Maximum Torque per Voltage Flux-Weakening strategy with speed limiter for PMSM drives

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Abstract—This paper presents an enhanced Flux Weakening (FW) control scheme for Permanent Magnet Synchronous Motors (PMSMs), focused on electric vehicle applications. The novelty of the proposed algorithm is the integration in a unified scheme of both the accelerator pedal as torque reference and the cruise speed limiter (CSL) as speed limit, without interfering between them until this limit is achieved. The dq-axis current references are calculated from the proposed algorithm by using a polar coordinate system and a per-unit system. The latter is based on the characteristic machine parameters which aim to ease and simplify the algorithm implementation. Moreover, it takes advantage of the smooth transition between the Low Back Electromotive Force (LBEF) zone and the FW zone thanks to a voltage loop which changes the current-vector angle. Another fundamental merit of the proposed scheme is its capacity to work in all the dq-plane throughout the Maximum Torque per Ampere (MTPA), Constant Torque (CT), Current and Voltage Limit (CVL), Maximum Torque per Voltage (MTPV) and Constant Speed (CS) strategies without switching the algorithm. Simulations and experimental results from a real exterior-rotor Interior Permanent Magnet Synchronous Motor (IPMSM) direct-drive e-motorbike verify the effectiveness of the proposed method.

Index Terms—Flux Weakening, Maximum Torque per Voltage, Permanent magnet motors, Torque control, Velocity control.

I. INTRODUCTION

NOWADAYS, the use of mild-hybrid and full-electric vehicles is rapidly increasing due to the importance of reducing greenhouse gas emissions. Another significant tendency in the automotive industry is the necessity to provide torque assistance and torque boosting in the powertrain. Moreover, the newest electric traction requires motors with high torque at low speeds and wide speed range at constant power. Based on the requirements mentioned above, a Permanent Magnet Synchronous Motor (PMSM) is an adequate type for this application [1].

Below the nominal speed, the Maximum Torque per Ampere (MTPA) strategy is used to minimize the required current. When the speed increases, the back electromotive force does so and, from a certain speed, it is mandatory to diminish the

stator magnet flux through the stator currents, increasing the negative d -axis current. Thenceforward, the motor is working in the denominated Flux Weakening (FW) zone.

By the authors' best knowledge, literature shows several FW algorithms working either at Constant Torque (CT) and Current and Voltage Limits (CVL) zones. These techniques have been widely classified and analyzed in [2], citing the most important characteristics of each one: Analytical Direct Calculation [3], [4], Direct Open Loop Algorithm with Experimental Look Up Tables (LUT) [5]–[7], Single Current Regulator (SCR) [8]–[10], Torque and Flux Control Method with LUT [11], [12], Unified Direct Flux Vector Control (UDFVC) in the stator flux frame [13], [14] and Vector Current Control (VCC) [15]–[21]. It should be highlighted that for electric vehicles, the need for a wide constant power speed range demands the control to work in the Maximum Torque per Voltage (MTPV) zone, so these methods are discarded.

Some authors have adapted the previously cited methods with the possibility to achieve the MTPV zone: Analytical Direct Calculation [22]–[24], Direct Open Loop Algorithm with Simulated or Experimental LUT [25]–[27], Torque and Flux Control Method with LUT [25], [28]–[30]. However, VCC is the most common one [31]–[38]. These methods may also be divided into three type of regulation: on the voltage magnitude [32], [34], [36]–[38], on the voltage error [31], [35] and on the duty cycles [33].

The main drawback of Analytical Direct Calculation is its parameter-variation sensitiveness, especially if high currents are required. Moreover, these control algorithms for an Interior Permanent Magnet Synchronous Motor (IPMSM) take many calculations in real-time, increasing the burden of the microcontroller. LUTs in Direct Open Loop and Torque and Flux Control Methods are convenient for considering both the magnetic saturation and resistor voltage drop, but they depend on previous experimental results or Finite Element Analysis.

In this paper, a VCC method that controls the voltage norm is used. The main characteristics of the analyzed methods are listed in Table I. Bolognani in [20] suggested that calculating the FW reference with the current angle, instead of the d -axis current, maximizes the dynamic performance. So, a lower voltage margin is considered, and high torque and efficiency are achieved. In none of the analyzed algorithms torque and speed references are contemplated in the same scheme.

As a result, this paper proposes an enhanced algorithm that considers the requirements of electric vehicles, converting the accelerator pedal and the cruise speed limiter (CSL) in reference signals of a unified scheme. This algorithm lets the

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TABLE I
MTPV ALGORITHMS COMPARISON

Reference	Reference System	Coordinate System	Transition to FW	Output Voltage Controller	Reference Signal
[32]	Real	Cartesian	Switched	dq -axis current	Speed
[34]	Real	Cartesian	Continuous	d -axis current	Speed
[36]	Real	Cartesian	Switched	d -axis current	Speed
[37]	Real	Cartesian	Continuous	d -axis current	Torque
[38]	Per-unit	Polar	Continuous	angle current	Speed

vehicle have the dynamics of the torque control until the speed limit is reached, by not interfering with the current controllers until this threshold is achieved. It smoothly changes to a FW zone, letting the motor work in all the possible trajectories: MTPA, CT, CVL, MTPV, and Constant Speed (CS). This soft transition is done by an explicit control scheme that avoids switching among the different trajectories. This transition is achieved thanks to a voltage loop controller, which gives the angle current variation from the MTPA curve to the FW zone. Moreover, it takes advantage of a per-unit system based on the motor parameters to reduce the computational burden of a fixed-point microcontroller in real-time, making it more attractive for industrial applications.

The content of the paper is organized in six sections. In Section II, a per-unit PMSM model is presented, as well as all the possible trajectories below and above the nominal speed. In Section III, the proposed FW strategy is explained. In Section IV and V, the simulation and experimental results are presented for a real exterior-rotor Interior Permanent Magnet Synchronous Motor (IPMSM) direct-drive e-motorbike. Finally, Section VI details the conclusion reached after this research was conducted.

II. ACHIEVABLE OPERATING POINTS

In Section II-A, a per-unit PMSM model and its electrical limits are presented. In Section II-B, the trajectory at low speed is described, and the base speed is defined. Finally, in Section II-C, all the FW trajectories are analyzed.

A. PMSM per-unit model in the Cartesian coordinate system

In accordance with [39], the PMSM electrical dynamic equations referred to the synchronous rotor frame with the d -axis aligned to the rotor flux, and by implementing the Park transformation which maintains the current and voltage modulus invariant, are

$$v_d(t) = R_s i_d(t) + L_d \frac{di_d(t)}{dt} - \omega_e(t) L_q i_q(t) \quad (1)$$

$$v_q(t) = R_s i_q(t) + L_q \frac{di_q(t)}{dt} + \omega_e(t) L_d i_d(t) + \omega_e(t) \lambda_m, \quad (2)$$

where $v_d(t)$, $v_q(t)$, $i_d(t)$ and $i_q(t)$ are the time-dependent voltage and current dq components, L_d and L_q are the motor inductance at the dq -axis, λ_m is the flux linkage due to the spinning of magnets, R_s is the winding phase resistance and $\omega_e(t)$ is the time-dependent electrical speed.

Fig. 1. Phasor diagram in dq -axis.

At steady-state conditions the voltage equations are simplified as

$$V_d = R_s I_d - \omega_e L_q I_q \quad (3)$$

$$V_q = R_s I_q + \omega_e L_d I_d + \omega_e \lambda_m, \quad (4)$$

where V_d , V_q , I_d and I_q are the non time-dependent voltage and current dq components, and ω_e is the non time-dependent speed.

The relationship between mechanical and electrical speed in a PMSM is expressed as

$$\omega_m = \frac{\omega_e}{pp}, \quad (5)$$

where ω_m is the mechanical speed and pp is the pair of poles.

Equations (3) and (4) may be represented in a diagram phasor as in Fig. 1. The proposed method is based on controlling the norm (I_s) and angle current (γ) instead of the conventional methods whose aim is to control I_d and I_q . It has to be noted that γ and δ (voltage angle) are referred with reference to the positive d -axis.

The electromagnetic torque (T_{em}) is expressed as

$$T_{em} = \frac{3}{2} pp (\lambda_m I_q + (L_d - L_q) I_d I_q). \quad (6)$$

A per-unit system is used to simplify the FW analysis. Base values are listed in Table II. The subscript N indicates the nominal value of the electrical speed. Fundamentally, the base current is defined as the short-circuit current (I_{sc}) of the machine. This current represents the flux generated in the stator by the winding currents with the same direction and value as the flux linkage of the permanent magnets. Moreover, the base torque is the corresponding value when the motor is considered an SPMSM and the base current is equal to I_{sc} . In Table II, the parameter ξ is defined as the saliency ratio.

Achievable operating points are restricted according to the current and voltage limits. The maximum available current ($i_{s,max}$) is expressed as

$$\sqrt{i_d^2 + i_q^2} = i_s \leq i_{s,max} = \frac{\min \{I_{motor,max}, I_{inv,max}, I_{bat,max}\}}{I_b}, \quad (7)$$

where i_d and i_q are the non time-dependent current dq components expressed in the per-unit system, i_s is the modulus current vector in the per-unit system, $I_{motor,max}$ is the maximum

TABLE II
BASE VALUES FOR AN SPMSM AND AN IPMSM

Symbol	Quantity	SPMSM ^a	IPMSM ^b
λ_b	base flux linkage	λ_m	
ω_b	base electrical rotational speed	$\omega_{e,N}$	
I_b	base current	$I_{sc} = \frac{\lambda_m}{L}$	$I_{sc} = \frac{\lambda_m}{L_d}$
V_b	base voltage	$\lambda_m \omega_{e,N}$	
Z_b	base impedance	$\omega_{e,N} L$	$\omega_{e,N} L_d$
T_b	base torque	$\frac{3}{2} \text{pp} \frac{\lambda_m^2}{L}$	$\frac{3}{2} \text{pp} \frac{\lambda_m^2}{L_d}$
ξ	saliency ratio	1	$\frac{L_q}{L_d}$

^a SPMSM: $L_d = L_q = L$. ^b IPMSM: $L_d \neq L_q$.

motor current, $I_{inv,max}$ is the maximum inverter current, $I_{bat,max}$ is the maximum battery current, and I_b is the base current value indicated in Table II. So, $i_{s,max}$, as well as i_s , can be higher or less than 1, depending on if the electrical and thermal battery, converter and motor limits are higher or less than I_{sc} .

The maximum available voltage ($v_{s,max}$) is expressed as

$$\sqrt{v_d^2 + v_q^2} = v_s \leq v_{s,max} = \frac{V_{bat}}{\sqrt{3} V_b}, \quad (8)$$

where v_d and v_q are the non time-dependent voltage dq components expressed in the per-unit system, v_s is the modulus voltage vector in the per-unit system, V_{bat} is the battery voltage, and V_b is the base voltage value indicated in Table II. A Space Vector Pulse Width Modulation (SVPWM) strategy is chosen in this paper.

Expressing (3) and (4) in the per-unit system and neglecting the resistor voltage drop, (8) yields to

$$\frac{(i_d + 1)^2}{\left(\frac{v_{s,max}}{\Omega_e}\right)^2} + \frac{i_q^2}{\left(\frac{v_{s,max}}{\Omega_e \xi}\right)^2} \leq 1, \quad (9)$$

where Ω_e is the electrical speed in the per-unit system. Also, the electromagnetic torque in the per-unit system (τ_{em}) is expressed as

$$\tau_{em} = i_q [1 + (1 - \xi) i_d]. \quad (10)$$

Fig. 2 and Fig. 3 depict the characteristic curves for an SPMSM and an IPMSM, respectively. The current limit (7) is seen as a circle centered in the coordinates $(i_d, i_q) = (0,0)$, with a radius equal to i_s .

The voltage limit (9), in the case of an SPMSM, depicts a circle centered in the coordinates $(i_d, i_q) = (-1,0)$ and with a radius equal to $\frac{v_s}{\Omega_e}$. In the case of an IPMSM, (9) describes an ellipse with the same center, in which semi-major and semi-minor axes are equal to $\frac{v_s}{\Omega_e}$ and $\frac{v_s}{\Omega_e \xi}$, respectively. If the resistor voltage drop is considered, the voltage curve suffers a translation and a rotation depending on i_d and i_q values, but this topic is out of the scope of this paper.

Fig. 2. Characteristic curves for SPMSMs with $i_s > 1$. The red line is the current limit circle. The blue line is the voltage limit circle. The magenta lines are the constant torque straight curves. The green line is the MTPA straight curve. The yellow line is the MTPV straight curve.

Fig. 3. Characteristic curves for IPMSMs with $i_s > 1$. The red line is the current limit circle. The blue line is the voltage limit ellipse. The magenta lines are the constant torque hyperbolic curves. The green line is the MTPA hyperbolic curve. The yellow line is the MTPV hyperbolic curve.

The torque curves, in the case of SPMSMs, are straight lines which only depend on the i_q values as it is described in (10) with $\xi = 1$. Nevertheless, in the case of IPMSMs, the torque curves are rectangular hyperbolas according to (10) with $\xi \neq 1$.

B. LBEF trajectory and base speed in the polar coordinate system

The attainable trajectories are different if the back electromotive force is lower or higher than the maximum synthesizable stator voltage. From now on, the first case is cited as Low Back Electromotive Force (LBEF) zone and the latter as FW zone. In [39, p. 303-327] some strategies at the LBEF zone have been compared: Constant Torque Angle Control (CTAC), Unit Power Factor Control (UPFC), Constant Stator Flux Control (CSFC) and MTPA.

In this paper, the MTPA algorithm is chosen to minimize the current needed for the demanded torque. This decision is supported by the necessity in vehicle applications to reduce the cross-sectional area of the cable conductors.

The MTPA condition for PMSMs is defined as $\frac{\partial \tau_{em}}{\partial \gamma} = 0$, expressing (10) in terms of i_s and γ . The MTPA angle value (γ_{MTPA}) is expressed as

$$\gamma_{MTPA} = \frac{\pi}{2} \quad (11)$$

$$\gamma_{MTPA} = \arccos \left(\frac{-1 + \sqrt{1 + 8i_s^2 (1 - \xi)^2}}{4i_s (1 - \xi)} \right), \quad (12)$$

represented in a green straight curve in Fig. 2 for SPMSMs (11) or in a green hyperbolic curve in Fig. 3 for IPMSMs (12).

The base speed is defined as the value when the voltage curve goes across the intersection point between the MTPA curve and the current circle (point Q in Fig. 2 and Fig. 3). From this value upwards, a FW technique has to be applied to maintain the work point inside the voltage curve. The stator voltage in the per-unit system at the base speed is expressed as

$$v_s = \sqrt{i_s^2 + 1} \quad (13)$$

$$v_s = \sqrt{\frac{(-1 + \sqrt{1 + 8(1 - \xi)^2 i_s^2})^2 (1 - \xi^2)}{16(\xi - 1)^2} - \frac{-1 + \sqrt{1 + 8(1 - \xi)^2 i_s^2}}{2(\xi - 1)} + (1 + \xi^2 i_s^2)}, \quad (14)$$

for SPMSM and IPMSM, respectively.

C. Flux Weakening trajectories

If the voltage curve center is located inside the current circle ($i_s > 1$), then the theoretical maximum achievable speed is infinite, as long as the MTPV curve is followed and no electromagnetic and mechanical losses are considered. This is represented in yellow as a straight curve in Fig. 2 for SPMSMs or as a hyperbolic curve in Fig. 3 for IPMSMs. On the contrary, if the voltage curve center is located outside the current circle ($i_s < 1$), then the maximum speed is finite at the point $(i_d, i_q) = (-i_s, 0)$.

The MTPV trajectory is defined as the curve where the maximum torque at minimum current is achieved, considering the voltage limit. The MTPV condition for PMSMs is defined as $\frac{\partial \tau_{em}}{\partial \delta} = 0$, expressing (10) in terms of i_s and δ . The MTPV current value (i_{MTPV}) is expressed as

$$i_{MTPV} = \frac{1}{\cos(\gamma)} \quad (15)$$

$$i_{MTPV} = \frac{-(2 - \xi)\cos(\gamma) - \sqrt{(2 - \xi)^2 \cos^2(\gamma) - 4(1 - \xi)(\cos^2(\gamma)(\xi^2 + 1) - \xi^2)}}{2(1 - \xi)(\cos^2(\gamma)(\xi^2 + 1) - \xi^2)}, \quad (16)$$

for SPMSM and IPMSM, respectively.

At low speed, the motor can work at any point of the MTPA curve, and when the voltage curve crosses the MTPA zone, a FW strategy has to be applied, maintaining the stator voltage inside its limit. Depending on the torque level, there is a speed range in which the torque value can be maintained. This zone is named Constant Torque Control (CTC). To sum up, all possible strategies at the FW zone are based on the voltage limit condition: Current and Voltage Limit Control (CVLC), CTC, and MTPV.

Depending on the value of i_s different trajectories can be followed. If $i_s > 1$, then the motor can follow trajectory 1.1, 1.2 or 1.3, as it is depicted in Fig. 4. However, if $i_s < 1$, the motor can only follow trajectory 2.1 and 2.2, as it is showed in Fig. 5.

III. FLUX-WEAKENING STRATEGY

This section presents the control algorithm that calculates the dq -axis current set-points from the torque and speed references. The algorithm, represented in Fig. 6, smoothly changes among all trajectories explained in Section II-B and II-C, and

(a) Trajectory 1.1.

(b) Trajectory 1.2.

(c) Trajectory 1.3.

Fig. 4. Trajectories at MTPA and FW zones for IPMSMs if $i_s > 1$.

it allows the motor to work with torque and speed reference in a unified scheme. The per-unit system representation facilitates the implementation in a fixed-point microcontroller.

In Section III-A, the voltage control loop is explained. In Section III-B, the FW trajectory decision is analyzed. Finally, in Section III-C the speed control loop is presented.

A. Voltage control loop

The current regulators give the voltage feedback to the voltage control loop, in which the main objective is to maintain the voltage modulus lower than the voltage limit (9), considering a security factor (K_{FW}). The onset of the FW zone is determined when the voltage modulus exceeds the maximum value.

Fig. 5. Trajectories at MTPA and FW zones for IPMSMs if $i_s < 1$.

The voltage control loop of conventional methods manages Δi_d . Its main drawback is that Δi_d has to increase at the same time that speed increases, except for the moment when the work point enters the MTPV trajectory for an IPMSM, when it has to decrease. This change on the PI control action for the same voltage error can be solved by changing the PI controller constants or letting an over-voltage during this transition. However, in the proposed method, the voltage control loop acts on $\Delta\gamma$, which always increases with speed, allowing the system to transition smoothly.

In this control scheme, $\Delta\gamma$, used to hold the work point inside the voltage limits, is added to γ_{MTPA} to achieve the current angle reference (γ^*). The saturation blocks are added to guarantee that γ^* stays in the range of $[\gamma_{\text{MTPA}}, \pi]$.

B. Flux Weakening trajectory decision

The trajectory is chosen from the torque reference (τ_{em}^*) and the current angle reference (γ^*). Mainly, this decision is made thanks to two dynamic saturations that shrink the current modulus calculated to achieve the reference value.

Firstly, the current modulus needed for the required torque ($i_{\tau_{\text{em}}}$) is calculated according to (10). From this value, γ_{MTPA} is calculated following (12). Moreover, the current needed to work in the MTPV zone is calculated by replacing γ for γ^* into (16), so limiting to the maximum current specified in (7).

These saturations lead to a smooth transition among the attainable zones as is shown in Fig. 7. It depicts four different cases to get the minimum current:

TABLE III
IPMSM CHARACTERISTIC PARAMETERS

Symbol	Value	Unit
R_s	17	m Ω
L_d	70	μH
L_q	79	μH
λ_m	0.023	(V·s)/rad
V_{bat}	48	V
$I_{s,\text{max}}$	467	A
pp	20	

- $\gamma_1 \rightarrow i_{\tau_{\text{em}},1} < i_{s,1} < i_{\text{MTPV},1}$: CTC
- $\gamma_2 \rightarrow i_{s,2} < i_{\tau_{\text{em}},2} < i_{\text{MTPV},2}$: CVLC
- $\gamma_3 \rightarrow i_{s,3} < i_{\text{MTPV},3} < i_{\tau_{\text{em}},3}$: CVLC
- $\gamma_4 \rightarrow i_{\text{MTPV},4} < i_{s,4} < i_{\tau_{\text{em}},4}$: MTPV

C. Speed control loop

The speed control loop independently acts to the torque reference through a PI controller that limits the speed to the reference value (Ω_m^*), enabling the motor to work in the Constant Speed Control (CSC) zone. The current mechanical speed in pu (Ω_m) is calculated from Ω_e using (5). The PI-controller output is used to limit the torque reference, acting as the upper torque saturation. The main advantage is that the dynamics of the rest of the control is not affected by the speed PI controller. It only takes action when the vehicle has the CSL switched on.

IV. SIMULATION RESULTS

Initially, an analytical study is carried out to compare the accomplishment of the proposed FW method compared with the conventional method. A direct drive IPMSM, which runs as the rear wheel of an e-motorbike, is used. The characteristic parameters are listed in Table III. The main parameters to define its characteristic curves, in the per unit-system, are $\xi = 1.1$ and $i_s = 1.4$.

Fig. 8 shows the torque-speed and power-speed curves corresponding to the conventional (MTPA + CVLC) and the proposed (MTPA + CVLC + MTPV) methods. The electromagnetic power is defined in the secondary y-axis as p_{em} . The proposed method achieves constant power during a wide speed range from $\omega_e = 1.7\omega_b$. Meanwhile, the output power decreases until it reaches 0 from this speed to $\omega_e = 4.2\omega_b$ in the conventional method. So, the presented algorithm obtains current torque during more speed range, therefore adapting to the electric vehicle needs.

Fig. 9 shows the evolution of i_d and γ regarding speed. Fig. 9a displays how i_d totally inverts its dynamics from $\omega_e = 1.7\omega_b$. This operation derives in a vortex evolution seen in the dq -trajectory, implying a higher value of K_{FW} . Meanwhile, Fig. 9b shows the continuous increase of $\Delta\gamma$. That is why the proposed voltage PI controller controls $\Delta\gamma$.

A Matlab-Simulink model is designed to validate the proposed method. A step torque reference of 1.1 pu, less than

Fig. 6. Control Algorithm Model. The reference values are represented in purple. The feedback values from the system are shown in red. The intermediate calculations are depicted in blue.

Fig. 7. Current Vector Selection for an IPMSM.

Fig. 8. Torque versus speed (red) and power versus speed (green) with conventional (dashed line) and proposed (solid line) FW methods.

the maximum value, with a speed limit of 1.31 pu, is applied, demonstrating that the proposed method can work in all possible trajectories. Fig. 10 demonstrates the control algorithm efficacy through the simulated dq -trajectory. Initially, the work point, represented by a cyan line, follows the MTPA strategy. Then, the speed increases, and the control forces to steadily

Fig. 9. Control variable of voltage PI controller versus speed with conventional (dashed line) and proposed (solid line) FW methods.

raise the angle γ , following the curves CTC, CVLC, and MTPV. From a specific speed value, the work point pursues the voltage ellipse, decreasing the applied torque, until the break-even point is reached.

Fig. 11 shows the torque response, analyzing the different trajectories in terms of torque. During the MTPA and CTC zones, the electric motor can apply the torque reference, whereas it can not do so in the CVLC, MTPV, and CSC zones because it maintains the work point inside the ellipse voltage. Finally, the applied torque is reduced until the system reaches the mechanical balance to comply with the speed limit, as can be seen in Fig. 12.

V. EXPERIMENTAL RESULTS

To verify the reliability of the method, as shown in Fig. 13, an experimental platform which has a DC motor as its load is set up. The test bench is composed of a 6.5 kW three-phase inverter, fed from a rated 48 V Lithium-ion battery, and a 2.5 kW and 510 rpm IPMSM. The inverter is managed by a fixed-point microcontroller where the algorithm has been implemented. The torque reference is given by a real throttle to emulate the operation of an electric motorbike.

Fig. 10. Representation in dq -axis of current limit (red), MTPA trajectory (green), MTPV trajectory (yellow), torque in steady-state (magenta), voltage ellipse considering the resistor drop at speed reference (blue), and simulated dq -trajectory (cyan).

Fig. 11. Simulation results of torque versus time. Torque reference (magenta, dashed line) and applied torque (magenta, solid line).

Fig. 12. Simulation results of mechanical speed versus time. Speed limit (blue, dashed line) and real speed (blue, solid line).

The motor can not work either at CVLC or MTPV zones due to the limitation of the elements available in the laboratory.

Fig. 13. Experimental setup of the IPMSM assembled with a DC motor, having the throttle signal as torque reference.

The real system has not enough inertia to accelerate slowly. As a consequence, if a high reference torque is applied, the motor rapidly accelerates to the speed limit. In Section IV, the inertia parameter is ten times higher than its real value, which is used in Section V, with the purpose to verify the effectiveness of the algorithm in all the speed range.

Initially, a step torque reference of 0.18 pu is applied to the machine. Also, the speed limit is fixed to 1.31 pu. At time 2 s, the torque reference is increased up to 0.36 pu; meanwhile, the speed limit is kept unchanged. At time 4 s, to analyze the performance of the voltage loop controller, the load is reduced in a third. The results are reported in Fig. 14 to Fig. 19.

Fig. 14 shows the work point in the dq plane. When the motor works below its base speed, the work point is sustained at the MTPA curve. If the speed reaches the limit at which a flux-weakening algorithm is needed, the voltage control loop increases the current angle, and the algorithm lets the motor to maintain the torque reference. Despite that, when the speed reference is overtaken, the reference torque is reduced until the speed reaches this value. At the moment the load is lessened in a third, the work point moves along the same ellipse, but the intersection point corresponds with a torque curve lower than previously.

Fig. 15 shows both the dq -axis reference and applied values. Until time 2.2 s, the work point resides in the MTPA curve, but once the speed increases until the voltage limit is achieved, the current angle rises by increasing the d -axis current following the CT zone. At time 2.6 s, the constant torque can not be followed, and the speed limit constraints the applied torque by reducing both the dq -axis currents. At time 4 s, as the load is decreased and to maintain the speed limit, the dq -axis currents are diminished to reduce the applied torque.

Fig. 16 shows the experimental phase currents during the test. Fig. 16a gives a notion of how the phase currents are modified during the transition between LBEF and FW zones, and when the work point works at a constant speed, including the load variation. Fig. 16b presents the transition between the Six-Step to the Continuous commutation mode. The electrical angle goes from a discrete to a continuous representation, by following a Linear Regression from the Least Squares Method explained in [40].

Fig. 17 shows the voltage waveform versus time. As can

Fig. 14. Representation in dq -axis of current limit (red), MTPA trajectory (green), MTPV trajectory (yellow), torque curve in steady-state (magenta), voltage ellipse considering the resistor drop at speed reference (blue), and simulated (cyan) and experimental (black) dq -trajectory.

Fig. 15. Comparison between the simulation and experimental results of the evolution of dq -axis current versus time. dq -axis current reference (blue and red, dashed line), simulation results (blue and red, solid line), and experimental results (black).

be seen at time 2.2 s, the voltage modulus gets its maximum value ($V_{bat}K_{FW}$). From that point, a flux-weakening technique is applied to maintain the voltage down this value. An adequate (K_{FW}) factor is essential to maintain the stability of the system control because there is a transient when the applied voltage is higher than its reference.

Fig. 18 shows the applied torque evolution at each zone. Until the maximum speed is not reached, the reference torque is applied. From that point, the applied torque decreases to maintain the speed limit even when the load torque is lessened.

Fig. 19 exhibits the evolution of the mechanical speed either in the simulation and the real model. The transition between the Six-Step mode and the Continuous mode provokes a slipping in the mechanical system between the test motor and the load, as can be seen at the beginning. At time 2.6 s, the speed limit is overcome until the applied torque is not reduced enough. At time 4 s, the load change lets to increase the speed until the applied torque is not reduced again.

To overcome the physical limitations of the setup, a Virtual

(a) Entire phase current waveforms.

(b) Zoom at the starting.

Fig. 16. Representation of phase current waveforms regarding time. Experimental phase A, B, and C currents represented in blue, orange and yellow waveforms, respectively.

Fig. 17. Experimental results of dq -axis voltage and modulus regarding time. dq -axis and modulus voltage experimental results in orange, blue and yellow, respectively. The black dashed line indicates the flux-weakening voltage limit.

CVLC zone and MTPV zone, called VCVLC and VMTPV, respectively, have been programmed in the power inverter. Mainly, these modifications are done by reducing the real I_{sc}

Fig. 18. Comparison between the simulation and experimental results of the evolution of torque versus time. Torque reference (magenta, dashed line), simulation results (magenta, solid line), and experimental results (black).

Fig. 19. Comparison between the simulation and experimental results of the evolution of speed versus time. Speed reference (blue, dashed line), simulation results (blue, solid line), and experimental results (black).

and $I_{s,max}$ to the virtual ones. It becomes in a hyperbola that represents the VMTPV zone nearer to the q -axis, and in a current limit circle that is shrunk to its center compared with the original circle. Fig. 20 shows the work point in the dq plane both for the conventional method (cyan line) and the proposed algorithm (black line). The proposed algorithm demonstrates its smooth performance to work from the MTPA zone to the FW zone, and also when changing from CTC, VCVLC, and VMTPV regions.

Therefore, the experimental results confirmed that the proposed method can work at all possible trajectories, if the virtual ones are considered. Moreover, it changes the chosen strategy smoothly, without switching between each of them, by adapting the control algorithm to the necessity of an electric vehicle, which has both the set-point from the gas pedal and CSL. The presented results in this article assure the correct operation of the voltage loop in conjunction with the speed regulator.

Fig. 20. Representation in dq -axis of virtual current limit (dashed-dot red), MTPA trajectory (green), VMTPV trajectory (dashed-dot yellow), torque reference curve (magenta), and conventional (cyan) and proposed (black) dq -trajectory.

VI. CONCLUSION

This paper has proposed an advanced flux-weakening control algorithm focused on the requirements of mild-hybrid and full-electric vehicles. The main feature of this method is to unify the torque and speed controllers without interfering between them until the maximum speed is reached. Besides, it exploits the polar coordinate representation and the per-unit system for generating the current vector reference. Furthermore, it smoothly changes the reference signals coming from the vehicle, beginning with the torque reference from the accelerator pedal up to the speed limit from the cruise speed limiter, according to the most confining signal.

Moreover, the algorithm can be used both for SPMSMs and IPMSMs, and it works in all possible trajectories as Maximum Torque per Ampere, Constant Torque, Current and Voltage Limit, Maximum Torque per Voltage, and Constant Speed with a smooth transition between the Low Back Electromotive Force and the Flux Weakening zones. This transition is managed by a voltage control loop which considers either the resistor voltage drop or the saturation effects.

Both simulation and experimental results prove the reliability and effectiveness of this algorithm on a 2.5 kW and 510 rpm IPMSM in a direct-drive e-motorbike of a 48-V system.

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