

Model Parameters of Electric Motors for Desired Operating Conditions

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Abstract—Researchers dealing with electric motor control simulations need motor parameters for some desired operating conditions. Despite such an obvious need, no algorithm yielding motor parameters can be found even for the basic set of desired “voltage, output power, speed and efficiency” in the literature. A lot of electric motor design methods exist; but all give the physical design parameters for manufacturing such as numbers and dimensions of slots, magnets and turns. They are usually based on design requirements that only experienced people can understand and the mentioned basic demand set is not completely included among them. This article covers the deficiency of the algorithms giving all the model parameters required for the control simulations for dc servo, induction, and synchronous motors according to simple design requirements that an inexperienced researcher can easily understand. A transformer design algorithm is also included. The induction motor and salient-pole synchronous motor algorithms are the main contributions. The propositions can be used even if the demands are given for generator mode with some care. These algorithms may also be considered as another kind of design and they may help to reduce physical designs to lower-level steps according to simple design requirements.

Index Terms—AC motors, DC motors, induction motors, permanent magnet motors, algorithm design and analysis.

I. INTRODUCTION

Researchers dealing with control simulations of electric motors usually need a suitable set of model parameters yielding operation conditions about a desired region. Since an arbitrary set of parameters is probably not reasonable, they seek for a set of parameters belonging to a real motor or at least a verified model in simulations. However, what they have found may not meet their requirements closely. In addition, they may not notice an abnormality of the simulation results due to a possible programming error for the set of parameters and the working conditions. Therefore, they really need some design algorithms which simply give model parameters for control simulations from the demanded working range.

There are several works on designs of the dc motors [1-3], induction motors [4-7], synchronous motors of permanent magnet (PMSM) [8-10], or wound rotor (WRSM) [11-13], and both cylindrical [9], [12] and salient-pole [10-11], [13] rotor types. They explain good methods to find parameters for the physical implementations and manufacturing with some improvements; however, they do not give all the model parameters suitable for simulations, sometimes not even a winding resistance. A website serves some computational tools for permanent magnet (PM) motor designers [14]. It computes the physical parameters

including most of the parameters required for simple model simulations online. However, such tools ask the users several choices on which inexperienced users have no idea even though explanatory images are provided. In addition, the users cannot start directly with the basic demands for the operating conditions such as the power, voltage, speed and efficiency. Therefore, although there are praiseworthy tools and algorithms for the electric motor design, the existing ones in the literature are not suitable for researchers' quick needs to achieve simple model parameters for their demanded operating ranges. I do not want to elongate the reference list as a study explaining design methods to suit the researchers' control simulation purposes is apparently a serious lack in the literature.

This paper helps the researchers to generate their own motor parameters according to their desired operating conditions. The presented algorithms are for dc servo motors, induction motors and synchronous motors with the PM or wound rotors of the salient pole or cylindrical types, and for transformers. These are another kind of design algorithms based on quite different criteria than the physical design criteria [15-16], since the proposed ones are for simulation and educational purposes. The transformer algorithm is included to illustrate that this kind of designs may also give opinion about some manufacturing parameter values.

Even though most formulas are well-known, it should be emphasized against underestimation of the contributions that without following the specially organized steps and control assumptions, it is almost impossible to reach a set of parameters satisfying the requirements. My rigorous literature survey has resulted in no result to find an algorithm which gives a set of model parameters for control simulations of dc servo, induction, and synchronous motors satisfying the very basic set of demands for “the operating power, voltage, speed and efficiency”. As the induction motors and salient-pole synchronous motors need elaborated algorithms, they are the main contributions of this study.

With some care, as will be described, the algorithms can also be used when the requirements are given for the generator mode. As most models assume, the core losses, hysteresis, saturation and armature reaction are ignored here. The used models for ac motors are based on 3-phase ↔ 2-phase (dq) transformation with the amplitude equivalence of the phase variables as mostly used in the literature. The algorithms are based on some preferences since any control methods and special selections for arbitrary assumptions can be preferred in the design procedure to satisfy the demanded operating conditions. Most of the algorithm formulas are presented in tables for simplicity. Then the motor models

are given in the normal form of differential equations, which are ready to simulate with solver programs.

II. DC SERVO MOTOR DESIGN

A. Theory

Since all the time (t) derivatives become zero in the steady state, the electrical and mechanical equations [17] of the motor become

$$v_a = R_a i_a + L_a \dot{i}_a + K_b \omega \rightarrow \underbrace{v_a i_a}_{P_i} = \underbrace{R_a i_a^2}_{P_{Cu}} + \underbrace{K_b \omega i_a}_{P_m} \quad (1)$$

$$K_b i_a = T_L + B_f \omega + J_i \dot{\omega} \rightarrow \underbrace{K_b \omega i_a}_{P_m} = \underbrace{T_L \omega}_{P_o} + \underbrace{B_f \omega^2}_{P_f} \quad (2)$$

if multiplied by i_a and ω respectively, where the parameters R_a and L_a are the armature resistance and inductance, K_b is the back emf or torque constant, B_f is the friction constant and J_i is the inertia; and the variables v_a and i_a are the applied armature voltage and current, ω is the angular rotor speed in rad/s, T_L is the load torque, P_i and P_o are the input and output powers, P_m is the electromechanical power, P_{Cu} and P_f are the loss powers due to the armature resistance and friction respectively.

The model has 5 parameters but 2 of them, L_a and J_i , have no effect in the steady state. In addition, there are 2 independent variables, v_a and T_L . Therefore, we can have 5 demands for the steady state and 2 demands for the transient, which are the electrical and mechanical time constants determined with L_a and J_i respectively.

B. Algorithm with an Example

TABLE I. REQUIREMENTS FOR DC SERVO MOTOR DESIGN ALGORITHM WITH AN EXAMPLE DESIGN

Symbol	Quantity Name	Example Value
v_a	armature voltage	100 V
n	rotor speed	2000 rpm
P_o	output power	1500 W
η	efficiency	0.90
k_{ml}	mechanical to total loss rate (P_f/P_{loss})	0.5
τ_{elc}	electrical time constant	0.15 s
τ_{mec}	mechanical time constant	0.25 s

TABLE II. ALGORITHM FOR OPERATING VALUES OF DC SERVO MOTOR

Symbol with Formula	Value for Example Design
$P_i = P_o/\eta$	1667 W
$P_{loss} = P_i - P_o$	167 W
$P_f = k_{ml} P_{loss}$	83.3 W
$P_{Cu} = P_{loss} - P_f$	83.3 W
$P_m = P_o + P_f$	1583 W
$\omega = (\pi/30) \cdot n$	209 rad/s
$T_L = P_o/\omega$	7.16 Nm
$i_a = P_i/v_a$	16.7 A

The algorithm for the requirements in Table I is given in Table II-III, which are mostly based on the power components shown in (1)-(2), and it can be trivially modified for some other requirements. For example, in each of (v_a, i_a, P_i), (P_o, P_i, η), (T_L, P_o, n), (k_{ml}, P_{loss}, P_f), (R_a, L_a, τ_{elc}) and (B_f, J_i, τ_{mec}) triple, if any two others are determined, the third can be easily found from the simple relation among them.

If the core loss is not ignored, it must also be subtracted from P_{loss} when calculating P_{Cu} .

TABLE III. ALGORITHM FOR MODEL PARAMETERS OF DC SERVO MOTOR

Symbol with Formula	Value for Example Design
$R_a = P_{Cu}/i_a^2$	0.300 Ω
$L_a = R_a \tau_{elc}$	0.045 H
$K_b = P_m/(\omega i_a)$	0.454 Nm/A = 0.454 Vs/rad
$B_f = P_f/\omega^2$	0.00190 Nm·s/rad
$J_i = B_f \tau_{mec}$	0.000475 kg·m ²

The operating values in Table II and the parameters in Table III are verified exactly for the simulation of the following dc servo motor model [17]:

$$\begin{bmatrix} \dot{i}_a \\ \dot{\omega} \end{bmatrix} = \begin{bmatrix} -R_a/L_a & -K_b/L_a \\ K_b/J_i & -B_f/J_i \end{bmatrix} \begin{bmatrix} i_a \\ \omega \end{bmatrix} + \begin{bmatrix} 1/L_a \\ 0 \end{bmatrix} v_a - \begin{bmatrix} 0 \\ 1/J_i \end{bmatrix} T_L \quad (3)$$

III. INDUCTION MOTOR DESIGN

A. Theory

The field oriented control (FOC) with short circuited rotor will be considered, where the rotor flux linkage vector is aligned with d-axis. In addition, the minimum stator rms current will be preferred for the required torque.

Since all the derivatives become zero in the steady state, the electrical equations [18] of the stator and rotor become

$$v_s = R_s i_s + \frac{d\psi_s}{dt} + j\omega_g \psi_s \rightarrow v_s = R_s i_s + j\omega_g \psi_s \quad (4)$$

$$v_r = 0 = R_r i_r + \frac{d\psi_r}{dt} + j(\omega_g - \omega_r) \psi_r \rightarrow 0 = R_r i_r + j(\omega_g - \omega_r) \psi_r \quad (5)$$

where $v_s = v_{sd} + jv_{sq}$, $i_s = i_{sd} + ji_{sq}$, $\psi_s = \psi_{sd} + j\psi_{sq}$
 $= L_s i_s + M i_r$, $v_r = v_{rd} + jv_{rq}$, $i_r = i_{rd} + ji_{rq}$ and $\psi_r = \psi_{rd} + j\psi_{rq} = L_r i_r + M i_s$ are the complex stator voltage, current and flux linkage, and the rotor's respectively with respect to a reference frame rotating with an arbitrary electrical angular speed ω_g ; R_s, L_s, R_r and L_r are the stator resistance and inductance, and the rotor's respectively; M is the mutual inductance between the stator and rotor, and ω_r is the rotor's electrical angular speed.

With the choice of ω_g satisfying $\psi_{rq} = 0$ as in the FOC, as can be seen from (4)-(5) or [19], we get $\psi_{rd} = M i_{sd}$ in the steady state. Considering $\psi_r = (L_r/M)(\psi_s - \sigma L_s i_s)$, steady state values

$$\psi_{sq} = \sigma L_s i_{sq}, \quad \psi_{sd} = L_s i_{sd} \quad (6)$$

are achieved, where $\sigma = 1 - M^2/(L_s L_r)$ is the leakage coefficient. Then (4) becomes

$$\begin{bmatrix} v_{sd} \\ v_{sq} \end{bmatrix} = \begin{bmatrix} R_s & -\omega_g \sigma L_s \\ \omega_g L_s & R_s \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} \quad (7)$$

in the steady state. Multiplying both sides by $(3/2) \begin{bmatrix} i_{sd} & i_{sq} \end{bmatrix}$ from left gives

$$\underbrace{\frac{3}{2}(v_{sd} i_{sd} + v_{sq} i_{sq})}_{P_i} = \underbrace{\frac{3}{2} R_s (i_{sd}^2 + i_{sq}^2)}_{P_{CuSt}} + \frac{3}{2} \omega_g \underbrace{L_s (1 - \sigma)}_{M^2/L_r} i_{sd} i_{sq} \quad (8)$$

where P_i is the stator input power and P_{CuSt} is the stator's resistive loss. The choice of

$$\omega_g = \omega_r + \frac{i_{sq}}{\tau_r i_{sd}} \quad (9)$$

forces $\psi_{rq} \rightarrow 0$ quickly according to the rotor's electrical time constant $\tau_r = L_r/R_r$, and makes (8)

$$P_i = P_{CuSt} + \frac{3M^2}{2L_r} \omega_r i_{sd} i_{sq} + \frac{3}{2} R_r \frac{M^2}{L_r^2} i_{sq}^2 \quad (10)$$

Another arbitrary choice is the angle of i_s with respect to d-axis of the reference frame without imposing a requirement for ψ_{rd} . A reasonable choice for this angle is 45° , i.e., $i_{sd} = i_{sq}$, which maximize the electromechanical torque T_e for a certain level of $\sqrt{i_{sd}^2 + i_{sq}^2}$ since T_e becomes proportional to $i_{sd} i_{sq}$ due to the choice of $\psi_{rq} = 0$, which also makes $\omega_g = \omega_s$, the synchronous speed in electrical rad/s. In other words, this choice provides that a certain level of T_e is obtained with the minimum level of stator rms current. Then from (9) and (10),

$$P_i - P_{CuSt} = \frac{3M^2}{2L_r} \omega_g i_{sq}^2 = \frac{3M^2}{2L_r} \frac{\omega_r}{1-s} i_{sq}^2 \quad (11)$$

where s is the slip. As can be seen from the single-phase equivalent circuit of the induction motor without the core loss in the steady state,

$$P_i - P_{CuSt} = \frac{P_{CuRot}}{s} \quad (12)$$

and according to (9), the choice $i_{sd} = i_{sq}$ occurs if

$$\tau_r = \frac{1-s}{s\omega_r} \quad (13)$$

As equating the right hand side of (11) to that of (12) and using (13), we find another relation for parameters from the operating values:

$$\frac{M^2}{L_r} = \frac{2\tau_r P_{CuRot}}{3i_{sq}^2} \quad (14)$$

In the induction motor design algorithms, the stator power factor $\cos\phi_1$ should not be a design criteria since it is equal to $\cos 45^\circ$ lagging for an idealized induction motor [20], where the leakage flux and stator resistance are zero, if the minimum stator rms current is applied for the required torque and about $\cos 45^\circ$ in most other cases. The reason is, as seen from (6), since $\psi_{sq}/\psi_{sd} = \sigma \approx 0$, ψ_s is nearly aligned with the d-axis, v_s is about 90° ahead of it, that is about 45° ahead of i_s when $i_{sd} = i_{sq}$. The exact value of $\cos\phi_1$ is difficult to determine directly but we can simply do it in two stages instead. First, the parameters are calculated with an arbitrary $\cos\phi_1$ value such as 0.7. According to the design criteria given in the next subsection, the stator current is inverse proportional to $\cos\phi_1$, then (M^2/L_r) is proportional to $\cos^2\phi_1$ by (14) and so are

$$R_s = \frac{2P_{CuSt}}{3(i_{sd}^2 + i_{sq}^2)} \text{ and } L_s = \frac{M^2}{(1-\sigma)L_r}. \text{ Therefore, the stator}$$

voltage calculated from (7) is proportional to $\cos\phi_1$. By the first stage with the arbitrary $\cos\hat{\phi}_1$ value, (7) will not give the demanded stator voltage probably; however, the correct $\cos\phi_1$ value can be then found using the proportionality,

and some of the parameters are again calculated accordingly.

B. Algorithm with an Example

To satisfy the requirements in Table IV, the calculations in Table V are done first, where the same symbols have the same meanings as defined in Section II.

TABLE IV. REQUIREMENTS FOR INDUCTION MOTOR DESIGN WITH AN EXAMPLE

Symbol	Quantity Name	Example Value
V_{s1}^{rms}	stator rms voltage per phase	220 V
P_o	output power	4000 W
n_r	speed	1400 rpm
η	efficiency	0.80
n_{pp}	number of pole pair	2
k_{CuSt}	stator's share in copper loss (P_{CuSt}/P_{Cu})	0.5
σ	leakage coefficient	0.002
N_{St}/N_{Rot}	stator/rotor turn ratio	3
k_{ml}	mechanical to total loss rate (P_f/P_{loss})	0.25
τ_{mec}	mechanical time constant	0.25 s

TABLE V. DIRECT CALCULATIONS FOR INDUCTION MOTOR DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
$P_i = P_o/\eta$	5000 W
$P_{loss} = P_i - P_o$	1000 W
$P_f = k_{ml}P_{loss}$	250 W
$P_{Cu} = P_{loss} - P_f$	750 W
$P_{CuSt} = k_{CuSt}P_{Cu}$	375 W
$P_{CuRot} = P_{Cu} - P_{CuSt}$	375 W
$\omega_{mec} = (\pi/30) \cdot n_r$	146.6 rad/s
$T_L = P_o/\omega_{mec}$	27.3 Nm
$\omega_r = n_{pp}\omega_{mec}$	293 rad/s
$s = P_{CuRot}/(P_i - P_{CuSt})$	0.0811
$\omega_s = \omega_g = \omega_r/(1-s)$	319 rad/s
$\tau_r = (1-s)/(s\omega_r)$	0.0387 s

TABLE VI. TEMPORARY VALUES FOR INDUCTION MOTOR DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
$\cos\hat{\phi}_1$	0.7
$\hat{i}_{sd} = \hat{i}_{sq} = \hat{i}_{s1}^{rms} = \frac{P_i}{3 \cdot V_{s1}^{rms} \cdot \cos\hat{\phi}_1}$	10.82 A
$\hat{R}_s = \frac{P_{CuSt}}{3(\hat{i}_{s1}^{rms})^2}$	1.067 Ω
(\hat{M}^2/\hat{L}_r) from (14)	0.0825 H
$\hat{L}_s = \frac{(\hat{M}^2/\hat{L}_r)}{1-\sigma}$	0.0827 H
\hat{v}_{sd} from (7)	10.98 V
\hat{v}_{sq} from (7)	297.02 V
$\hat{V}_{s1}^{rms} = \sqrt{\frac{\hat{v}_{sd}^2 + \hat{v}_{sq}^2}{2}}$	210.17 V

Next, 2-stage calculations are done. In the first stage, temporary values, which are denoted by capped symbols, are found with an arbitrary $\cos\hat{\phi}_1$ (0.7 for the example), as

in Table VI. In the second stage, some operating values and parameters are calculated as in Table VII satisfying the demands exactly. Some additional operating values can also be calculated as shown in Table VIII.

TABLE VII. FINAL CALCULATIONS OF PARAMETERS FOR INDUCTION MOTOR DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
$\cos \varphi_1 = \left(\frac{V_{s1}^{rms}}{\hat{V}_{s1}^{rms}} \right) \cdot \cos \hat{\varphi}_1$	0.7327
$i_{sd} = i_{sq} = I_{s1}^{rms} = \frac{P_i}{3 \cdot V_{s1}^{rms} \cdot \cos \varphi_1}$	10.34 A
$R_s = \frac{P_{CuSt}}{3(I_{s1}^{rms})^2}$	1.169 Ω
(M^2/L_r) from (14)	0.09040 H
$L_s = \frac{(M^2/L_r)}{1-\sigma}$	0.09058 H
$L_r = L_s / (N_{St}/N_{Rot})^2$	0.01006 H
$R_r = L_r / \tau_r$	0.2604 Ω
$M = \sqrt{(M^2/L_r)L_r}$	0.03016 H
$B_f = P_f / \omega_{mec}^2$	0.0116 Nm·s/rad
$J_i = B_f \tau_{mec}$	2.91×10^{-3} kg·m ²

TABLE VIII. CALCULATIONS OF SOME OPERATING VALUES FOR INDUCTION MOTOR DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
v_{sd} from (7)	11.49 V
v_{sq} from (7)	310.91 V
$\Psi_{sd} = L_s i_{sd}$	0.9365 Vs
$\Psi_{sq} = \sigma L_s i_{sq}$	0.00187 Vs
$f_s = \omega_s / (2\pi)$ = stator frequency	50.78 Hz
$i_{rd} = (\Psi_{sd} - L_s i_{sd}) / M$	0 A
$i_{rq} = (\Psi_{sq} - L_s i_{sq}) / M$	-30.99 A
$\Psi_{rd} = (\Psi_{sd} - \sigma L_s i_{sd}) L_r / M$	0.3119 Vs
$\Psi_{rq} = (\Psi_{sq} - \sigma L_s i_{sq}) L_r / M$	0 Vs

C. Model to Simulate

The parameter set can be used with any form of model; for example, arranging the model differential equations in [18] into the normal form, (15) is obtained in the synchronous reference frame with short-circuited rotor, and with the stator current and rotor flux linkage as the electrical state variables.

$$\frac{d}{dt} \begin{bmatrix} i_{sd} \\ i_{sq} \\ \Psi_{rd} \\ \Psi_{rq} \\ \omega_r \end{bmatrix} = \begin{bmatrix} -\frac{L_r R_s + M^2/\tau_r}{\sigma L_s L_r} i_{sd} + \omega_g i_{sq} + \frac{M}{\tau_r \sigma L_s L_r} \Psi_{rd} + \frac{M}{\sigma L_s L_r} \omega_r \Psi_{rq} + \frac{1}{\sigma L_s} v_{sd} \\ -\frac{L_r R_s + M^2/\tau_r}{\sigma L_s L_r} i_{sq} - \omega_g i_{sd} + \frac{M}{\tau_r \sigma L_s L_r} \Psi_{rq} - \frac{M}{\sigma L_s L_r} \omega_r \Psi_{rd} + \frac{1}{\sigma L_s} v_{sq} \\ \frac{M}{\tau_r} i_{sd} - \frac{1}{\tau_r} \Psi_{rd} + (\omega_g - \omega_r) \Psi_{rq} \\ \frac{M}{\tau_r} i_{sq} - \frac{1}{\tau_r} \Psi_{rq} - (\omega_g - \omega_r) \Psi_{rd} \\ \frac{3}{2} \frac{n_{pp}^2 M}{J_i L_r} (\Psi_{rd} i_{sq} - \Psi_{rq} i_{sd}) - \frac{B_f}{J_i} \omega_r - \frac{n_{pp}}{J_i} T_L \end{bmatrix} \quad (15)$$

Alternatively, a doubly fed induction motor model as in (16) can also be used with the parameters found by this algorithm; however, the operating values of the algorithm are for zero rotor voltages v_{rd} , v_{rq} . Equation (16) is obtained arranging the model differential equations in [21] into the normal form.

$$\frac{d}{dt} \begin{bmatrix} i_{sd} \\ i_{sq} \\ i_{rd} \\ i_{rq} \\ \omega_r \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{\sigma L_s} i_{sd} + \left(\omega_g + \frac{M^2}{\sigma L_s L_r} \omega_r \right) i_{sq} + \frac{MR_r}{\sigma L_s L_r} i_{rd} + \frac{M}{\sigma L_s} \omega_r i_{rq} + \frac{1}{\sigma L_s} v_{sd} - \frac{M}{\sigma L_s L_s} v_{rd} \\ -\left(\omega_g + \frac{M^2}{\sigma L_s L_r} \omega_r \right) i_{sd} - \frac{R_s}{\sigma L_s} i_{sq} - \frac{M}{\sigma L_s} \omega_r i_{rd} + \frac{MR_r}{\sigma L_s L_r} i_{rq} + \frac{1}{\sigma L_s} v_{sq} - \frac{M}{\sigma L_s L_s} v_{rq} \\ \frac{MR_s}{\sigma L_s L_r} i_{sd} - \frac{M}{\sigma L_s} \omega_r i_{sq} - \frac{R_r}{\sigma L_r} i_{rd} + \left(\omega_g - \frac{1}{\sigma} \omega_r \right) i_{rq} - \frac{M}{\sigma L_s} v_{sd} + \frac{1}{\sigma L_r} v_{rd} \\ \frac{M}{\sigma L_r} \omega_r i_{sd} + \frac{MR_s}{\sigma L_s L_r} i_{sq} - \left(\omega_g - \frac{1}{\sigma} \omega_r \right) i_{rd} - \frac{R_r}{\sigma L_r} i_{rq} - \frac{M}{\sigma L_s} v_{sq} + \frac{1}{\sigma L_r} v_{rq} \\ \frac{3}{2} \frac{n_{pp}^2 M}{J_i} (\Psi_{rd} i_{sq} - \Psi_{rq} i_{sd}) - \frac{B_f}{J_i} \omega_r - \frac{n_{pp}}{J_i} T_L \end{bmatrix} \quad (16)$$

D. Equivalent Circuit and Additional Values:

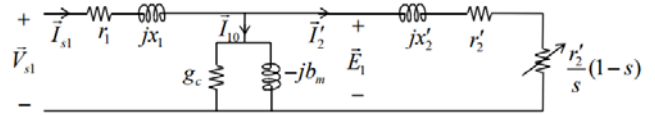


Figure 1. Induction motor equivalent circuit, which is reduced to single phase and referred to the stator

The parameters can also be converted to the parameters of the single-phase equivalent circuit (Fig. 1) as will be shown in Table IX.

TABLE IX. EQUIVALENT CIRCUIT PARAMETERS FOR INDUCTION MOTOR

Symbol with Formula	Value for Example Design
$r_1 = R_s$	1.169 Ω
$r_2 = R_r$	0.2604 Ω
$k = \sqrt{1-\sigma}$ couplage coefficient	0.9990
$L_m = (N_{St}/N_{Rot})M$ magnetizing inductance	0.09049 H
$L_{1l} = (1-k)L_s$ stator leakage inductance	90.63 μ H
$L_{2l} = (1-k)L_r$ rotor leakage inductance	10.07 μ H
$x_1 = \omega_s L_{1l}$ stator leakage reactance	28.92 m Ω
$x_2 = \omega_s L_{2l}$ rotor leakage reactance at f_s	3.213 m Ω
$b_m = 1/(\omega_s L_m)$ magnetizing susceptance	34.63 mS
g_c (ignored) core losses conductivity	0 S
$r_2' = (N_{St}/N_{Rot})^2 \cdot r_2$	2.343 Ω
$x_2' = (N_{St}/N_{Rot})^2 \cdot x_2$	28.92 m Ω

All these parameters and operating conditions are exactly verified by both the simulation of (15) and the equivalent circuit computations.

IV. PMSM DESIGN

A. Theory

To develop a design algorithm for the PMSM, the stator flux orientation will be considered, where the stator flux linkage vector's component sourced by the permanent magnet (Φ_{PM}) is aligned with d-axis. In addition, the minimum stator rms current will be preferred for the required torque.

Stator equation [22] is similar to that of the induction motor with ω_r replaced for ω_g . Since all the derivatives become zero in the steady state, the stator equation becomes

$$v_s = R_s i_s + \frac{d\Psi_s}{dt} + j\omega_r \Psi_s \rightarrow v_s = R_s i_s + j\omega_r \Psi_s, \quad (17)$$

where

$$\Psi_s = \Psi_{sd} + j\Psi_{sq} = (L_{sd} i_{sd} + \Phi_{PM}) + jL_{sq} i_{sq}, \quad (18)$$

L_{sd} and L_{sq} are d- and q-axis synchronous inductances, which

are different for salient-pole machines, and similar symbols have similar meanings as those of the induction motors. Then at the equilibrium,

$$\begin{bmatrix} v_{sd} \\ v_{sq} \end{bmatrix} = \begin{bmatrix} R_s & -\omega_r L_{sq} \\ \omega_r L_{sd} & R_s \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \begin{bmatrix} 0 \\ \omega_r \Phi_{PM} \end{bmatrix}. \quad (19)$$

Multiplying both sides by $(3/2)[i_{sd} \ i_{sq}]$ from left gives the input power:

$$P_i = \frac{3}{2}(v_{sd}i_{sd} + v_{sq}i_{sq}) = \frac{3}{2}R_s(i_{sd}^2 + i_{sq}^2) + \frac{3}{2}\omega_r(L_{sd} - L_{sq})i_{sd}i_{sq} + \frac{3}{2}\omega_r\Phi_{PM}i_{sq}, \quad (20)$$

where the first term in the right hand side is P_{Cu} . Since the electromechanical torque is

$$T_e = \frac{3}{2}n_{pp}(L_{sd} - L_{sq})i_{sd}i_{sq} + \frac{3}{2}n_{pp}\Phi_{PM}i_{sq} \quad (21)$$

and $\omega_{mec} = \omega_r/n_{pp}$, sum of the other two terms in the right hand side of (20) equals the electromechanical power ($P_m = T_e\omega_{mec} = P_o + P_f$). In order to obtain maximum T_e from a certain level of stator rms current $I_{s1}^{rms} = \sqrt{(i_{sd}^2 + i_{sq}^2)/2}$,

substituting $i_{sq} = \sqrt{2(I_{s1}^{rms})^2 - i_{sd}^2}$ and equating the derivative of T_e with respect to i_{sd} to zero, we need to solve

$$\frac{\partial T_e}{\partial i_{sd}} = \frac{3}{2}n_{pp}(L_{sd} - L_{sq})\sqrt{2(I_{s1}^{rms})^2 - i_{sd}^2} - \frac{3}{2}n_{pp}\frac{\Phi_{PM}i_{sd}}{\sqrt{2(I_{s1}^{rms})^2 - i_{sd}^2}} = 0 \quad (22)$$

for i_{sd} . Using $k_{TPM} = \frac{\Phi_{PM}}{(L_{sd} - L_{sq})i_{sd} + \Phi_{PM}}$ which is defined as the ratio of the torque produced due to the permanent magnet to total T_e , and $\frac{\Phi_{PM}}{(L_{sd} - L_{sq})i_{sd}} = \frac{k_{TPM}}{1 - k_{TPM}}$ in (22),

$$i_{sd}^2 = \frac{2}{2 + \frac{\Phi_{PM}}{(L_{sd} - L_{sq})i_{sd}}}(I_{s1}^{rms})^2 = \frac{2 - 2k_{TPM}}{2 - k_{TPM}}(I_{s1}^{rms})^2 \quad (23)$$

$$i_{sq}^2 = \frac{2}{2 - k_{TPM}}(I_{s1}^{rms})^2 \quad (24)$$

Since Φ_{PM} is a positive parameter,

$$i_{sq} = I_{s1}^{rms} \cdot \sqrt{\frac{2}{2 - k_{TPM}}} \cdot \text{sign}(k_{TPM}T_e) \quad (25)$$

$$i_{sd} = I_{s1}^{rms} \cdot \sqrt{\frac{2 - 2k_{TPM}}{2 - k_{TPM}}} \cdot \text{sign}\left(\frac{(1 - k_{TPM})T_e}{i_{sq}}\right) \quad (26)$$

The algorithm to decide the PMSM parameters according to the desired operating conditions is straightforward for the cylindrical rotor type since $k_{TPM} = 1$ as $L_{sd} = L_{sq}$. Equating $2(V_{s1}^{rms})^2 = (v_{sd}^2 + v_{sq}^2)$ by using (19) gives

$$L_{sd} = L_{sq} = \sqrt{\left(\frac{V_{s1}^{rms}}{\omega_r I_{s1}^{rms}}\right)^2 - \left(\frac{R_s}{\omega_r}\right)^2} - \frac{1}{2}\left(\frac{\Phi_{PM}}{I_{s1}^{rms}}\right)^2 - \frac{\sqrt{2}R_s\Phi_{PM}}{\omega_r I_{s1}^{rms}} \quad (27)$$

for the cylindrical rotor PMSM. However, a nonlinear equation of k_{TPM} with very complicated coefficients should be solved for the salient-pole type. Instead of analytically solving that complicated equation, a loop algorithm is recommended in order to determine k_{TPM} . The loop

algorithm can be the Newton-Rampson's method but the derivative is replaced by the numerical approximation calculated from the last two iterations. Then the other parameters can be determined.

B. Algorithm with an example

To satisfy the requirements in Table X, the calculations in Table XI are done first, where the same symbols have the same meanings as defined in previous sections.

Then, if the rotor is cylindrical, i.e. $k_{dq} = 1$, the other parameters and some operating values are found as in Table XII.

TABLE X. REQUIREMENTS FOR PMSM DESIGN ALGORITHM WITH AN EXAMPLE DESIGN

Symbol	Quantity Name	Example Value
V_{s1}^{rms}	stator rms voltage per phase	220 V
P_o	output power	4000 W
n_r	speed	1400 rpm
η	efficiency	0.80
n_{pp}	number of pole pair	2
k_{dq}	saliency ratio (L_{sd}/L_{sq}) (for salient-pole)	5/3
$\cos\phi_1$	power factor	0.8
k_{ml}	mechanical to total loss rate (P_f/P_{loss})	0.25
τ_{mec}	mechanical time constant	0.25 s

TABLE XI. DIRECT CALCULATIONS FOR PMSM DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
$P_i = P_o/\eta$	5000 W
$P_{loss} = P_i - P_o$	1000 W
$P_f = k_{ml}P_{loss}$	250 W
$P_{Cu} = P_{loss} - P_f$	750 W
$I_{s1}^{rms} = P_i/(3V_{s1}^{rms}\cos\phi_1)$	9.47 A
$\omega_{mec} = (\pi/30)\cdot n_r$	146.6 rad/s
$\omega_r = n_{pp}\omega_{mec}$	293 rad/s
$T_L = P_o/\omega_{mec}$	27.3 Nm
$B_f = P_f/\omega_{mec}^2$	0.0116 Nm·s/rad
$J_i = B_f\tau_{mec}$	2.91×10^{-3} kg·m ²
$R_s = P_{Cu}/(3(I_{s1}^{rms})^2)$	2.7878 Ω
$T_e = (P_i - P_{Cu})/\omega_{mec}$	28.99 Nm

TABLE XII. CALCULATIONS OF SOME PARAMETERS AND OPERATING VALUES FOR CYLINDRICAL ROTOR PMSM DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
$k_{TPM} = 1$	1
$i_{sd} = 0$	0 A
$i_{sq} = \sqrt{2} \cdot I_{s1}^{rms}$	13.39 A
$\Phi_{PM} = 2T_e/(3n_{pp}i_{sq})$	0.7215 Vs
L_{sd} from (27)	47.54 mH
$L_{sq} = L_{sd}$	47.54 mH
v_{sd} from (19)	-186.7 V
v_{sq} from (19)	248.9 V

For the salient-pole motors ($k_{dq} \neq 1$), the following algorithm with a loop is proposed:

- Step 1: Assign a stopping ϵ value for $|e_v|$, the absolute error for v_{s1}^{rms} demand, for example $\epsilon = 10^{-6}$ V.
- Step 2: Assign a limit for $|\Delta k_{TPM}|$, absolute change of k_{TPM} in a step, for example $\Delta k_{max} = 0.02$.

Step 3: Initiate following arbitrarily about the example values $k_{TPM} = 0.5$, $\Delta k_{TPM} = 0.0001$, $e_v = 0.3V$, $e_v^{old} = 0.5V$

Step 4: while $|e_v| > \epsilon$,

Step 4.a: $\Delta k_{TPM} = -\Delta k_{TPM} \frac{e_v}{e_v - e_v^{old}}$

Step 4.b: If $|\Delta k_{TPM}| > \Delta k_{max}$, then $\Delta k_{TPM} = \text{sign}(\Delta k_{TPM}) \cdot \Delta k_{max}$

Step 4.c: $k_{TPM} = k_{TPM} + \Delta k_{TPM}$, $e_v^{old} = e_v$

Step 4.d: Calculate i_{sq} and i_{sd} from (25) and (26)

Step 4.e: $\Phi_{PM} = \frac{2k_{TPM}T_e}{3n_{pp}i_{sq}}$

Step 4.f: $L_{sq} = \frac{2(1-k_{TPM})T_e}{3(k_{dq}-1)n_{pp}i_{sd}i_{sq}}$, $L_{sd} = k_{dq}L_{sq}$

Step 4.g: Calculate v_{sd} and v_{sq} from (19)

Step 4.h: $e_v = V_{s1}^{rms} - \sqrt{(v_{sd}^2 + v_{sq}^2)}/2$

end while

This algorithm yields the parameters and operating values in Table XIII for the example. They are verified exactly with simulations

TABLE XIII. CALCULATIONS OF SOME PARAMETERS AND OPERATING VALUES FOR SALIENT-POLE ROTOR PMSM DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
k_{TPM}	0.890
i_{sd}	4.21 A
i_{sq}	12.71 A
Φ_{PM}	0.6767 Vs
L_{sd}	49.50 mH
L_{sq}	29.70 mH
v_{sd}	-98.97 V
v_{sq}	294.97 V

C. Model to simulate

The parameter set can be used with any form of the model, for example, (28), which is in the synchronous reference frame with the stator current and rotor speed as the electrical state variables. This is obtained arranging the model differential equations in [22] into the normal form.

$$\frac{d}{dt} \begin{bmatrix} i_{sd} \\ i_{sq} \\ \omega_r \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{L_{sd}}i_{sd} + \frac{L_{sq}}{L_{sd}}\omega_r i_{sq} + \frac{1}{L_{sd}}v_{sd} \\ -\frac{R_s}{L_{sq}}i_{sq} - \frac{L_{sd}}{L_{sq}}\omega_r i_{sd} - \frac{\Phi_{PM}}{L_{sq}}\omega_r + \frac{1}{L_{sq}}v_{sq} \\ \frac{3n_{pp}^2(L_{sd}-L_{sq})}{2J_i}i_{sd}i_{sq} + \frac{3n_{pp}^2\Phi_{PM}}{2J_i}i_{sq} - \frac{B_f}{J_i}\omega_r - \frac{n_{pp}}{J_i}T_L \end{bmatrix} \quad (28)$$

V. WRSM DESIGN

A. Theory

To decide the WRSM parameters for certain operating values, the same method is included as the PMSM design replacing P_{Cu} and Φ_{PM} with P_{CuRot} and Mi_f respectively, where i_f is the rotor current and M is the mutual inductance between the stator and rotor. Similarly P_i in I_{s1}^{rms} and T_e formulas is replaced with just the stator's input power

$P_{iSt} = P_i - P_{CuRot}$. In addition, given any two of the desired values of v_f , i_f and $k_{rl} = P_{CuRot}/P_{loss}$, the third is found from their steady state relation, $v_f = R_f i_f$, where v_f and R_f are the voltage and resistance of the rotor. To determine the rotor inductance L_f , an additional requirement is assigned for a measure of the leakage between a stator phase and rotor windings with

$$\sigma_f = 1 - \frac{3M^2}{2L_{sd}L_f} \quad (29)$$

This measure is somewhat complicated than the usual leakage coefficient due to the saliency of the rotor but still complies $0 \leq \sigma_f \leq 1$ since L_{sd} is 3/2 times a stator phase's self inductance in the case of the best alignment with the rotor and no-leakage [23]. Then, we get

$$L_f = \frac{3M^2}{2(1-\sigma_f)L_{sd}} \quad (30)$$

B. Algorithm with an example

1) Requirements:

Without losing the generalization, just not to write the same steps here again as in the PMSM design, the same requirements are assumed with slight differences that while P_o , $P_{iSt} = P_i - P_{CuRot}$, P_{CuRot} and P_f are desired the same as before, $k_{rl} = 0.2$ is chosen, which means $P_i = 5250W$, $P_{loss} = 1250W$, $P_{CuRot} = 250W$, $k_{ml} = 0.2$ and $\eta = 0.7619$ are desired. Let the additional requirements be $v_f = 24V$ and $\sigma_f = 0.02$.

2) Calculations:

Now, all the other values in calculation parts given in PMSM section are found the same regarding Φ_{PM} as Mi_f . Then,

$$i_f = \frac{P_{CuRot}}{v_f} = \frac{250W}{24V} = 10.42A \quad (31)$$

$$R_f = \frac{v_f}{i_f} = \frac{24V}{10.42A} = 2.304\Omega \quad (32)$$

For the cylindrical rotor case ($k_{dq} = 1$),

$$M = \frac{\Phi_{PM}}{i_f} = \frac{0.7215Vs}{10.42A} = 69.27mH \quad (33)$$

and by (30), $L_f = 154.5mH$.

For the salient-pole case with $k_{dq} = 5/3$.

$$M = \frac{\Phi_{PM}}{i_f} = \frac{0.6767Vs}{10.42A} = 64.97mH \quad (34)$$

and by (30), $L_f = 130.5mH$.

C. Model to simulate

The parameter set can be used with any form of the model, for example, the following model in the synchronous reference frame with the stator current and rotor speed as the electrical state variables.

$$\frac{d}{dt} \begin{bmatrix} i_{sd} \\ i_{sq} \\ \omega_r \end{bmatrix} = \begin{bmatrix} \frac{1}{1.5M^2 - L_{sd}L_f} (R_s L_f i_{sd} - L_{sq} L_f \omega_r i_{sq} - R_f M i_f - L_f v_{sd} + M v_f) \\ -\frac{L_{sd}}{L_{sq}} \omega_r i_{sd} - \frac{R_s}{L_{sq}} i_{sq} - \frac{M}{L_{sq}} \omega_r i_f + \frac{1}{L_{sq}} v_{sq} \\ \frac{1}{1.5M^2 - L_{sd}L_f} (-1.5R_s M i_{sd} + 1.5L_{sq} M \omega_r i_{sq} + R_f L_{sd} i_f + 1.5M v_{sd} - L_{sd} v_f) \\ \frac{3n_{pp}^2(L_{sd}-L_{sq})}{2J_i} i_{sd} i_{sq} + \frac{3n_{pp}^2 M}{2J_i} i_{sq} i_f - \frac{B_f}{J_i} \omega_r - \frac{n_{pp}}{J_i} T_L \end{bmatrix} \quad (35)$$

This is the normal form of the model differential equations in [24], where the flux linkage variables are

$$\left. \begin{aligned} \Psi_{sd} &= L_{sd}i_{sd} + Mi_f \\ \Psi_{sq} &= L_{sq}i_{sq} \\ \Psi_f &= L_f i_f + \frac{3}{2}Mi_{sd} \end{aligned} \right\} \quad (36)$$

and Ψ_f is the rotor winding's flux linkage.

VI. MODIFICATIONS FOR GENERATORS

In the generator mode, the input power and the motor's shaft output power become negative with their definitions according to the motor mode. Although the negative of the shaft output power with the motor mode definition is the generator's shaft input power, the negative of the input power with the motor mode definition is not the generator's output power if there is an excitation current applied. Therefore, when using the presented algorithms for the generator mode demands, negative of the generator's desired output power is added to the excitation power and used as the input power in the algorithms. For a wound rotor synchronous generator example, let the design requirements be 1300W of the gross shaft input power, 1000W of the net electrical stator output power and 100W of the excitation (rotor) input power. Then any two of the input power $P_i = -900W$, the output power $P_o = -1300W$ and the efficiency $(-1300)/(-900) = 1.444$ are used as the design requirements in the algorithms even though the generator's efficiency is $900/1300 = 0.692$ actually. For doubly-fed induction machines, the power input from the rotor is also considered as if the excitation power, which also becomes negative if the positive power is extracted from the rotor's electrical terminals.

Induction machine design according to the generator mode demands requires two further actions. First, initial value $\cos\hat{\phi}_1$ must be taken a negative value such as -0.7 . Second, not to get a negative value from (13) with a negative slip, τ_r must be taken negative of it, which means $i_{sd} = -i_{sq}$ is applied.

VII. TRANSFORMER DESIGN

TABLE XIV. REQUIREMENTS FOR TRANSFORMER DESIGN ALGORITHM WITH AN EXAMPLE DESIGN

Symbol	Quantity Name	Example Value
V_1	primary rms voltage per phase	231 V
V_2	secondary rms voltage per phase	560 V
f	frequency	50 Hz
S_2	secondary apparent power per phase	12 kVA
k_{Cu}	copper loss to total loss rate (P_{Cu}/P_{loss}) at rated resistive load	0.5
k_{Cu2}	secondary copper loss to total copper loss rate (P_{Cu2}/P_{Cu}) at rated resistive load	0.4
η	efficiency at rated resistive load	0.92
k	coupling coefficient	0.99
α_{EV2}	leading angle of \vec{E}_2 with respect to \vec{V}_2	5°

An algorithm for transformer parameters based on the requirements in Table XIV is presented in Table XV for educational needs. For example, to assess the student's capability on vector algebra at an exam, the lecturer may want α_{EV2} angle not to be negligible. Explanations are not

given for most of the formulas and symbols as they are well-known. Their organization is the algorithm.

TABLE XV. CALCULATIONS FOR TRANSFORMER DESIGN ALGORITHM

Symbol with Formula	Value for Example Design
$\omega = 2\pi f$	314 rad/s
$I_2 = S_2 / V_2$	21.43 A
$P_1 = S_2 / \eta$	13.043 kW
$P_{loss} = P_1 - S_2$	1043 W
$P_{Cu} = k_{Cu}P_{loss}$	522 W
$P_{Cu2} = k_{Cu2}P_{Cu}$	209 W
$P_{Cu1} = P_{Cu} - P_{Cu2}$	313 W
$r_2 = P_{Cu2} / I_2^2$	0.4545 Ω
$P_{Fe} = P_{loss} - P_{Cu}$	522 W
$x_2 = (V_2 / I_2 + r_2) \tan(\alpha_{EV2})$	2.326 Ω
$E_2 = \sqrt{(V_2 + r_2 I_2)^2 + (x_2 I_2)^2}$	572 V
$g'_c = P_{Fe} / E_2^2$	1.595 mS
$x'_m = x_2 \cdot k / (1 - k)$	230.3 Ω
$b'_m = 1 / x'_m$	4.342 mS
$\vec{I}'_{10} = (g'_c - jb'_m)(E_2 \angle \alpha_{EV2})$	(1.1252 - j 2.3945) A
$\vec{I}'_1 = I_2 + \vec{I}'_{10}$	(22.5538 - j 2.3945) A
$I'_1 = \vec{I}'_1 $	22.68 A
$r'_1 = P_{Cu1} / I_1'^2$	0.6086 Ω
$x'_1 = x'_m (1 - k)$	2.3029 Ω
$\vec{V}'_1 = (r'_1 + jx'_1)\vec{I}'_1 + (E_2 \angle \alpha_{EV2})$	(588.98 - j 100.33) V
$V'_1 = \vec{V}'_1 $	597.46 V
$a = V_1 / V'_1$	0.387
$r_1 = a^2 r'_1$	0.0910 Ω
$x_1 = a^2 x'_1$	0.3442 Ω
$g_c = g'_c / a^2$	10.67 mS
$b_m = b'_m / a^2$	29.05 mS
$x_m = a^2 x'_m$	34.425 Ω
$I_1 = I'_1 / a$	58.66 A
$E_1 = aE_2$	221 V
$L_{1l} = x_1 / \omega$	1.096 mH
$L_{2l} = x_2 / \omega$	7.404 mH
$L_m = x_m / \omega$	109.6 mH
$M = L_m / a$	283.41 mH
$L_1 = L_m + L_{1l}$	110.67 mH
$L_2 = (M/a) + L_{2l}$	740.43 mH
$\sigma = 1 - M^2 / (L_1 L_2)$	0.0198

The presented algorithms in this paper may help manufacturing purposes designs. For the transformer design example, assuming $\mu_r = 900$, $h^2/A = 133$, magnetic flux density $B = 1.2$ T, stack factor = 0.8, resistivity of copper wire $\rho_{Cu} = 1.8 \times 10^{-8}$ Ωm , we find the main manufacturing parameters approximately:

- number of turns for primary winding

$$N_1 = \frac{\sqrt{2} \times 0.8 \times 133 \pi f B (L_m + L_{1l})^2}{(\mu_r \mu_0)^{2/3} E_1^{1/3}} = 107$$

- number of turns for secondary winding $N_2 = N_1/a = 277$

- length of magnetic flux path $h = \frac{0.8 \times (L_m + L_{II})}{\mu_r \mu_0 N_1^2} = 113 \text{ cm}$

- cross sectional area of core $A = h^2/133 = 97 \text{ cm}^2$

- length of primary wire $l_{w1} = 4N_1 h/\sqrt{133} = 42.2 \text{ m}$

- cross sectional area of primary wire

$$A_{w1} = \rho_{Cu} \times l_{w1}/r_1 = 8.3 \text{ mm}^2$$

- length of secondary wire $l_{w2} = 4N_2 h/\sqrt{133} = 109.0 \text{ m}$

- cross sectional area of secondary wire

$$A_{w2} = \rho_{Cu} \times l_{w2}/r_2 = 4.3 \text{ mm}^2$$

These calculations are not precise since the coupling coefficient is assumed to be true and some length and area errors due to bending and isolations are ignored; however, they give a fairly close opinion for the physical design.

VIII. CONCLUSION

Easy-to-use formulas and algorithms have been presented to achieve basic model parameters for dc servo motors, induction motors, PMSMs, WRSMs and transformers. Design requirements are mostly the operating conditions. The other design requirements such as the turn ratio, time constants, leakage coefficients etc. are simple to decide for an inexperienced researcher. The resulting sets of model parameters satisfy the desired operating conditions exactly for the assumed models. The algorithms are also applicable to the generator mode demands.

Although the presented design algorithms do not yield most of the manufacturing parameters, they will be helpful to decide them too since the operating values for the demands are also found. Transformer example has been extended to that level to illustrate that possibility. Even if it is more difficult for motors, a quick opinion for the physical sizes can be inferred with the proposed algorithms.

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