross section area S_{slot} . We adopt a reasonable fill factor k_{fill} (0.3 to 0.5 for round wires and up to 0.7 for rectangular conductors). Hence, $S_{slot} = N_{cs} \cdot S_c / k_{fill}$.

For a trapezoidal slot shape (straight tooth) one can solve the equation below to compute h_d :

$$h_d^2 + \frac{N_s}{\pi} (\tau_u - l_d) h_d - \frac{N_s}{\pi} S_{slot} = 0 \quad \text{with} \quad \tau_u = \frac{2\pi R}{N_s} \quad (5)$$

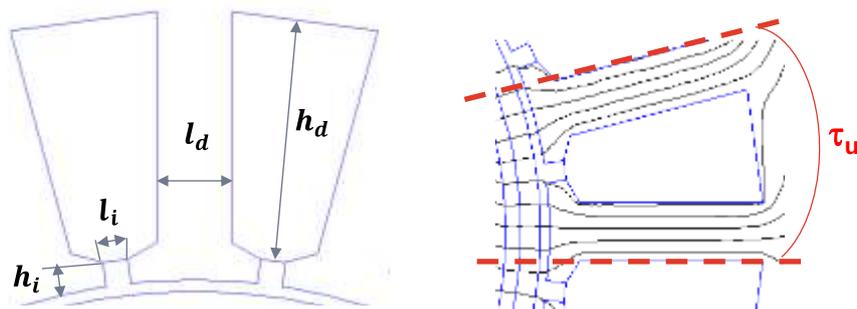


Figure 1 : 2D view of stator tooth pitch: (left) straight tooth and slot dimensions (right) flux lines under a tooth pitch

Moreover, we need to define the slot opening dimension (width l_i and height h_i) to have the final tooth height. This is a rather difficult problem which depends on the saturation of iron in the slot opening area. Finite element calculation is needed to determine a precise slot opening area shape.

For sizing the yoke, one selects the desired peak flux density B_{cmax} in the yoke. As shown in Figure 2, the yoke (or back-iron) height h_c is obtained by assuming that half the flux under a pole (calculated using a constant B) in the air-gap entirely closes through the back-iron. This leads to:

$$h_c = \frac{\pi R}{\sqrt{2p}} \frac{B}{B_{cmax}} \quad (6)$$

Finally, the outer stator radius is given by: $R_o = R + h_i + h_d + h_c$.

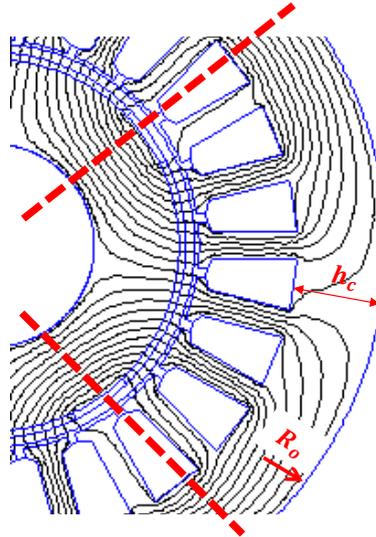


Figure 2 : 2D view of stator pole pitch showing back-iron dimensions and flux lines

4th step: losses and thermal criteria

To calculate the Joule losses generated in the stator winding, we need to estimate the end-winding wires length. Let's note L' the length of half a coil. The ratio L'/L usually varies between 1.2 - 2. This ratio decreases when increasing the pole-pair number p . Then, we can compute the phase resistance $R_{ph} = 2\rho_{cu}NL'/S_c$

The Joule losses are:

$$p_j = 3R_{ph}I^2 \quad (7)$$

The dimensions of the laminations being known, we can determine the weight of the teeth and yoke. Hence, the iron losses are roughly estimated using "Steimetz" like formulas:

$$p_{Fe} = k_p m \left(\frac{B_{peak}}{B_0} \right)^2 \left(\frac{F}{F_0} \right)^2 \quad (8)$$

Where:

m (kg): weight of the tooth or the yoke

k_p (W/kg): specific losses of the considered lamination at a base peak flux density B_0 and frequency F_0 (usually, $B_0 = 1$ or 1.5 T and $F_0 = 50$ or 60 Hz).

B_{peak} (T): peak flux density in the tooth (B_{dmax}) or the yoke (B_{cmax}).

At high frequency operation, this formula underestimates the iron losses of the machine, so correction factors are frequently used. It is then necessary to have precise data from lamination's manufacturer who gives the specific losses for various peak flux densities and frequencies.

The principal limitation when sizing any electrical machines is often a thermal one. Thus, an electromagnetic-thermal coupling is needed most of the time to achieve a robust machine sizing. A simplified thermal network is proposed in his paper.

2.2 Equivalent thermal circuit

Owing its good accuracy and low CPU time, a lumped parameter thermal model of the machine is built. Each part of the machine consists of an elementary block composed of a number of independent unidirectional lumped thermal circuits. A unidirectional thermal circuit is shown in Fig. 3.a, it is associated to a thermal conductivity λ in $\frac{W}{m.K}$, a power source P in W and a thermal capacity C in $\frac{J}{m.K}$. This lumped model contains 3 thermal resistances and allows the determination of the average temperature (and the maximal temperature if needed) [4]. A 3D lumped parameters thermal circuit is then obtained by connecting of the three unidirectional ones to the average temperature node of the elementary block, Fig.3.b. This cartesian representation of the unidirectional thermal circuit requires a linearization of the geometry of the different parts of the machines.

To have an equivalent thermal network of the overall machine, the thermal circuit of a given block is connected to its neighboring ones at their contact surface. By applying Kirchhoff's nodal law, one can write a linear system of equations whose solution allows the determination of the nodal temperatures.

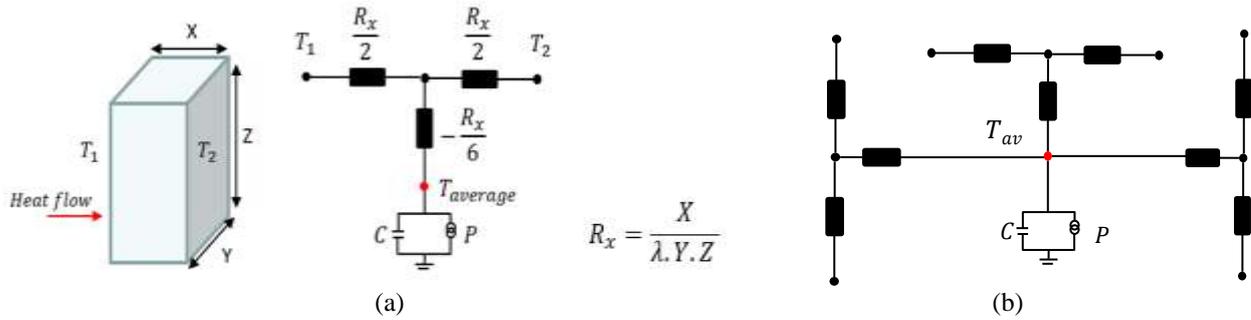


Figure 3. Elementary block and its equivalent 1D (a) and 3D (b) thermal circuits

3. Magneto-thermal sizing of a surface mounted PMSM

A surface mounted permanent magnet synchronous machine (PMSM) is considered owing it excellent performances which are required in aerospace applications. The electromagnetic computation of this machine can be easily done analytically under some assumptions.

3.1 Analytical computation of air-gap flux density

The procedure described above allows sizing a polyphase stator of an ac machine for a known magnetic loading B . In the case of a surface mounted PMSM, B is created by the rotor so it is important to determine the magnets and back-iron dimensions.

Figure 4 represents a cross section view of a surface mounted PMSM. We assume:

- a smooth airgap thus neglecting slotting effects
- infinite iron permeability
- radially magnetized magnets having a remanence B_{rem} and unity relative recoil permeability

By analytically solving the PDE arising from a 2D magnetostatic problem [5], the rms value B of the fundamental flux density computed on the stator bore radius R is:

$$B = \frac{4B_{rem}}{\pi\sqrt{2}} \frac{p}{p^2 - 1} \frac{(p-1)R_m^{2p} - (p+1)R_r^{2p} + 2R_m^{p-1}R_r^{p+1}}{R^{2p} - R_r^{2p}} \left(\frac{R}{R_m}\right)^{p-1} \quad \text{with } p > 1 \quad (9)$$

Equation (9) is then combined to equations (1)-(8) to achieve an analytical magnetic computation of the considered machine.

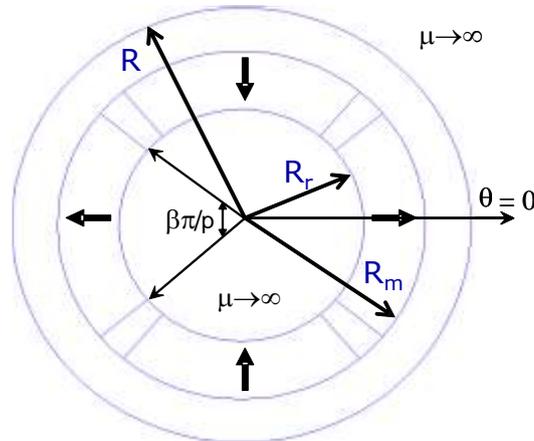


Figure 4. 2D cross section view of surface mounted PMSM used for the analytical modelling

3.2 Simplified thermal network of PMSM

A 1D thermal network is constructed by assuming that the heat flow in the machine is mainly radial. We only consider forced air cooling, so the generated heat in the machine is removed through its external stator surface (finned if any) and inner rotor surface in the case of a hollow shaft.

As shown in Figure 5, a slot pitch of the PMSM is considered and the active parts are decomposed into 5 conductive thermal blocks. For simplicity reasons, the actual cylindrical machine geometry is linearized so we can represent each block by the unidirectional thermal circuit of Figure 3. The convection coefficients h_e (airgap) and h_{ext} can be estimated by simple relations [4], [6].

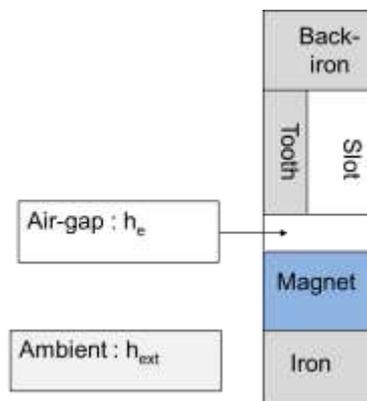


Figure 5. Tooth-pitch of PMSM used for thermal model

3.3 Material selection

Achieving high-power densities for aerospace applications in the frame of IMOTHEP project [7] requires the use of advanced materials as described hereafter.

Stator windings conductors

High frequency currents cause high ac copper losses so Litz conductors to reduce them are considered. The insulation class adopted here allows a maximum operating temperature of about 230 °C [8]. The slot copper fill factor is set to $k_{fill}=0.5$.

Magnetic circuit laminations

Cobalt-Iron based alloy “Vacoflux48” is considered for the stator laminations while 0.35mm “Vacodur S+” laminations are used in the rotor owing their high yield strength (~800 MPa) suitable for high-tip-speed rotor designs. The saturation flux density of Vacoflux 48 is close to 2.3 T.

Regarding the specific iron losses of Vacoflux48 which depend on the lamination thickness, the manufacturer [9] provides some values with respect to maximal flux density and frequency. We adopted here Vacoflux laminations with a thickness of 0.055mm for which we find some data regarding their specific losses in [10]. Specific losses are given for flux densities up to 2T and frequencies up to 2 kHz. For an iron flux density of 2T, we have approximated the specific losses $K_p [W/kg]$ vs. operating frequency $f [Hz]$ as a second order polynomial:

$$K_p = 3.2310^{-5} f^2 - 0.01187 f + 14.58 \quad (10)$$

In the simulation studies, we have used (10) obtained for 2T to estimate the iron losses of the machine.

Permanent Magnets

The RECOMA 35E SmCo magnet grade is considered for the rotor magnets [11]. The residual flux density is high enough (1.1 T) at an operating temperature of 250 °C. To reduce eddy current loss, these magnets can also be laminated in the axial direction and the magnet slices coated with a thin insulation [12]. Obviously, circumferential segmentation allows further reduction of the eddy current loss.

4. Development of an optimization tool under Calc spreadsheet application

LibreOffice (successor to OpenOffice) is a free office suite. It includes the spreadsheet application Calc which is a powerful and feature-rich tool to perform statistical and scientific calculations [13].

4.1 Computational tool

As shown in Figure 6, Calc's solver allows solving optimization problems in which the optimum value of a particular spreadsheet cell has to be calculated based on constraints provided in other cells. The DEPS Evolutionary Algorithm is used.

Both magnetic and thermal computation are performed in two separate sheets and coupled together in the solver interface.

For the thermal computation, we solve a (12x12) linear system using MINVERSE (for matrix inversion) and MMULT (for matrices multiplication) functions.

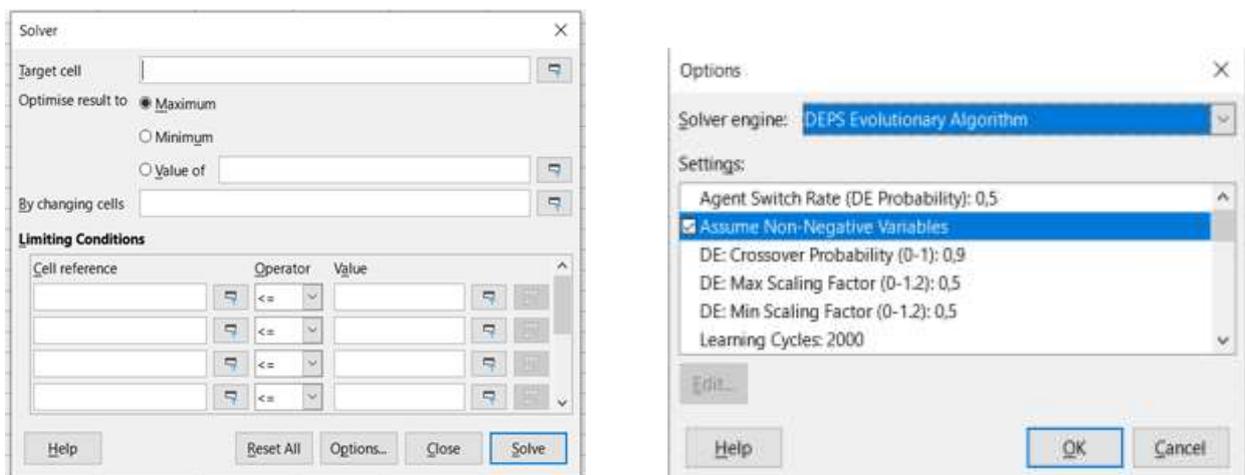


Figure 6. Calc solver window showing optimization features

4.2 Application example

A PMSM rated 500 kW-20000 rpm has been optimized. The following main constraints are considered:

- Air cooling with $h_{ext}=200$ W/m²K
- Airgap = 5 mm (includes PMs retaining sleeve)
- Winding heating is limited to 140 °C

The main results of the conducted optimization are shown in Table 1. The objective is to minimize the active mass. An active power density of about 9.77 kW/kg is obtained. The cpu time is about 30 s.

Table 1: Main results for a 500 kW PMSM

Power	Speed	Active diameter	Active length	Active weight	Efficiency
(kW)	(rpm)	(m)	(m)	(kg)	(%)
500	20.000	0.225	0.444	51.2	98.5

5. Conclusion

A powerful computational tool is developed Under LibreOffice Calc spreadsheet free application for the magneto-thermal design and optimization of PMSMs. Such a tool is well suited for collaborative works needed by the IMOTHEP project where different users can easily share and integrate new data.

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