Normalised hybrid flux weakening strategy for automotive asymmetrical dual three-phase IPMSMs

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Abstract—In this paper, a hybrid flux weakening control strategy based on a vector space decomposition (VSD) model is presented for asymmetrical dual three-phase interior permanent magnet synchronous machines (IPMSMs). This strategy integrates a voltage magnitude feedback loop, which provides robustness against deviations in the electrical parameters of the machine, by acting on the angle of the current vector; and a feed-forward to improve the controller action. In parallel, the current vector is analytically determined in a normalised system. Simulation results that evaluate the performance of the controller over standardised driving cycles demonstrate the validity of the proposal.

Index Terms—Dual three-phase, IPM, EV, flux weakening

I. INTRODUCTION

Interior Permanent magnet synchronous machines (IPMSMs) are widely used in both Electric Vehicle (EV) and Hybrid EV (HEV) applications. These machines are compact and exhibit high power density and high efficiency [1]. In addition, the reduced per-phase power, torque ripple, low harmonic content and improved fault tolerance of multi-phase technologies over conventional three-phase systems make them attractive for applications with high power and reliability requirements [2].

Dual three-phase configuration is one of the most studied multi-phase architecture. Depending on the phase arrangement between the three-phase sets, they can be classified as symmetrical (shifted by 0 or π/3) or asymmetrical (shifted by π/6). With the latter configuration (Fig. 1), sixth harmonic torque pulsations produced by the two winding sets are in anti-phase and, therefore, they are cancelled [3].

Regardless of such phase shift, vector control strategies for dual three-phase IPMSMs can be broadly divided into two categories: the ones using two individual current controllers, based on the double dq synchronous frames model with mutual coupling voltages between both frames [4], [5]; and the ones based on vector space decomposition (VSD) control, with two currents regulators per each generated sub-plane [6], [7]. The latter techniques are prevailing since there is not mutual coupling between sub-planes and, as a result, a more accurate torque regulation can be provided [8].

To extend the operating speed range of an EV drive, which is constrained by the battery voltage, flux weakening (FW) needs to be applied. FW strategies can be classified into the feed-forward, where Look-up Tables (LUTs) or analytical calculations are used [9]–[11], feedback [12], [13], and hybrid [14] families. For example, the Vector Current Control (VCC) approach is the most commonly adopted feedback-based FW alternative, where three types of regulation are possible: on voltage magnitude [15]–[17], voltage error [10], and duty cycles [18].

A hybrid flux weakening control strategy is proposed, which uses a VCC to regulate the voltage magnitude considering the current angle variation as output, as suggested by Bolognani in [15]. In addition, to improve the performance of the FW algorithm, a feed-forward technique is integrated, where current trajectories are analytically determined in a normalised system depending on the angle position of the current vector [17], [19].

This paper is organised as follows: In section II, the mathematical model of a dual-three phase IPMSMs is briefly introduced. Achievable operating regions are introduced in Section III. The hybrid flux weakening control based on a normalised system is developed in Section IV. Finally, simulation results are presented and discussed in Section IV and conclusions are provided.
According to the VSD theory [6], the six-dimensional machine, represented in the ABC-XYZ frame, can be decomposed into three orthogonal sub-spaces: $\alpha\beta, z_1z_2$ and $o_1o_2$. By applying the transformation matrix $T_6$ of (2), harmonics in the ABC-XYZ frame are mapped into different sub-planes:

The fundamental and $(12k \pm 1)$th, $k = 1, 2, \ldots$ harmonics are mapped into $\alpha\beta$; the $(6k \pm 1)$th, $k = 1, 3, 5, \ldots$ into $z_1z_2$; and the $(3k)$th, $k = 0, 1, 3, 5, \ldots$ into $o_1o_2$ sub-plane. Thus,

$$[T_6]^T = \begin{bmatrix}
1 & \cos(4\theta_s) & \cos(8\theta_s) & \cos(12\theta_s) & \cos(16\theta_s) & \cos(20\theta_s) \\
1 & \sin(4\theta_s) & \sin(8\theta_s) & \sin(12\theta_s) & \sin(16\theta_s) & \sin(20\theta_s) \\
0 & \cos(4\theta_s) & \cos(8\theta_s) & \cos(12\theta_s) & \cos(16\theta_s) & \cos(20\theta_s) \\
0 & \sin(4\theta_s) & \sin(8\theta_s) & \sin(12\theta_s) & \sin(16\theta_s) & \sin(20\theta_s) \\
1 & \cos(2\theta_s) & \cos(4\theta_s) & \cos(6\theta_s) & \cos(8\theta_s) & \cos(10\theta_s) \\
0 & \sin(2\theta_s) & \sin(4\theta_s) & \sin(6\theta_s) & \sin(8\theta_s) & \sin(10\theta_s) \\
1 & 1 & 1 & 1 & 1 & 1 \\
0 & 0 & 0 & 0 & 0 & 0
\end{bmatrix}$$

where $\theta_s = \pi/6$.

The variables in the $\alpha\beta$ sub-plane can be converted into the $dq$ synchronous frame by applying the standard Park transformation, and the variables in $z_1z_2$ can be converted to a new frame, designated as $d_zq_z$ [20]:

$$[F_d] = \begin{bmatrix}
\cos \theta_e \\
\sin \theta_e
\end{bmatrix} \begin{bmatrix}
\cos \theta_e \\
\sin \theta_e
\end{bmatrix} \begin{bmatrix}
F_{\alpha} \\
F_{\beta}
\end{bmatrix} = T_{dq} F_{\alpha\beta},$$

where $F$ can be voltage ($v$), current ($i$) or flux ($\psi$). Then, the $(6k \pm 1)$th, $k = 1, 3, 5, \ldots$ harmonics in $z_1z_2$ sub-plane are converted to $(6k)$th harmonics in the $d_zq_z$-frame:

$$[F_{d_z}] = \begin{bmatrix}
\cos \theta_e \\
\sin \theta_e
\end{bmatrix} \begin{bmatrix}
\cos \theta_e \\
\sin \theta_e
\end{bmatrix} \begin{bmatrix}
F_{d_1} \\
F_{q_1}
\end{bmatrix} = T_{d_zq_z} F_{d_1q_1},$$

The mathematical model of a dual three-phase IPMSM by means of VSD is detailed in [8], where the voltage equations in the $dq$-frame ($\alpha\beta$ sub-plane) and $d_zq_z$-frame ($z_1z_2$ sub-plane) can be expressed as:

$$[v_d] = \begin{bmatrix}
R_s + L_d \dot{q} \\
R_s + L_q \dot{q}
\end{bmatrix} \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \dot{\omega}_e \begin{bmatrix}
-L_d q_i \\
-L_q i_d
\end{bmatrix} + \lambda_m,$$

$$[v_d] = \begin{bmatrix}
R_s + L_d \dot{q} \\
R_s + L_q \dot{q}
\end{bmatrix} \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \omega_e \begin{bmatrix}
-L_d q_i \\
-L_q i_d
\end{bmatrix} + \lambda_m,$$

where $\lambda_m$ is the PM flux linkage and $\omega_e$ is the electrical speed. If full mutual coupling between the two sets and full mutual coupling between phases in each set are assumed, the equivalent inductances $L_d^{eq}$ and $L_q^{eq}$ in the $dq$-frame, and $L_{d_z}$ and $L_{q_z}$ in the $dq_z$-frame can be simplified as:

$$L_d^{eq} = L_{sl} + 3L_d, L_q^{eq} = L_{sl} + 3L_q,$$

$$L_{d_z} = L_{sl}, L_{q_z} = L_{sl}.$$
theoretical maximum achievable speed is not finite. Otherwise, the maximum speed is finite at point \((\vec{i}_d, \vec{i}_q) = (-1, 0)\), see Fig. 2.

On the other hand, by normalising (9), the electromagnetic torque \(\tau_{em}\) is expressed as:

\[
\tau_{em} = \vec{i}_q[1 + (1 - \xi)\vec{i}_d],
\]

where several \(\vec{i}_d\) and \(\vec{i}_q\) combinations that lead to the same torque value can be obtained. Thus, proper current trajectory calculations are fundamental if losses are required to be reduced. In accordance with (10) and (11), four regions can be distinguished for a synchronous machine: Maximum Torque per Ampere (MTPA) region, Flux Weakening (FW) region (without and with torque reduction) and Maximum Torque per Voltage (MTPV) region, also known as deep flux weakening region.

A maximum torque per applied current modulus is guaranteed in the MTPA trajectory, and ohmic losses are minimised, which prevail at low speeds [23]. A MTPA point is applicable provided that it is not located beyond the ellipse related to the voltage limit. Otherwise, flux weakening mode is reached and a new current angle and current magnitude must be calculated to guarantee minimum phase currents. The FW trajectory (Fig. 2(a)) is referred to the vectors intersecting the reference torque curve \(G_1\) and \(G_2\) points. The reference torque is maintained through this trajectory until reaching the current limit, where it decreases. Finally, the MTPV trajectory is defined as the curve where the maximum torque is reached with a minimum voltage \(H\) point which intersects a maximum torque curve for a voltage limit is shown in Fig. 2(b). This trajectory minimises magnetic losses, which are prevalent at high speeds.

IV. Flux weakening control based on a normalised system

A. General description

In this paper, a hybrid flux weakening control strategy that considers all the aforementioned operation regions and determines the current reference trajectories from equations expressed in a normalised system is proposed. This algorithm is integrated in the Field Oriented Control (FOC) as shown in Fig. 3. Note that only the \(dq\) sub-plane is used by the controller.

Considering \(\vec{i}_d\) and \(\vec{i}_q\) in terms of angle and magnitude, an angular displacement \((\Delta \gamma_f)\) is added to the MTPA angle \((\gamma_M)\) to obtain a reference current angle \((\gamma^*)\), which is used to maintain the stator voltage modulus inside the voltage limit (Fig. 4(a)). Thus, the operating range of \(\gamma^*\) is bounded by \([\gamma_M; \pi]\).

The term \(\Delta \gamma_f\) (Fig. 4(a)) is determined from a feed-forward block, which calculates the theoretical displacement of the current angle \((\Delta \gamma_a)\) from the torque reference \((\tau_{em}^r)\) and the normalised electrical speed \(\omega_e\); and a voltage feedback controller, providing \(\Delta \gamma_r\). The latter aims to compensate errors in the determination of the analytic FW trajectory due to possible parameter deviations in the drive system.

Regarding the current reference magnitude determination, magnitudes that correspond to the MTPV trajectory \((i_{mtpv})\) and to the magnitude required to achieve the set-point torque \((i_{tem})\) are determined (Fig. 4(b)). Depending on the position of \(\gamma^*\), two dynamic saturation blocks provide a smooth transition between all the operating regions. To obtain the final reference current magnitude \(i_{tem}^*\), without exceeding the \(i_{max}\) limit, \(i_{tem}\) is saturated by \(i_{mtpv}\) as upper limit, provided that the latter does not exceed the current limit \(i_{max}\). Otherwise, the upper limit is equal to \(i_{max}\).

B. MTPA angle calculation

For synchronous machines, the MTPA currents \((\vec{i}_{d,M} \text{ and } \vec{i}_{q,M})\) are obtained for a given torque reference [23]:

\[
\tau_{em}^* = \frac{\vec{i}_{q,M}}{2}[1 + \sqrt{1 + 4(1 - \xi)\vec{i}_{d,M}^2}],
\]

\[
\tau_{em}^* = \frac{\vec{i}_{d,M}}{1 - \xi}[1 + (1 - \xi)\vec{i}_{d,M}]^3.
\]

In this way, the MTPA angle \((\gamma_M)\) is obtained as:

\[
\gamma_M = \arccos \frac{\vec{i}_{d,M}}{\vec{i}_{q,M}} \rightarrow \gamma_M^2 = \vec{i}_{d,M}^2 + \vec{i}_{q,M}^2.
\]
C. Feed-forward angular displacement calculation

With respect to the the optimal currents in FW, these can be determined by considering the following optimisation problem [23]:

$$\min \left( L_d^{eq} e^{eq}_d \left[ i_d + \frac{\lambda_m}{P_d} \right] \right)^2 + (L_q^{eq} i_q)^2 \leq (V_{max}/\omega_e)^2, \quad (17)$$

subject to $T_{em} = 3N_p I_m + (L_d^{eq} - L_q^{eq}) i_d i_q$.

and a fourth order polynomial is obtained in terms of $i_d$ [23]. This polynomial can be normalised with the base values indicated in Table I:

$$\bar{i}_d^2 + A_n \bar{i}_d^4 + B_n \bar{i}_d^3 + C_n \bar{i}_d + D_n = 0, \quad (18)$$

where,

$$A_n = \frac{2(2 - \xi)}{1 - \xi}, \quad B_n = \frac{1}{(1 - \xi)^2} + \frac{4}{1 - \xi} + 1 - \bar{v}_{max}^2/\Omega_e^2, \quad (19)$$

$$C_n = 2 \left[ \frac{1}{(1 - \xi)^2} + \frac{1}{1 - \xi} - (1 - \bar{v}_{max}^2/\Omega_e^2) \right], \quad (20)$$

$$D_n = \frac{1}{(1 - \xi)^2} \left[ 1 + \bar{v}_{em}^2 - \bar{v}_{max}^2/\Omega_e^2 \right]. \quad (21)$$

Analytical solution of (18) can be obtained by means of Ferrari’s method, thoroughly detailed in [23]:

$$\left( \bar{i}^{FW}_d, \bar{i}^{FW}_q \right) = \left( -\frac{A_n}{4} - \frac{\eta}{2} + \frac{\mu}{2} \right) + (1 + (1 - \xi) \bar{i}^{FW}_d). \quad (22)$$

On the other hand, normalising the MTPV trajectories defined in [23], the following is obtained:

$$\bar{i}^{MTPV}_d = -\xi + \sqrt{\xi^2 + 8(\bar{v}_s/\Omega_e)(1 - \xi)}/4(1 - \xi) - 1, \quad (23)$$

$$\bar{i}^{MTPV}_q = \frac{1}{\Omega_e} \sqrt{(\bar{v}_s/\Omega_e)^2 - (\bar{i}^{MTPV}_d + 1)^2}. \quad (24)$$

Once both FW and MTPV currents have been obtained, it is necessary to define the angle variation in all operating regions:

$$\Delta \gamma_{MTPA} = 0, \quad \Delta \gamma_{FW} = \arctan(\bar{i}^{FW}_q/\bar{i}^{FW}_d) + \pi - \gamma_M, \quad (25)$$

$$\Delta \gamma_{MTPV} = \arctan(\bar{i}^{MTPV}_q/\bar{i}^{MTPV}_d) + \pi - \gamma_M. \quad (26)$$

Therefore, $\Delta \gamma^a$ is determined depending on the following conditions:

$$\Delta \gamma^a = \begin{cases} \Delta \gamma_{MTPA} & \text{if } \Delta \gamma_{FW} \leq 0 \text{ and } |i_d^{FW}| \leq i_d^{FW} \leq |i_d^{FW}|, \\ \Delta \gamma_{FW} & \text{if } \Delta \gamma_{FW} > 0 \text{ and } i_d^{FW} \leq |i_d^{FW}|, \\ \Delta \gamma_{MTPV} & \text{otherwise.} \end{cases} \quad (27)$$

It is worth highlighting that $\Delta \gamma^a$ is calculated without taking into account the voltage drop produced by $R_s$. Thus, a feedback voltage loop is included to compensate stator voltage excesses caused by neglecting the influence of $R_s$ or produced due to electrical parameter deviations.

Finally, $\gamma^s$ is defined as the sum of the analytically calculated angular displacement ($\Delta \gamma^a$), the MTPA angle ($\gamma_M$) and the output of the integral controller ($\Delta \gamma^c$).

D. Current magnitude calculation

Finally, both $i_{mtpv}$ and $i_{tem}$ magnitudes are calculated. For $i_{tem}$ determination, the total angular displacement $\Delta \gamma^f$ needs to be considered. If $\Delta \gamma^f = 0$, $i_{tem}$ must have the value corresponding to the MTPA trajectory for $\gamma^s$, which is obtained by solving $\partial \tau_{em}/\partial \gamma = 0$; otherwise, $i_{tem}$ is related to the FW region, and the voltage limit has to be taken into account. Thus, $i_{tem}$ is obtained by applying Ferrari’s method to (18), providing $i_d$ in terms of angle and magnitude.

On the other hand, the MTPV condition for synchronous machines is expressed as $\partial \tau_{em}/\partial \delta = 0$. Considering the currents in (13) with respect to $i_s$ and $\delta$, where the latter is the angle of the voltage vector referred to positive d axis [19], voltages in normalised dq-frame are rewritten as:

$$\bar{v}_d = \bar{v}_s \cos \delta = -\bar{v}_s \Omega_e \xi - \Omega_e \xi \bar{v}_s \sin \gamma, \quad (28)$$

$$\bar{v}_q = \bar{v}_s \sin \delta = \Omega_e \text{sat} + \Omega_e (\bar{i}_d + 1) + \Omega_e (\bar{i}_d \sin \gamma + 1).$$

and, in consequence,

$$\frac{\partial \tau_{em}}{\partial \delta} = \frac{(1 - \xi)}{\Omega_e^2} [\bar{v}_s^2 \cos^2 \delta - \bar{v}_s^2 \sin^2 \delta] - \bar{v}_s \sin \delta = 0. \quad (29)$$

By replacing voltage vector terms with expressions in terms of $i_s$ and $\gamma$, (29) can be rewritten as:

$$\frac{(1 - \xi)}{\xi} \left[ (\bar{i}_s \sin \gamma)^2 \bar{v}_s^2 - (\bar{i}_s \cos \gamma + 1)^2 \right] = \bar{i}_s \cos \gamma + 1. \quad (30)$$
Therefore, (30) is solved as a function of $\gamma$, and the MTPV current modulus is defined, in terms of cosines, as:

$$i_{\text{mtpv}} = \frac{-(2-\xi)\cos(\gamma) - \sqrt{(2-\xi)^2\cos(\gamma)^2 - 4(1-\xi)(\cos(\gamma)^2(\xi^2+1)-\xi^2)}}{2(1-\xi)(\cos(\gamma)^2(\xi^2+1)-\xi^2)},$$

(31)

V. SIMULATION RESULTS

A detailed model of an asymmetrical dual three-phase IPMSM drive, whose most relevant parameters are listed in Table II, has been implemented in the Matlab/Simulink environment together with the proposed controller. The field weakening algorithm has been tested under close-to-real driving conditions. In particular, the torque and speed profiles have been calculated for a given EV which is circulating under the New European Driving Cycle (NEDC). Considering the high computational burden of the model, the model has been loaded into an OPAL-RT OP4510 real-time digital platform (Intel Xeon E3 v5 CPU, 4 core, 8 MB cache, 3.5 GHz) and executed in the simulation mode (no real-time). This way, the simulation of the complete NEDC cycle has been carried out in less than 1 hour of execution time. As the field weakening algorithm only enters when the vehicle operates at high speeds, the presented simulations only focus on the extra urban driving profile interval of the NEDC.

Fig. 5(a) shows how, depending on the operation conditions of the drive system, the current set-point generation block calculates online the current reference angle $\gamma^* = \gamma_M + \Delta \gamma^f$ and magnitude $i^*_q$. Note how, in this particular drive system, the MTPV trajectory is outside the maximum current circle and, thus, this mode of operation does not enter. Once transformed into the dq frame, the d- and q-axis reference currents are determined and regulated by means of the FOC algorithm (Fig. 5(b)). This way, the stator voltage $V_{dq}$ is kept below the maximum limit $V_{\text{lim}}$ during the whole driving profile (Fig. 5(c)).

VI. CONCLUSIONS

A hybrid flux weakening strategy has been proposed for the calculation of current trajectories in dual three-phase IPMSMs. Polar coordinates and the normalised system representation are exploited by the algorithm to generate the current vector reference. This strategy can operate in all achievable regions (MTPA, FW, and MTPV) and correct operation is guaranteed by the feedback block even if the stator resistance is neglected or electrical parameter deviations occur. The effectiveness of the algorithm is proven by results in a driving cycle.

It is important to point out that, although this algorithm has been developed with asymmetric dual three-phase IPMSMs in mind, it can be incorporated into the control of symmetric ones or even for three-phase systems.

Currently, the dual three-phase IPMSM simulated in this work is under construction and it is expected that the proposal will be experimentally validated in the near future.

<p>| Table II: Parameters of the Dual Three-Phase IPMSM Drive. |
|---------------------------------|---------------------------------|</p>
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of pole pairs $N_p$</td>
<td>19</td>
</tr>
<tr>
<td>Stator resistance $R_s$</td>
<td>61.43 [mΩ]</td>
</tr>
<tr>
<td>Equivalent d-axis inductance $L_{dq}^e$</td>
<td>1.00 [mH]</td>
</tr>
<tr>
<td>Equivalent q-axis inductance $L_{eq}^q$</td>
<td>1.35 [mH]</td>
</tr>
<tr>
<td>PM flux linkage $\lambda_m$</td>
<td>0.038 [Wb]</td>
</tr>
<tr>
<td>DC bus nominal voltage $V_{dc}$</td>
<td>400 [V]</td>
</tr>
<tr>
<td>Maximum torque $T_{max}$</td>
<td>54 [N.m]</td>
</tr>
<tr>
<td>Maximum mech. speed $\omega_m$</td>
<td>5000 [rpm]</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>25 [kHz]</td>
</tr>
<tr>
<td>Simulation step</td>
<td>1 [μs]</td>
</tr>
</tbody>
</table>

(a) Angle and magnitude determination provided by the proposed set-point generation algorithm.

(b) dq currents.

(c) Flux weakening control.

Fig. 5. OPAL-RT results for NEDC.
ACKNOWLEDGEMENTS

This work has been supported by the “Ministerio de Ciencia e Innovación” of the Government of Spain through the PID2020-115126RB-I00 project, by the Government of the Basque Country through project KK-2020/00077 and by the European Union’s Horizon 2020 research and innovation programme under project FITGEN grant agreement No 824335.

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